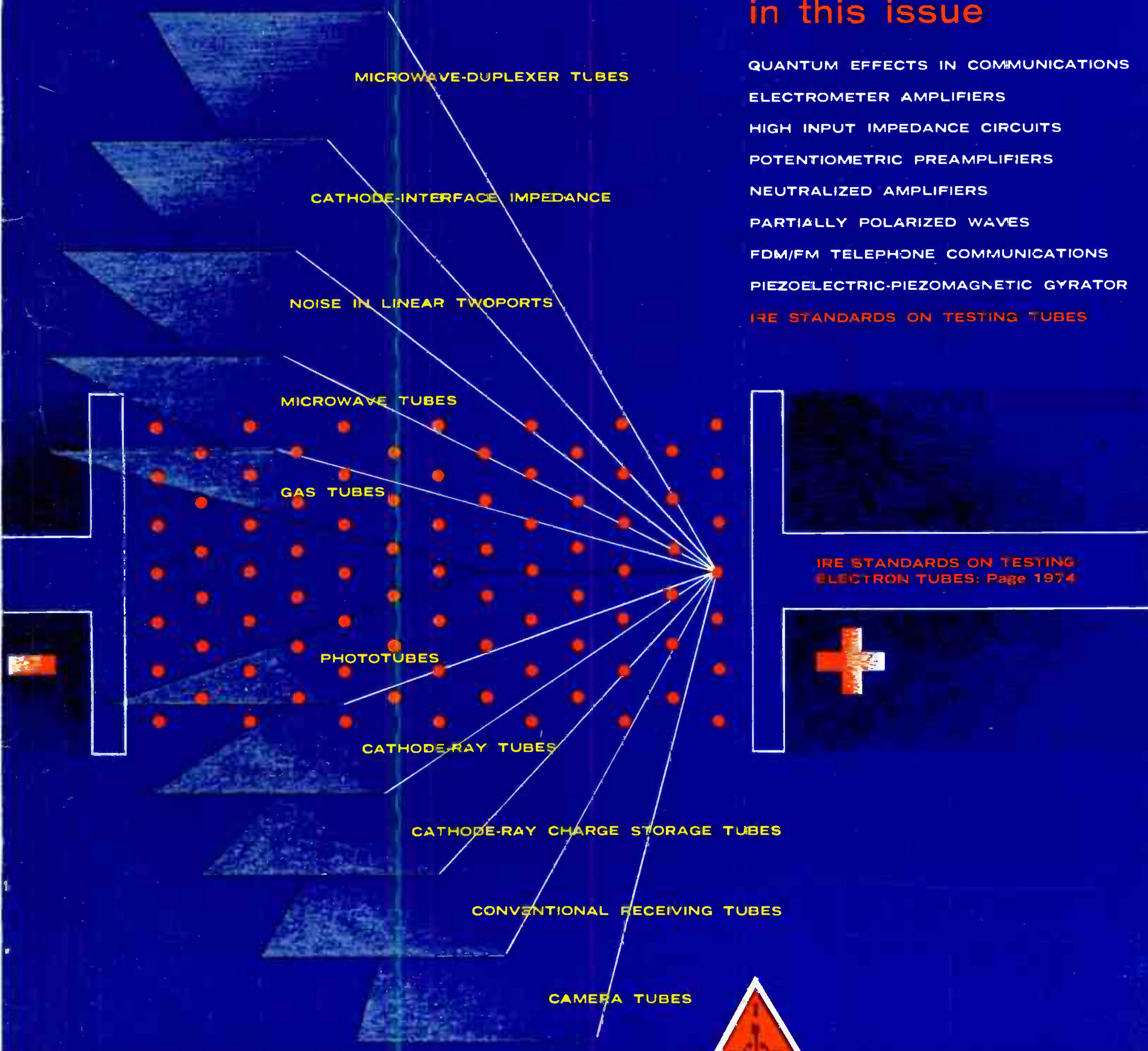
september 1962

the institute of radio engineers

Proceedings of the IRE

in this issue



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Excellent Response	twice as good at low end
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High Efficiency	up to 30% better ... compare DCR
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Rugged	Grade 4, completely metal cased
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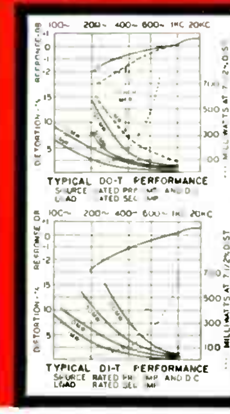
Printed Circuit Use	(solder melting) nylon insulated leads
Suited to Clip Mounting	use Augat #6009-8A clip

TRANSFORMERS PICTURED: DO-T, DI-T



TRANSFORMERS

DO-T No.	Pri. Imp.	D.C. Ma.‡ in Pri.	Sec. Imp.	Pri. Res. DO-T	Pri. Res. DI-T	Mw Level	DI-T No.
DO-T44	80 CT 100 CT	12 10	32 split 40 split	9.8	11.5	500	DI-T44*
DO-T29	120 CT 150 CT	10 10	3.2 4	10		500	
DO-T12	150 CT 200 CT	10 10	12 16	11		500	
DO-T13	300 CT 400 CT	7 7	12 16	20		500	
DO-T19	300 CT	7	600	19	20	500	DI-T19
DO-T30	320 CT 400 CT	7 7	3.2 4	20		500	
DO-T43	400 CT 500 CT	8 6	40 split 50 split	46	50	500	DI-T43*
DO-T42	400 CT 500 CT	8 6	120 split 150 split	46		500	
DO-T41	400 CT 500 CT	8 6	400 split 500 split	46	50	500	DI-T41*
DO-T2	500 600	3 3	50 60	60	65	100	DI-T2
DO-T20	500 CT	5.5	600	31	32	500	DI-T20
DO-T4	600	3	3.2	60		100	
DO-T14	600 CT 800 CT	5 5	12 16	43		500	
DO-T31	640 CT 800 CT	5 5	3.2 4	43		500	
DO-T15	800 CT 1070 CT	4 4	12 16	51		500	
DO-T32	800 CT 1000 CT	4 4	3.2 4	51		500	
DO-T21	900 CT	4	500	53	53	500	DI-T21
DO-T3	1000 1200	3 3	50 60	115	110	100	DI-T3
*DO-T45	1000 CT 1250 CT	3.5 3.5	16,000 split 20,000 split	120		100	
DO-T16	1000 CT 1330 CT	3.5 3.5	12 16	71		500	
DO-T33	1060 CT 1330 CT	3.5 3.5	3.2 4	71		500	
DO-T5	1200	2	3.2	105	110	100	DI-T5
DO-T17	1500 CT 2000 CT	3 3	12 16	108		500	
DO-T22	1500 CT	3	600	86	87	500	DI-T22
DO-T34	1600 CT 2000 CT	3 3	3.2 4	109		500	
DO-T37	2000 CT 2500 CT	3 3	8000 split 10,000 split	195	180	100	DI-T37*
DO-T18	7500 CT 10,000 CT	1 1	12 16	505		100	
DO-T35	8000 CT 10,000 CT	1 1	3.2 4	505		100	

DO-T No.	Pri. Imp.	D.C. Ma.‡ in Pri.	Sec. Imp.	Pri. Res. DO-T	Pri. Res. DI-T	Mw Level	DI-T No.
*DO-T48	8,000 CT 10,000 CT	1 1	1200 CT 1500 CT	640		100	
*DO-T47	9,000 CT 10,000 CT	1 1	9000 CT 10,000 CT	850		100	
DO-T6	10,000	1	3.2	790		100	
DO-T9	10,000 12,000	1 1	500 CT 600 CT	780	870	100	DI-T9
DO-T10	10,000 12,500	1 1	1200 CT 1500 CT	780	870	100	DI-T10
DO-T25	10,000 CT 12,000 CT	1 1	1500 CT 1800 CT	780	870	100	DI-T25
DO-T38	10,000 CT 12,000 CT	1 1	2000 split 2400 split	560	620	100	DI-T38*
DO-T11	10,000 12,500	1 1	2000 CT 2500 CT	780	870	100	DI-T11
DO-T36	10,000 CT 12,000 CT	1 1	10,000 CT 12,000 CT	975	970	100	DI-T36
DO-T1	20,000 30,000	.5 .5	800 1200	830	815	50	DI-T1
DO-T23	20,000 CT 30,000 CT	.5 .5	800 CT 1200 CT	830	815	100	DI-T23
DO-T39	20,000 CT 30,000 CT	.5 .5	1000 split 1500 split	800		100	
DO-T40	40,000 CT 50,000 CT	.25 .25	400 split 500 split	1700		50	
*DO-T46	100,000 CT	0	500 CT	7900		25	
DO-T7	200,000	0	1000	8500		25	
DO-T24	200,000 CT	0	1000 CT	8500		25	
DO-T500	Power DO-T, Pri 28V 380-1000 cycles, sec 6.3V @ 60 ma						

INDUCTORS

DO-T No.		DO-T	DI-T	DI-T No.
*DO-T50 (2 wdgs.)	§.075 Hy 10 ma, .06 Hy 30 ma §§.018 Hy 20 ma, .015 Hy 60 ma	10.5 2.6		
DO-T28	.3 Hy 4 ma, .15 Hy 20 ma	25		
	.1 Hy 4 ma, .08 Hy 10 ma		25	DI-T28
DO-T27	1.25 Hys 2 ma, .5 Hy 11 ma	100		
	.9 Hy 2 ma, .5 Hy 6 ma		105	DI-T27
DO-T8	3.5 Hys 2 ma, 1 Hy 5 ma	560		
	2.5 Hys 2 ma, .9 Hy 4 ma		630	DI-T8
DO-T26	6 Hys 2 ma, 1.5 Hys 5 ma	2100		
	4.5 Hys 2 ma, 1.2 Hys 4 ma		2300	DI-T26
*DO-T49 (2 wdgs.)	§20 Hys 1 ma, 8 Hys 3 ma §§5 Hys 2 ma, 2 Hys 6 ma	5100 1275		
DO-TSH	Drawn Hipermalloy shield and cover 20 30 db			DI-TSH

‡DCMA shown is for single ended useage (under 5% distortion 100MW 1KC) . . . for push pull, DCMA can be any balanced value taken by .5W transistors (under 5% distortion—500MW 1KC) DO-T & DI-T units designed for transistor use only. Pats. Pend.
§Series connected; §§Parallel connected → *Units newly added to series

VOLUME 50, NUMBER 9

September, 1962

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE

contents

COVER

Publication of a new IRE Standard on methods of testing electron tubes is symbolized by a diode, which is basic to all types of tubes, and ten leaves representing the ten parts of this comprehensive document. For information about the Standards and how to obtain it, see p. 1974.

SEPTEMBER

published monthly by The Institute of Radio Engineers, Inc.

Proceedings of the IRE

continued

SEPTEMBER

PROCEEDINGS OF THE IRE, published monthly by The Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N. Y. Cable Address: IRADENGS Manuscripts should be submitted in triplicate to the Editorial Department. Correspondence column items should not exceed four double-spaced pages (illustrations count as one-half page each). 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.

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Get nearly twice the capacitance of older designs in Sprague's new high-gain etched-foil TANTALEX® Capacitors

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HIGH CAPACITANCE Tubular Tantalex Capacitors with almost double the capacitance of standard etched-foil tantalum capacitors have been developed by the Sprague Electric Company to meet the needs of design engineers.

A new etching technique, the result of an intensive research program, gives considerably higher effective surface area to the capacitor electrodes *without sacrifice in reliability or in any of the electrical parameters* by which foil tantalum capacitors are usually judged.

Unlike other "high capacitance" foil tantalums, Sprague Tantalex Capacitors continue to maintain their rigid standards for shelf and service life under severe environmental conditions. Certain performance characteristics have actually been tightened. For example, allowable leakage current has now been halved, making the use of these capacitors possible in many new applications.

Etched-foil Tantalex Capacitors are available in two operating temperature ranges—polarized Type 112D and non-polarized Type 113D for −55 C to +85 C operation, as well as polarized Type 122D and non-polarized Type 123D for −55 C to +125 C operation.

The Foil-type Tantalex Capacitor Line also includes conventional low-gain etched-foil and plain-foil capacitors in both polarized and non-polarized construction, providing a foil tantalum capacitor for every application.

For complete technical data on 85 C capacitors, request Engineering Bulletin 3601B. For the full story on capacitors for 125 C operation, write for Engineering Bulletin 3602B. Address Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

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As was mentioned last month, multiple simultaneous beams from a lossless passive antenna must be orthogonal in space if they are to be free from crosstalk or coupling effects. This prevents one from specifying independently the beam shape and crossover levels in such a system. This month Warren White discusses methods whereby acceptable solutions can be found for a majority of cases.

Orthogonality in Multiple Beam Antennas (PART II)

As we mentioned last month, if multiple beams are to be radiated simultaneously from a lossless passive antenna, then for the beams to be decoupled, they must be orthogonal in space. The exact relation to be satisfied is

$$\int_{-\frac{\pi}{2}}^{\frac{\pi}{2}} \cos\phi\, d\phi \int_{0}^{2\pi} d\theta\, F_j(\theta, \phi) \cdot F_k^*(\theta, \phi) = 0 \text{ if } j \neq k$$

EQUATION I

Since the proof is given elsewhere,[1] it will not be repeated here. F_j (θ, ϕ) is the far field radiation pattern of the j^{th} beam. In the completely general case, F_j must be regarded as a vector quantity with complex elements. One element for example could give the phase and amplitude of the vertically polarized component and the other gives similar data for the horizontally polarized component. Of course, if the polarization is everywhere vertical, for example, we can disregard the second vector component and regard the F_j (θ, ϕ) as complex scalor quantities.

It is easy to see that when the phase centers coincide so that there is no phase difference between F_j and F_k, then if we have a substantial overlap between adjacent beams and negligible or non-existent side lobes, the beams will not be orthogonal. The relation seems to say that to achieve orthogonality between overlapping beams, it is essential to have substantial side lobe regions in which the phase is reversed. Another method, however, is to displace the phase centers. This will introduce a phase factor under the integral sign in equation (1) and will affect the value of the integral. As a practical matter, the beams will be orthogonal if the phase centers are displaced to the point where the apertures no longer overlap. One method of obtaining orthogonal operation then is to use independent apertures for adjacent beams. In Figure 1, we show a system using two apertures with the odd-numbered beams being radiated from the left and the even-numbered beams from the right. In cases where the side lobe requirements are more severe, three or more independent apertures may be required. If we are forming beams in a two-dimensional bundle instead of simply a one-dimensional stack, the number of apertures required becomes greater yet.

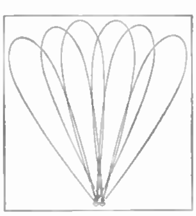

FIGURE 1

Since the use of multiple apertures is not likely to be cheap, it becomes desirable to investigate other methods of decoupling. As it turns out, the requirement for orthogonality is based on the fact that in a lossless passive system, the output power must be strictly equal to the input power. If we permit the use of lossy elements or active elements, decoupled operation can be readily achieved with non-orthogonal beams.

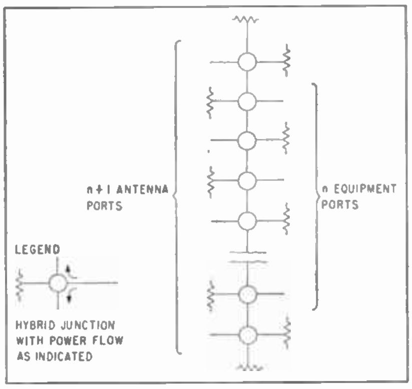

FIGURE 2

Figure 2 shows one type of lossy network capable of providing non-orthogonal beams. We assume the n + 1 ports on the left are connected to antenna ports that provide sin x/x or similar beams that are orthogonal but that do not have satisfactory crossover or side lobe levels. By exciting two of these ports simultaneously, we achieve a composite beam in which the side lobes are reduced. Furthermore, the beam formed by exciting ports 1 and 2 has a good crossover with the beam formed by exciting ports 2 and 3. With the hybrid network shown, energy fed to one of the ports on the right will excite two adjacent ports on the left and adjacent ports on the right are decoupled despite the fact that they feed the same antenna port. The price paid for this type of operation is that half of the input power winds up being dissipated in the terminations and only half is radiated. More elaborate networks can be devised that provide still better side lobe and crossover performance but result in a still greater loss in the terminations.

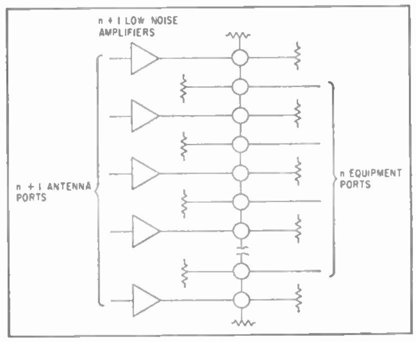

FIGURE 3

In receiving applications, the presence of lossy terminations is doubly damaging, since not only is the signal attenuated but additional noise is introduced. For this reason, a third alternative, that of active circuit decoupling is likely to be attractive. One method of achieving this type decoupling is shown in Figure 3. This network is similar to that of Figure 2 except for the amplifiers inserted between the antenna proper and the network. These amplifiers established the signal-to-noise ratios before the losses are introduced. Friis and Feldman[2] describe an alternative technique applicable to arrays. Here a receiver front end is provided for each element of the array and the beam-forming network operates at IF. Again the signal-to-noise ratio is established before any losses are introduced.

Doubtless other methods of operating with non-orthogonal beams can be found. Orthogonality might be achieved by putting adjacent beams on different frequencies or at different times. It is hoped, however, that an understanding of the problem can avoid effort wasted in attempting the impossible.

[1] W. D. White, "Pattern Limitations in Multiple Base Antennas," IRE Trans. on Antennas and Propagation, Vol. AP-10, No. 4, July 1962.

[2] H. T. Friis and C. B. Feldman, "A Multiple Unit Steerable Antenna for Shortwave Reception," BSTJ, Vol. 16, No. 3, p. 337-419, July 1937.

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

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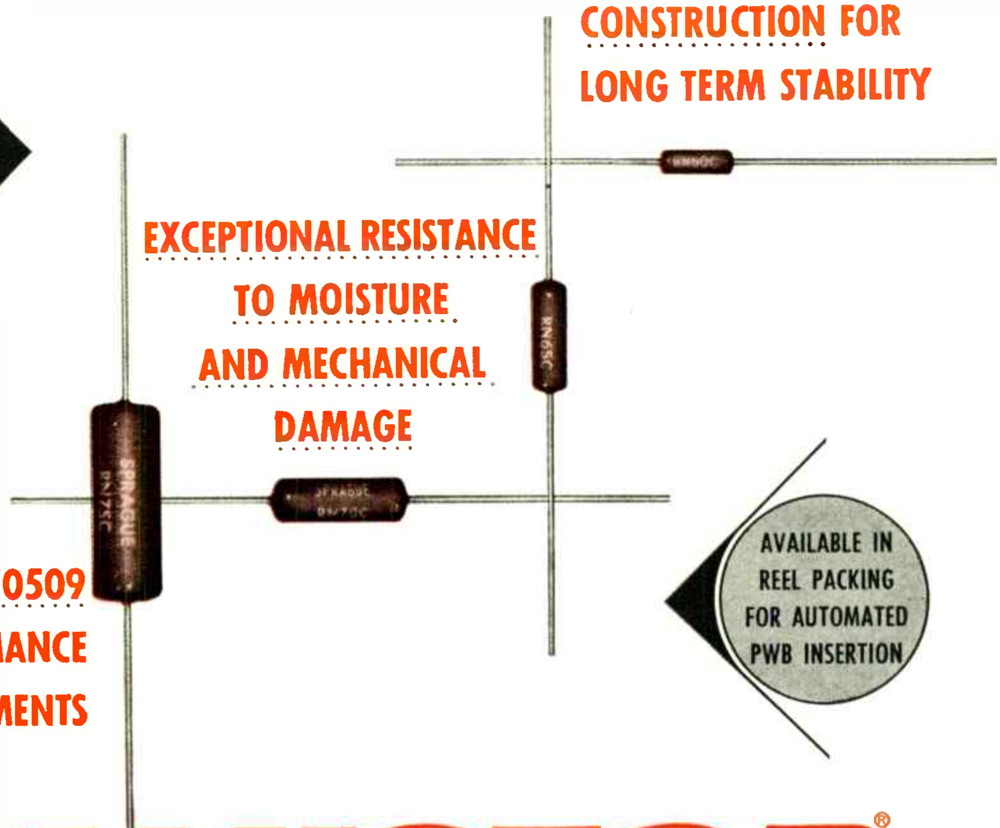

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Providing close accuracy, reliability and stability with low controlled temperature coefficients, these molded case metal-film resistors outperform precision wirewound and carbon film resistors. Prime characteristics include minimum inherent noise level, negligible voltage coefficient of resistance and excellent long-time stability under rated load as well as under severe conditions of humidity.

Close tracking of resistance values of 2 or more resistors over a wide temperature range is another key performance characteristic of molded-case Filmistor Metal Film Resistors. This is especially important where they are used to make highly accurate ratio dividers.

Filmistor Metal Film Resistors, in ⅛, ¼, ½ and 1 watt ratings, surpass stringent performance requirements of MIL-R-10509D, Characteristics C and E. Write for Engineering Bulletin No. 7025 to: Technical Literature Section, Sprague Electric Co., 235 Marshall Street, North Adams, Mass.

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NEWS FROM
BELL LABORATORIES

A simple, highly sensitive microwave amplifier

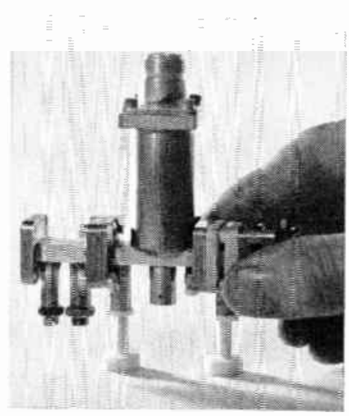
Bell Laboratories' Michael Chruney adjusts waveguide assembly (in circle) housing the diode. After adjustment the entire parametric amplifier will be immersed in liquid nitrogen in dewar at left. The new amplifier operates at 4170 megacycles (center of band) and provides an almost flat gain of 38 db over a 50-megacycle band with a noise figure of approximately 0.6 db.

Bell Laboratories engineers have developed an extremely sensitive parametric amplifier which approaches the maser in sensitivity. Both will be used in experiments with Telstar, the Bell System's experimental communications satellite.

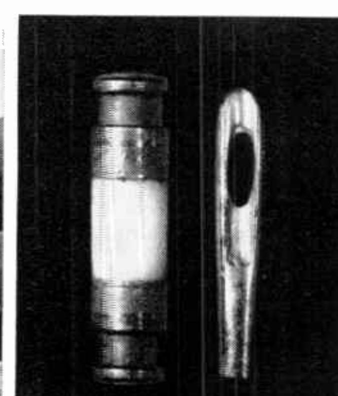
Heart of the parametric amplifier is a newly developed semiconductor diode with very low intrinsic noise. Previously, the sensitivity of such amplifiers at microwave frequencies was severely limited by the unwanted noise generated in their diodes. The new diode, no bigger than the eye-end of a needle, solved this problem.

Our engineers also devised new circuitry to stabilize precisely the output of the klystron (microwave generator) supplying power for the amplifier. To reduce further the intrinsic noise of the amplifier, they immersed the diode and its circuits in liquid nitrogen, utilizing a new cooling arrangement which economically maintains a low temperature for many days without attention.

The new amplifier fills a need in the communications field for a simple microwave amplifier of high sensitivity in applications for which the higher sensitivity of the maser does not justify its additional complication.



Close-up of the waveguide assembly, in which Bell Telephone Laboratories' newly developed diode is located.



Heart of amplifier—a hermetically sealed gallium arsenide diode—is compared with eye of average-sized sewing needle.

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BEFORE ORBIT...

Telstar's first 1300 miles were on the ground

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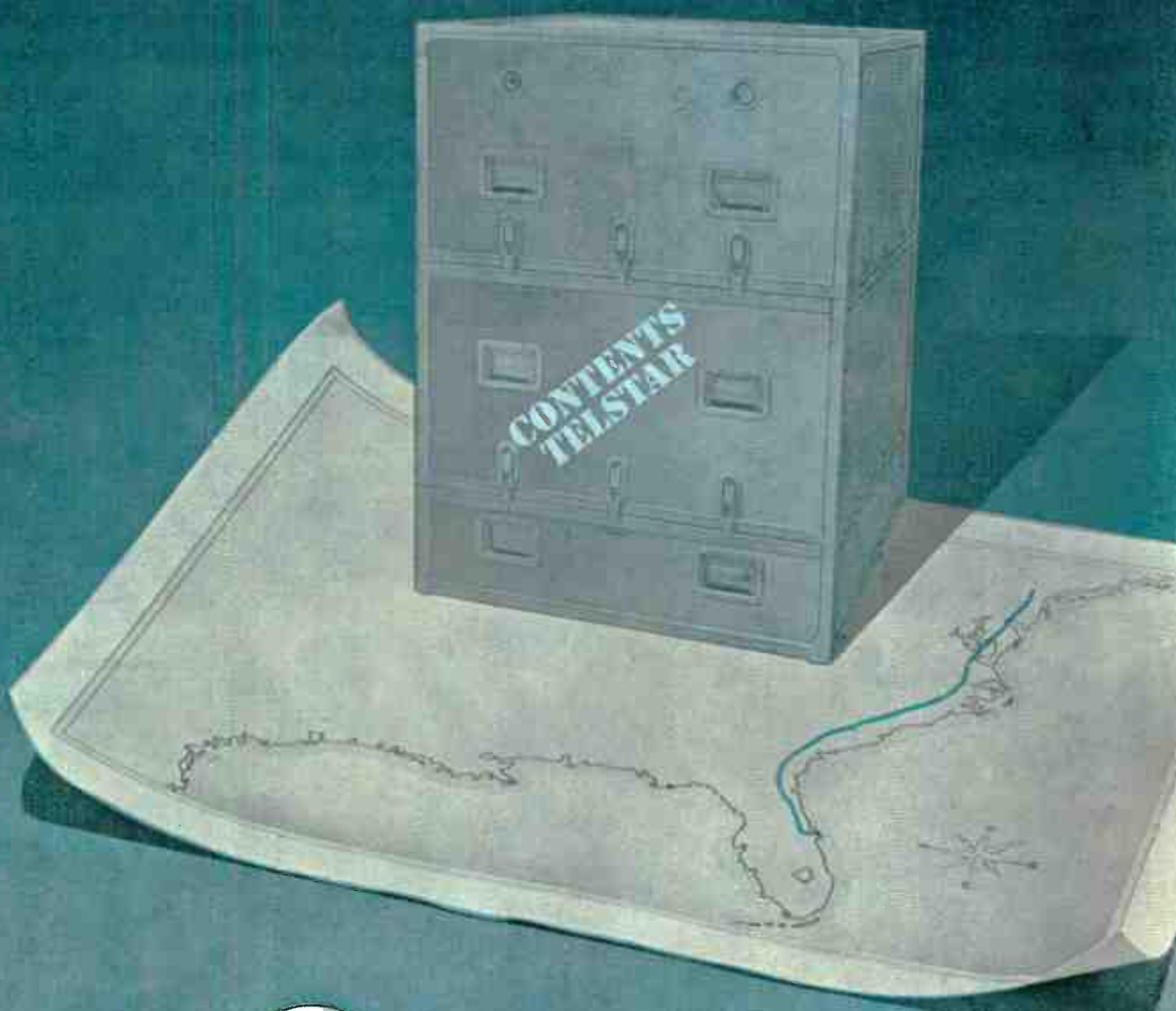

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Meetings with Exhibits

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

September 21-22, 1962

Conference on Communications, Roosevelt Hotel, Cedar Rapids, Iowa

Exhibits: Mr. Richard L. Jaycox, P.O. Box 948, Cedar Rapids, Iowa

October 1-3, 1962

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

Exhibits: Mr. Charles Glaviano, 45 Meadow Dr., Rome, N.Y.

October 2-4, 1962

Seventh National Symposium on Space Electronics & Telemetry. Fontainebleau Hotel, Miami Beach, Fla.

Exhibits: Mr. Charles H. Doersam, Jr., Instruments for Industry, 101 New South Road, Hicksville, L.I., N.Y.

October 8-10, 1962

National Electronics Conference, McCormick Place, Chicago, Ill.

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

October 15-18, 1962

Symposium on Space Phenomena & Measurement, Statler-Hilton Hotel, Detroit, Mich.

Exhibits: Mr. J. B. Bullock, University of Michigan, Ann Arbor, Mich.

November 1-2, 1962

Sixth National Conference on Product Engineering and Production. Jack Tar Hotel, San Francisco, Calif.

Exhibits: Mr. W. Dale Fuller, Lockheed Missiles & Space Div., P.O. Box 504, Sunnyvale, Calif.

November 4-7, 1962

Fifteenth Annual Conference on Electronic Techniques in Medicine & Biology, Conrad Hilton Hotel, Chicago, Ill.

Exhibits: Professional Associates, Inc., 6520 Clayton Rd., Saint Louis 17, Mo.

November 5-7, 1962

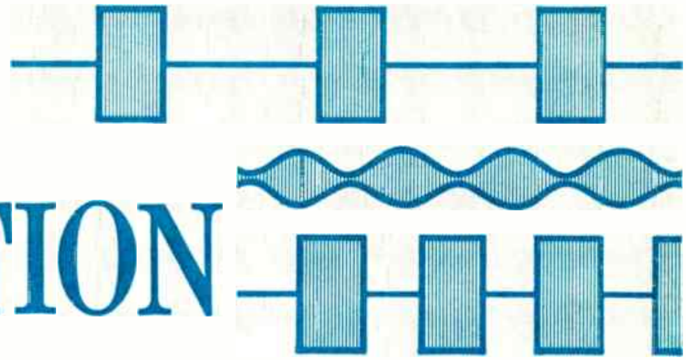
NEREM (Northeast Research & Engineering Meeting), Commonwealth Armory & Somerset Hotel, Boston, Mass.

Exhibits: Mr. S. K. Gibson, Instruments of New England, 108 Greenwood Lane, Waltham 5, Mass.

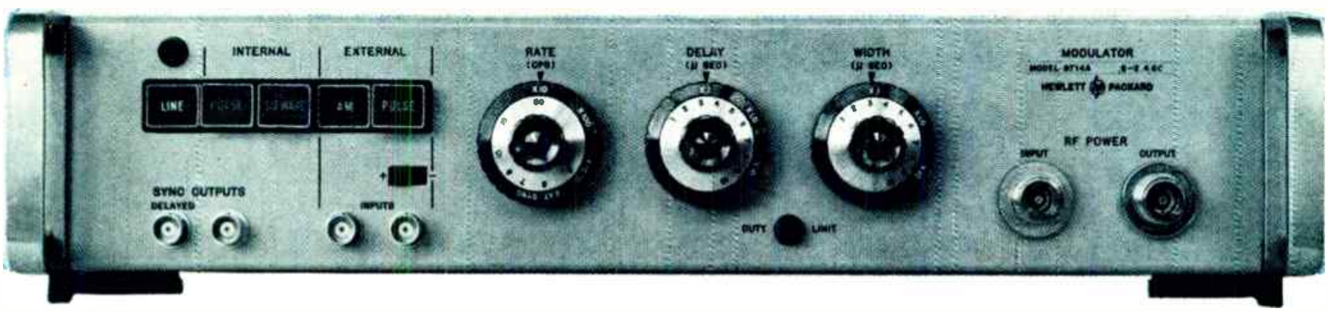
(Continued on page 10A)

from any signal source 800 to 2,400 MC
from any cw signal to 1 watt

PERFECT MODULATION



free of jitter and incidental FM



New ⓗⓟ 8714A Modulator delivers fast, jitter-free and FM-free pulses from any signal source or generator from 800 to 2,400 MC. A rise time of 20 nsec and fall time of 10 nsec are the result of using an absorption modulator composed of electrically controlled PIN diodes. Since the signal source, either klystron or BWO, is always operated in the cw mode, and the PIN diodes present a good match under all modulation conditions, pulses from the 8714A are free from incidental FM and jitter.

The 8714A has internal modulation generators providing square wave and pulse modulation 50 cps to 50 KC, with continuously variable width and delay controls. The internal square wave and pulse generators may also be externally synchronized at rates up to 1 MC for non-periodic modulation. External modulation capability in pulse and AM modes to at least 1 MC is also provided.

Push button function selection, simple controls and versatile rack or bench mount make the 8714A easy to use. Solid state design and rugged construction assure years of convenient and trouble-free service.

SPECIFICATIONS

Frequency range:	800 to 2,400 MC
RF input power:	Maximum 1 watt
SWR:	1.5 maximum at minimum attenuation, 2.0 maximum at 80 db attenuation
Impedance:	50 ohms nominal
On-off ratio:	Better than 80 db
Rise time:	20 nsec
Decay time:	10 nsec
Internal repetition rate:	Variable from 50 cps to 50 KC, 3 decade ranges
Jitter:	< 1.0 nsec
Internal pulse width:	Variable from 0.1 μsec to 100 μsec in 3 decade ranges
External sync:	0 to 1 MC depending on width/delay ratio
External AM mod:	Max. freq., 1 MC sinusoidal
External pulse mod:	Max. rep. rate, 1 MC
Dimensions:	16¾" wide x 4" high x 18⅜" deep
Weight:	23 lbs.
Price:	$850.00

7879 *Data subject to change without notice. Price f.o.b. factory*

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Meetings with Exhibits

(Continued from page 8A)

November 12-15, 1962

Eighth Annual Conference on Magnetism & Magnetic Materials, Penn-Sheraton Hotel, Pittsburgh, Pa.

Exhibits: Mr. J. L. Whitlock, John Leslie Whitlock Associates, 253 Waples Mill Rd., Oakton, Va.

November 16-17, 1962

Communications Symposium, Queen Elizabeth Hotel, Montreal, P.Q., Canada

Exhibits: Mr. Arthur H. Gregory, Northern Electric Co., Ltd., 1600 Dorchester Blvd. W., Montreal, P.Q., Canada

November 19-20, 1962

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

Exhibits: Dr. Arthur Goldsmith, Wilcox Electric Co., 1400 Chestnut, Kansas City 27, Mo.

December 4-6, 1962

Fall Joint Computer Conference, Sheraton Hotel, Philadelphia, Pa.

Exhibits: Mr. R. A. C. Lane, RCA Building 204-1, Camden 8, N.J.

December 6-7, 1962

13th National Conference on Vehicular Communications, Mayfair Hotel, Los Angeles, Calif.

Exhibits: Mr. Leslie M. Walker, Los Angeles County Dept. of Communications, 500 West Temple St., Los Angeles 12, Calif.

January 30-February 1, 1963

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

Exhibits: IRE Los Angeles Office, 1435 La Cienega Blvd., Los Angeles, Calif.

February 11-15, 1963

Third International Symposium on Quantum Electronics, UNESCO Bldg., Paris, France

Exhibits: Madame Cauchy, 7 rue de Madrid, Paris 8, France

March 25-28, 1963

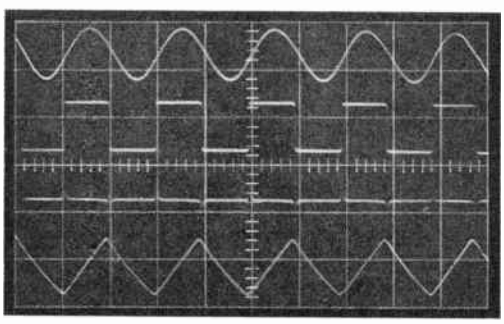
International Radio & Electronics Show and IRE International Convention, New York Coliseum and Waldorf-Astoria Hotel, New York. N.Y.

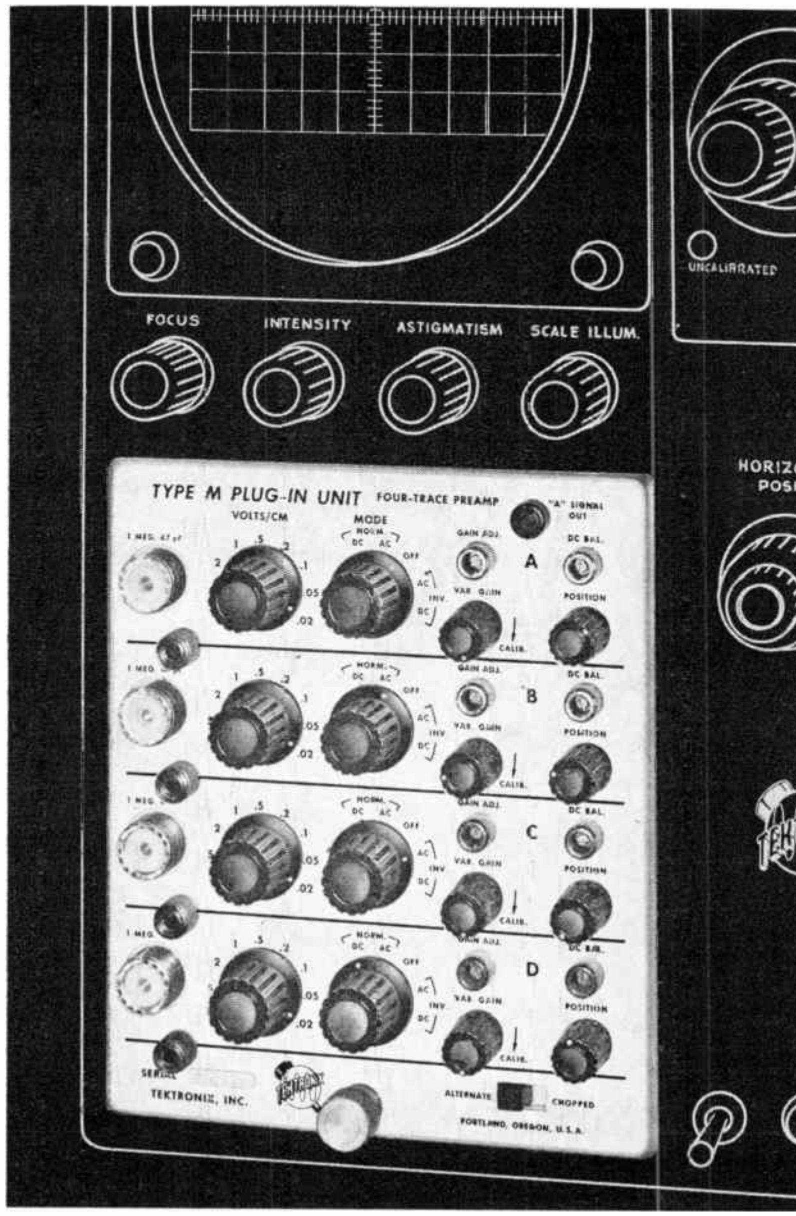
Exhibits: Mr. William C. Copp, IRE Advertising Dept., 72 West 45th St., New York 36, N.Y.

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

NEW
4-TRACE
PREAMPLIFIER

for Tektronix Oscilloscopes that accept letter-series plug-in units





TYPE M UNIT Seventeenth in the letter-series of plug-in units, the new Type M Unit adds multiple-trace displays to the wide range of applications possible with your Tektronix oscilloscope.

With a Type M Unit, you can observe up to four signals—either separately, or in any combination.

Independent controls for each amplifier channel permit you to position, attenuate, invert input signals as desired.

Other convenient preamplifier features—such as triggered or free-running electronic switching . . . ac-coupling or dc-coupling . . . and, after initial hookup, little or no cable switching—ideally suit the new Type M Unit for multiple-trace presentations in the laboratory or in the field.

For a demonstration of this new 4-Trace Preamplifier, please call your Tektronix Field Engineer.

CHARACTERISTICS

Operating Modes—Any combination of one to four channels electronically switched—at the end of each sweep or at a free-running rate of about 1 mc (1 µsec width samples). Or each channel separately. *Channel Sensitivity*—20 mv/cm to 10 v/cm in 9 calibrated steps. Continuously variable uncalibrated between steps, and to 25 v/cm. *Channel A Signal*—available on front panel for optimum triggering in some applications. *Frequency Response and Risetime*—With Type 540-Series and Type 550-Series Oscilloscopes dc to 20 mc, 17 nsec. With Type 531A, 533A, 535A Oscilloscopes dc to 14 mc, 25 nsec. *Constant Input Impedance*—at all attenuator settings.

Type M Plug-in Unit $525 00

U. S. Sales Price f.o.b. Beaverton, Oregon

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Model 451

750 to 2750 Mc—
only $1325!

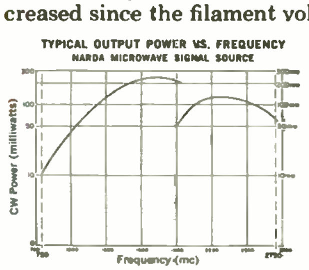
Now! Up to 300 mw of cw or modulated microwave signal from one package—one knob!

Narda Microwave SIGNAL SOURCE

You'd pay more for a universal klystron power supply alone! Yet, for just $1325 you get a complete laboratory and test department signal source, covering 750 to 2750 mc, and all in one compact package!

Power? Plenty! Up to 300 mw! (See typical curve below.) And, power's adjustable over a 50 db range. Tuning? One knob does it with a direct-reading frequency indicator tape. Modulation? Included! Square wave and pulse modulation in the most widely-used combinations are built in; also provision for external modulation—AM and FM.

Reliability? Life? Provided for! Klystron life and stability are increased since the filament voltage remains constant despite line voltage fluctuations. The cavity features non-contacting tuning for longer life and trouble-free operation.



The Model 451 comes complete with klystron, wide band cavity, 50 db range attenuator, solid state power supply and modulator. And it's available *now!* For complete specs (and catalog of other Narda precision products) write today. Address: Dept. PIRE-2-8.

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Plainview, L. I., New York
Area Code 516 GE 3-9000

IRE International News

Current IRE Statistics

(As of April 30, 1962)

Membership—100,075
Sections*—111
Subsections*—33
Professional Groups*—29
Professional Group Chapters—299
Student Branches†—229

* See this issue for a list.
† See June, 1962 for a list.

Calendar of Coming Events and Author's Deadlines*

1962

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

Sept. 3–7: Nat'l Advanced Technology, Management Conf., World Fair Grounds, Opera House, Seattle, Wash.

Sept. 3–7: 45th Internat'l. Congress on Microwave Tubes, Kurhaus Hotel, Scheveningen, Netherlands.

Sept. 4–7: ACM Nat'l. Conf., Hotel Syracuse, Syracuse, N. Y.

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

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

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

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

Oct. 1–3: 8th Nat'l. Communications Symp., Hotel Utica, Utica, N. Y.

Oct. 2–4: Nat'l. Symp. on Space Elec. and Telemetry, Fountainebleau Hotel, Miami Beach, Fla.

Oct. 7–12: AIEE 1962 Fall General Meeting—3rd Ann. Symp. on Switching, Circuit Theory and Logical Design, Chicago, Ill.

Oct. 8–10: Nat'l. Elec. Conf., McCormick Pl., Chicago, Ill.

Oct. 12–13: 7th Annual North Carolina Section Symp., Greensboro Coliseum, Greensboro, N. C.

Oct. 15–17: URSI-IRE Fall Meeting, Ottawa, Canada.

Oct. 15–18: Symp. on Space Phenomena and Measurement, Statler-Hilton, Detroit, Mich.

Oct. 22–24: ECCANE (East Coast Conf. on Aerospace and Navigational Elec.) Emerson Hotel, Baltimore, Md.

Oct. 25–27: Electron Devices Meeting, Sheraton Park Hotel, Wash., D. C.

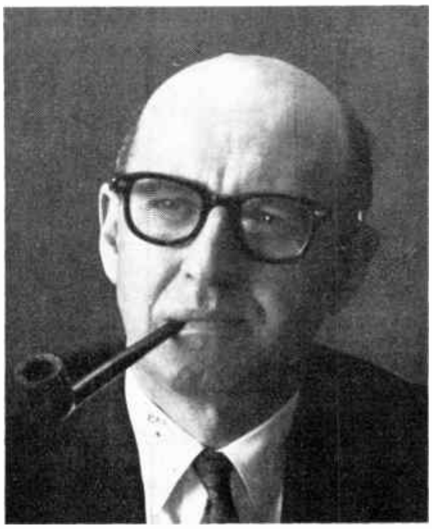
* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Donald G. Fink Appointed General Manager of IEEE

The Presidents of IRE and AIEE have announced the appointment of Donald G. Fink as General Manager of the Institute of Electrical and Electronic Engineers (IEEE), scheduled to be formed in January, 1963, when the IRE and AIEE are merged. He is now Director of the Philco Scientific Laboratory and will remain in that post until his successor is appointed. Warren H. Chase, President of AIEE, and Patrick E. Haggerty, President of IRE, stated that Mr. Fink was the unanimous choice of a 14-man Merger Committee appointed by the Boards of the two societies.

As General Manager, he will be the chief staff officer responsible for the day-to-day operation of the world's largest engineering society, with an estimated membership of 160,000. Among his responsibilities will be



Donald G. Fink (*above*) recently appointed General Manager of the Institute of Electrical and Electronic Engineers (IEEE).

the supervision of the publication of technical periodicals which, for AIEE and IRE, now total forty.

Dr. George W. Bailey, Executive Secretary of the IRE, will continue with the IEEE for at least two years as Consultant.

Mr. Fink has combined notable careers in technical publishing, government service, and industrial research. He received the B.S. degree from M.I.T. and the M.S. degree from Columbia University, both in electrical engineering. After a year on the research staff of M.I.T., he joined the editorial staff of *Electronics*, becoming Editor-in-Chief in 1946. While on leave of absence from his editorial duties during World War II, he became Head of the Loran Division of the M.I.T. Radiation Laboratory and, in 1943, was appointed Expert Consultant, in the field of electronic navigation and radar on the staff of the Secretary of War. For his overseas wartime service he was awarded the Medal of Freedom. In 1946 he served as Civilian Consultant at the Bikini atom bomb tests.

In 1952 he joined the Philco Research Laboratories as Director of Research, and in 1959 he was put in charge of all of the company's research activities. In 1961 he was appointed Vice President for Research. Since the Ford Motor Company acquired Philco in 1961, he has continued to head research activities, as Director of the Philco Scientific Laboratory.

Since the war, Mr. Fink has been advisor and consultant to the Federal Communications Commission, National Bureau of Standards, U. S. Senate, Department of State, and Department of the Army. Appointed to the Army Scientific Advisory Panel in 1957, he is presently Vice Chairman of the Communications and Electronics Subpanel. From 1950 to 1952 he was Vice Chairman of the National Television System Committee.

Mr. Fink is a Fellow of both the AIEE and the IRE, and is a member of the AIEE Electronics Committee. He was Editor of the PROCEEDINGS OF THE IRE in 1956 and 1957. In 1958, while he was President of the IRE, he participated in discussions with the President of the AIEE, setting up closer ties between the two societies which later led to the merger plans.

Final Call for Papers Automatic Control Conference

The Fourth Joint Automatic Control Conference will be held at the University of Minnesota, Minneapolis, Minn. on June 19–21, 1963. Prospective authors are invited to submit abstracts of 100 words by September 30, and manuscripts by November 15, 1962.

Papers are invited on control theory, applications, and components. Particular efforts are being made to include a broad range of application papers, and one or more applications symposia are being developed. Components papers, also, are especially invited.

The sponsoring societies of the JACC are the American Institute of Chemical Engineers (which has prime responsibility in 1963), the American Institute of Electrical Engineers, the American Society of Mechanical Engineers, the IRE, and the Instrument Society of America. Abstracts and papers may be submitted through the member society headquarters with the designation "for 1963 JACC" or to the Program Chairman, Professor Otis L. Updike, Department of Chemical Engineering, University of Virginia, Charlottesville, Va. Further details on paper submission will be supplied after abstracts are received. The early deadline schedule has been established to permit full preprinting of conference papers.

Papers prepared for the Congress of the International Federation for Automatic Control in Basle may be presented also at the JACC, and will be preprinted in abstract only to conform with IFAC requirements.

Mrs. David K. Barton holds the M. Barry Carlton Award for 1962 presented to her husband (*right*), RCA Missile and Surface Radar Department, Moorestown, N. J., for his paper, "The Future of Pulse Radar for Missile and Space Range Instrumentation," published in the October, 1961 issue of the IRE TRANSACTIONS ON MILITARY ELECTRONICS. The presentation was made at MIL-E CON, 1962 by William L. Doxey, National Chairman, 1961–1962.

CALL FOR PAPERS MIL-E-CON 1963

The 1963 National Winter Convention on Military Electronics will be held at the Ambassador Hotel, Los Angeles, Calif., on January 30–February 1, 1963. The Convention will be keynoted, as in past years, by an opening panel session conducted by senior members of the Department of Defense.

In order to assure the greatest possible scope and depth of technical papers, several Secret sessions will be held, in addition to Confidential and Unclassified sessions. Papers in the following fields will be presented:

Systems: Ballistic Missile Systems (S); Space Systems (C); Tactical Warfare Systems (C); Antisubmarine Warfare Systems (S); Ballistic Missile Defense Systems (S); Airborne Systems (including fire control and reconnaissance) (C); Command and Control Systems (C).

Technologies: Radar (C); Infrared (C); Lasers (U); Communication and Telemetry (U); Information and Data Processing (U); Guidance and Navigation (C); Instrumentation (U); Displays and Man-Machine Design (U); Aerospace Ground Support Equipment (U); Microelectronics (U); Fuel Cells and Space Power Supplies (U).

General: Program Management Techniques (U); Space Environmental Testing (U); Feedback from the Military User (U).

The Classified sessions will be sponsored by the U. S. Air Force Systems Command.

Unclassified and classified papers are invited for presentation. Authors should submit a 100-word unclassified abstract, a 500-word summary, and a short biography. Authors of classified papers are responsible for obtaining all necessary clearances. All papers should be sent to Dr. F. P. Adler, Manager, Space Systems Div., Hughes Aircraft Company, Culver City, Calif., no later than October 15, 1962.

Plans are being made for the publication of the Convention *Proceedings* which will include all unclassified papers and unclassified abstracts of classified papers. Authors will be contacted at a later date regarding publication.

CALL FOR PAPERS THIRD PICA CONFERENCE

The Third Power Industry Computer Conference, sponsored by the American Institute of Electrical Engineers, will be held on April 24–26, 1963 in Phoenix, Ariz. New concepts of computer applications, the integrated computer approach, and the future impact of computers on electric power system engineering will be the general themes of the Conference.

The aim of the Conference will be to probe current thinking of the direction that computer activities in the electric power industry will take in the coming decade. Invited papers by recognized leaders in their fields will set the stage for each of several sessions covering aspects of this general field. The remaining sessions will be devoted to reporting on the present state of the application of computers to the major areas of system planning, operation, control, and design.

Prospective authors should submit a title and a 150-200-word abstract as soon as possible to: G. W. Stagg, Program Chairman, American Electric Power Service Corp., 2 Broadway, New York 8, N. Y. Papers presented at the Conference are eligible for AIEE *Transactions* status, and thus the standard AIEE publication specifications apply. The deadline for complete papers will be January 24, 1963.

Calendar of Coming Events and Author's Deadlines*

(*Continued from page 14A*)

Oct. 30–31: Conf. on Spaceborne Computer Engineering, Disneyland Hotel, Anaheim, Calif.

Nov. 1–2: 6th Nat'l. Conf. on Product Engr. and Production, Jack Tar Hotel, San Francisco, Calif.

Nov. 4–7: 15th Annual Conf. on Engineering in Medicine and Biology, Conrad Hilton Hotel, Chicago, Ill.

Nov. 5–7: NEREM (Northeast Res. and Engr. Meeting), Commonwealth Armory, Somerset Hotel, Boston, Mass.

Nov. 12–14: Radio Fall Meeting, King Edward Hotel, Toronto, Ont., Canada.

Nov. 12–15: 8th Annual Conf. on Magnetism and Magnetic Mat., Penn-Sheraton, Pittsburgh, Pa.

Nov. 16–17: 2nd Canadian IRE Communications Symposium, Queen Eliz. Hotel, Montreal, P.Q., Canada.

Nov. 19–20: MAECON (Mid-America Electronics Conf.), Hotel Continental, Kansas City, Mo.

Nov. 28–30: 1962 Ultrasonics Symp., Columbia Univ., New York, N. Y.

Dec. 4–6: FJCC (Fall Joint Computer Conf.), Sheraton Hotel, Philadelphia, Pa.

Dec. 6–7: 13th Nat'l. Conf. on Vehicular Communications, Disneyland Motel, Los Angeles, Calif.

1963

Jan. 8–10: Millimeter and Submillimeter Conf., Cherry Plaza Hotel, Orlando, Fla. (DL*: Sept. 15, 1962, J. J. Gallagher, MP-172-Box 5837, The Martin Co., Orlando, Fla.)

Jan. 21–24: 9th Nat'l. Symp. on Reliability and Quality Control, Sheraton Palace Hotel, San Francisco, Calif.

Jan. 30–Feb. 1: 4th Winter Convention on Military Electronics, Ambassador Hotel, Los Angeles, Calif.

Feb. 11–15: 3rd Internat'l. Symp. on Quantum Electronics, UNESCO Bldg., Paris, France. (DL*: Oct. 1, 1962, Madame Cauchy, Secretaire 3ème Congress d'Electronique Quantique, 7 rue de Madrid, Paris 8ème, France.)

Feb. 20–22: Internat'l. Solid State Circuits Conf., Sheraton Hotel and Univ. of Pa., Phila., Pa. (DL*: Nov. 1, 1962, S. K. Ghandi, Philco Scientific Lab., Blue Bell, Pa.)

Mar. 25–28: IRE International Convention, Coliseum and Waldorf Astoria Hotel, New York, N. Y. (DL*: Oct. 19, 1962, Dr. D. B. Sinclair, IRE, 1 E. 79 St. New York 21, N. Y.)

Apr. 17–19: Southwestern IRE Conf. and Elec. Show, Dallas Memorial Auditorium, Dallas, Tex.

Apr. 17–19: Internat'l. Tech. Conf. on Non-Linear Magnetics, Shoreham Hotel, Washington, D. C. (DL*: Nov. 5, 1962, J. J. Suozzi, BTL Labs., Whippany, N. J.)

* DL = Deadline for submitting abstracts.

Call for Papers

1963 IRE International Convention

March 25–28, 1963

Waldorf-Astoria Hotel and the New York Coliseum, New York, N. Y.

Prospective authors are requested to submit all of the following information by the

Deadline Date of October 19, 1962

1. 100-word abstract *in triplicate*, title of paper, name and address
2. 500-word summary *in triplicate*, title of paper, name and address
3. Indicate the technical field in which your paper should be classified:

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

Note: Only original papers, not published or presented prior to the 1962 IRE International Convention, will be considered. *Any necessary military or company clearance of papers must be granted prior to submission.*

Address all material to: Dr. Donald B. Sinclair, Chairman
1963 Technical Program Committee
The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

IRE President Patrick E. Haggerty (*left*) presenting a $500.00 scholarship fund check from the Canaveral Section to Dr. Jerome P. Keuper (*right*), President of Brevard Engineering College. The fund is to be used for deserving scholars in space education. Scene: IRE Canaveral Section Annual Meeting, Officers' Club, Patrick Air Force Base, Fla., June 15, 1962.

New IRE Section and Region Established

The IRE Executive Committee approved the establishment of a new Section, to be known as the United Kingdom Section with territory consisting of: Channel Islands, Isle of Man, Northern Ireland, and United Kingdom of Great Britain.

A new Region, Region 9, has been established with territorial boundaries based on the European Broadcasting Area, defined in the Radio Regulations of the ITU, as follows: "The European Broadcasting Area is bounded on the West by the Western boundary of Region 1 (of ITU), on the East by the meridian 40° East of Greenwich and on the South by the parallel 30° North so as to include the western part of the U.S.S.R. and the territories bordering the Mediterranean. . . . In addition, Iraq is included . . . ."

The IRE Sections included in Region 9 are Benelux, Egypt, France, Geneva, Israel, Italy, and United Kingdom.

Obituary

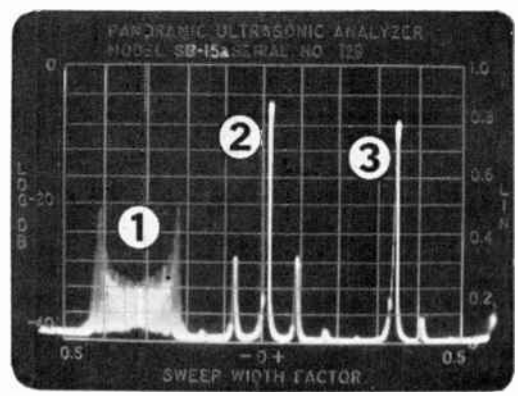
Howard P. Corwith (M'26–SM'43–F'52) (L), Research and Engineering Vice President of Western Union for ten years before his retirement on December 31, 1959, died of a heart attack on June 10, 1962. For 43 years he contributed substantially to the company's advances in research, especially in microwave beam, facsimile and submarine cable communications. Under his leadership, research and development had a prominent role in revolutionizing Western Union's operations during the decade in which they became automatic, electronic and ultra modern.

Born at Water Mill, N. Y., on December 3, 1892, he received his engineering degree at Cornell University, Ithaca, N. Y., in 1916 and joined Western Union as an engineering assistant. He served as an Ensign in the U. S. Navy during World War I. He established Western Union's Electronic Laboratories at Water Mill in 1925 for research and development of radio and electronic devices, and was in charge of the laboratories until 1943. In 1946 he was appointed Director of Research and became Vice President in July, 1949. Before his retirement, he had been a Director of Microwave Associates, Inc.; Technical Operations, Inc.; Dynametrics Corporation; and Teleprinter Corporation, all affiliates of Western Union in electronics and allied fields.

Mr. Corwith was a past President and Director of the New York Electrical Society and a Fellow of the American Institute of Electrical Engineers.

measure / analyze, 100 cps - 600 kc signals quickly, easily, with one compact instrument

PANORAMIC'S SB-15a ULTRASONIC SPECTRUM ANALYZER



Lab setup shows SB-15a versatility. (1) FM display measures dynamic deviation. (2) & (3) are AM and SSB signals, respectively, with sine wave modulation.

Panoramic's advanced Model SB-15a automatically and repetitively scans spectrum segments from 1 kc to 200 kc wide through the entire range (100 cps to 600 kc) . . . plots frequency and amplitude along the calibrated X and Y axes of a long persistence CRT, or on a 12 x 4½" chart (optional RC-3a/15). Sweep rates are adjustable from 1 to 60 cps.

Adjustable resolution enables selection and detailed examination of signals as close as 100 cps. Self-checking internal frequency markers every 10 kc. Also internal amplitude reference • Only 8¾" high, the SB-15a is completely self-contained, needs no external power supply or regulator.

PANORAMIC PRESENTATION MEANS

• quick signal location, minimum chance of missing weak signals or holes in spectrum • faster measurements—no tedious point-by-point plots • reliable spotting of low level discrete signals in noise • positive identification and dynamic analysis of all types of modulation

ALL THESE APPLICATIONS . . .

• Noise, vibration, harmonic analysis • Filter & transmission line checks • Telemetry analysis • Communication System Monitoring . . . and more • Power Spectral Density Analysis (with Model PDA-1 Analyzer) • Frequency Response Plotting (with Model G-15 Sweep Generator)

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dependable CERTIFIED SPECIFICATIONS for accurate data

the pioneer is the leader

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SUMMARY OF SPECIFICATIONS

Frequency Range: 100 cps to 600 kc.
Sweepwidth: Variable, calibrated from 1 kc to 200 kc.
Center Frequency: Variable, calibrated from 0 to 500 kc.
Markers: Crystal controlled, 10 kc and 100 kc plus harmonics.
IF Bandwidth: Variable, 100 cps thru 4 kc.
Sweep Rate: Variable, 1 cps to 60 cps.
Amplitude Scales: Linear, 40 db log (extendable to 60 db) and 2.5 db expanded.
Sensitivity: 200 µv to 100 v full scale deflection.
Accuracy: ±0.5 db.
Input Impedance: 55 k ohms.

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Eighth National Communications Symposium

HOTEL UTICA AND MUNICIPAL AUDITORIUM, UTICA, N. Y., OCTOBER 1–3, 1962

The Eighth National Communications Symposium, sponsored by the Rome-Utica Section of the IRE Professional Group on Communications Systems, will be held at the Hotel Utica and the Municipal Auditorium, Utica, N. Y. on October 1–3, 1962. The program is as follows:

Monday, October 1

Session 1—RF Power Generation Techniques

"Triode Oscillators for Applications in the 10-Gc Region," *J. B. Quirk, General Electric Co., Owensboro, Ky.*

"Practical Design Techniques for Solid-State Microwave Generators," *D. O. Fairley, Lenkurt Electric Co., Inc., San Carlos, Calif.*

"High Power Solid-State Amplifier Developments," *M. I. Jacob, Westinghouse Electric Corp., Baltimore, Md.*

"Optical Maser Systems for Interplanetary Communication," *V. J. Corcoran and F. W. Griffith II, Griffith Electronics, Elmwood Park, Ill.*

Monday Afternoon

Session 2A—Active Satellite Systems

"Frequency and Time Control for Multiple-Access Synchronous Satellite Communication Systems," *D. P. Peterson and J. A. Stewart, General Telephone and Electronics Labs., Inc., Menlo Park, Calif.*

"A Medium-Capacity Fixed Ground Station for Satellite Communications," *W. Glomb and B. Cooper, ITT Federal Labs., Nutley, N. J.*

"Communications Link Between Synchronous Satellites," *C. M. Johnson, IBM Communications Center, Rockville, Md.*

"System Organization for General Communication Via Medium Altitude Satellites," *D. G. C. Luck, Radio Corporation of America, Princeton, N. J.*

Session 2B—Coding and Data Processing Techniques

"Speech Data Processing in Real Time," *H. A. Straight, Melpar, Inc., Falls Church, Va.*

"Estimation of Message Errors for Control of a Variable Rate Binary Communications System," *N. G. Davies, Defense Research Telecommunications Establishment, Ottawa, Ontario, Canada.*

"Performance vs Complexity of Some New Decoders for the Binary Erasure Channel," *M. E. Mitchell, General Electric Co., Ithaca, N. Y.*

"An Interative-Code Communications System With Noisy Delayed Decision Feedback," *I. Jacobs and A. Levesque, Sylvania Electric Products Inc., Waltham, Mass.*

Tuesday, October 2

Session 3A—Passive Satellite Communication Systems

"The Instrumentation of an Array of Reflectors for Passive Satellite Communications," *A. E. Ruvin and S. W. Gery, Airborne Instrument Lab., Melville, N. Y.*

"Performance of a Rake System Over the Orbital Dipole Channel," *P. Bello and H. Raemer, Sylvania Electric Products Inc., Waltham, Mass.*

"The Floyd Satellite Communications Terminal," *L. C. Parode, R. C. Winterbottom, and L. W. Wilson, Hughes Aircraft Corp., Los Angeles, Calif.; and A. Feiner, Rome Air Development Center, Rome, N. Y.*

"Floyd Satellite Communications Terminal Monopulse Tracking Receiver as a System Logic Element," *T. F. Haggai and D. E. Miller, Hughes Aircraft Corp., Los Angeles, Calif.; and G. F. Negus, Rome Air Development Center, Rome, N. Y.*

Session 3B—Classified

Moderator: *H. A. Wheeler, President, Wheeler, Labs., Inc., Great Neck, N. Y.*

"State of the Art—Application of Bionics (Machine Intelligence) to Communications Systems," *E. B. Carne, Manager, Advanced Computer Labs., Melpar, Inc., Falls Church, Va.*

"Survivable Low-Frequency Long-Range Communications Systems," *S. Sensiper, Director, Survivable Communications Systems Lab., Space Electronics Corp., Glendale, Calif.*

"Performance of Correlations Systems with Band Width Exceeding the Coherent Propagation Band Width," *R. L. Freuberg and S. Stein, Applied Research Lab., Sylvania Electronic Systems, Waltham, Mass.*

"Applications of Statistical Communications Theory," *B. Goldberg, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

A paper by *Colonel Harold Johnson, Headquarters, USAF (AFOAC).*

Additional papers to be announced.

Tuesday Afternoon

Session 4A—Tropospheric Scatter Techniques

"AN/FRC-68 Angle Diversity Communication System," *D. Surenian and I. A. Fantera, ITT Labs., Nutley, N. J.*

"Theoretical Study of Dual Rate Transmission Over Gaussian Multiplicative Circuits," *P. A. Bello and W. M. Cowan, Jr., Sylvania Electric Products Inc., Waltham, Mass.*

"Effects of Amplitude and Phase Fluctuations on the Performance of Digital Data Systems in Fading FDM-FM and FDM-SSB Systems," *H. D. Becker, Cornell Aeronautical Lab., Buffalo, N. Y.*

"Analysis of Random Access Discrete Address System," *H. Magnuski and W. D. DeHart, Motorola Inc., Chicago, Ill.*

"Multiple Order Diversity System," *W. S. Patrick and M. J. Wiggins, Martin-Marietta Corp., Orlando, Fla.*

Session 4B—Propagation

"Polarization Considerations in Space Communications," *B. C. Potts, Ohio State University, Columbus, Ohio.*

"The Optimum Bit Length for HF-Transmission," *H. Feige-Kollmann, Collins Radio Co., Newport Beach, Calif.*

"Underground Communications," *R. M. Wundt and D. A. Boots, Sylvania Electric Products Inc., Waltham, Mass.*

"Lithosperic Communications," *G. L. Brown and A. F. Gangi, Space-General Corp., El Monte, Calif.*

Wednesday, October 3

Session 5A—Antennas

"Microwave System Engineering Using Large Passive Reflectors," *Maj. M. L. Norton, USASRDL, Fort Monmouth, N. J.*

"A New Approach to Loss in Antenna Gain and Tropospheric System Design," *J. L. Levatich, The Bendix Corp., Baltimore, Md.*

"Far-Field Wide-Band Distortion Patterns of Antennas," *D. J. Lewinski and H. Rosenblatt, The Martin Co., Baltimore, Md.*

"Adaptive Antenna Systems for Communications," *R. G. Roush, Electronic Communications Inc., Timonium, Md.*

Session 5B—Classified

Panel Discussion: "Frequency Assignment vs Common Spectrum Channelization."

Moderator: *D. C. Ports, Jansky and Bailey, Inc., Alexandria, Va.*

Panelists: *W. B. Bruene, Collins Radio Co., Cedar Rapids, Iowa; J. F. Byrne, Motorola, Inc., Riverside, Calif.; L. A. de Rosa, ITT Communications, Inc., Paramus, N. J.; J. J. Downing, Lockheed Missiles and Space Div., Palo Alto, Calif.; D. Haas, The Martin Co., Baltimore, Md.; and R. L. San Soucie, Sylvania Electric Products, Inc., Waltham, Mass.*

Wednesday Afternoon

Session 6A—Military Communication Systems

"Airborne UHF Multiplex Communications System," *J. R. Mensch and C. C. Pearson, Electronic Communications, Inc., St. Petersburg, Fla.*

"Design Considerations for a Switched Mobile Communication System," *R. Steel, Motorola Inc., Chicago, Ill.*

"A New Tactical Transmission Facility" *J. Bailey and V. Colaguori, U. S. Army Signal Research and Development Lab., Fort Monmouth, N. J.*

"The AN/GSC-5( ) Digital Communications System," *V. Dowdell, R. Fox, L. Doubleday and C. Foster General Dynamics/Electronics, Rochester, N. Y., and Rome Air Development Center, Rome, N. Y.*

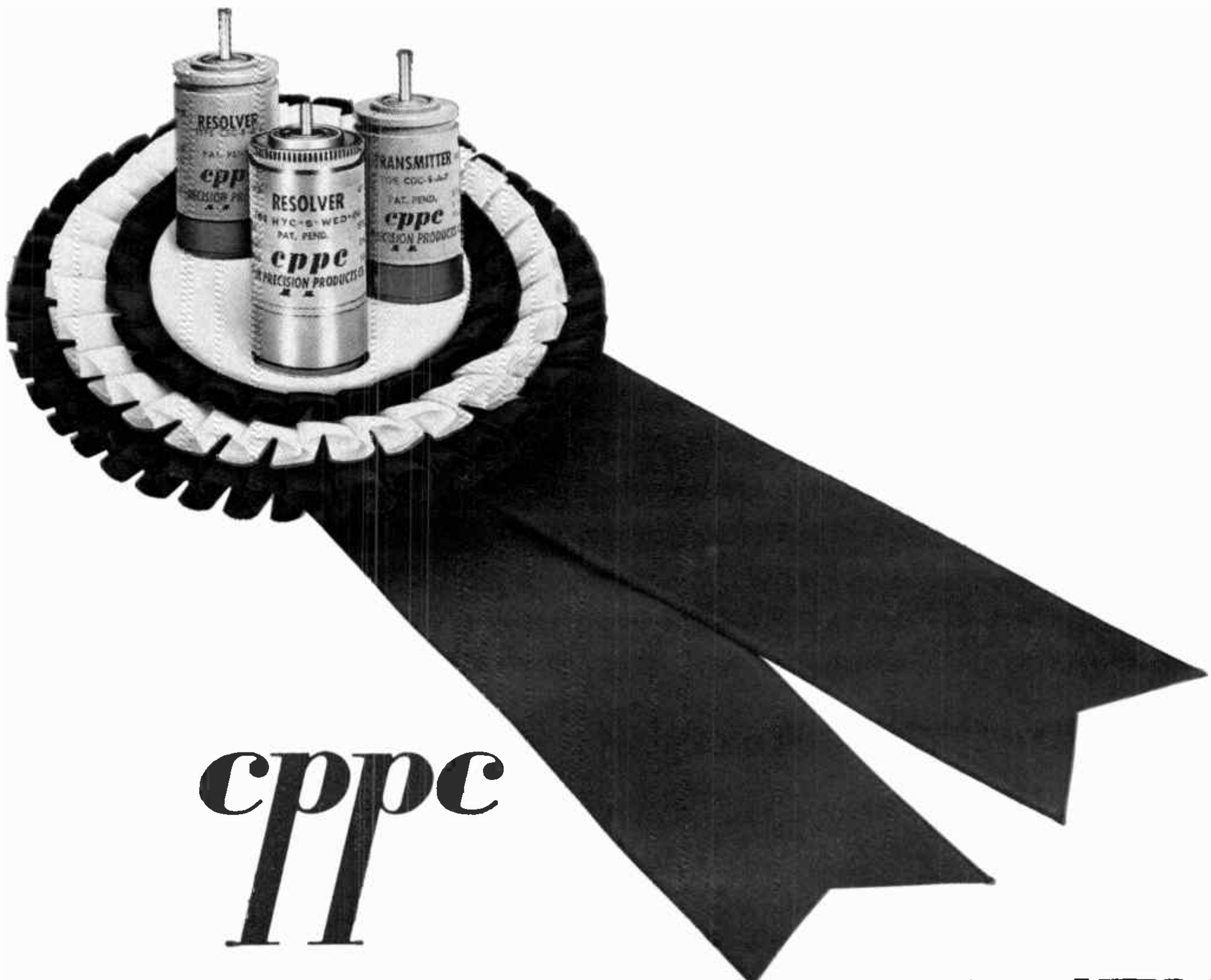
Session 6B—Receiving Techniques

"A Simple Adaptable Receiver with Pilot Tones for Analog Transmission Through a Selective Fading Channel," *D. A. Chesler, Sylvania Electric Products Inc., Waltham, Mass.*

"Optimum Decision and Scanning Techniques for Synchronization," *J. Z. Grayum, Philco Corp., Blue Bell, Pa.*

"Extending the Threshold of FM Receivers Using Feedback," *J. Page, General Electric Co., Syracuse N. Y.*

"A Phase-Locked Detector for FM Multiplex Applications," *D. W. Ford, Philco Corp., Philadelphia, Pa.*



cppc

continues to stress QUALITY

The temptation is present in these days of lowering prices and shrinking profit margins to build a "cheaper" product. CLIFTON PRECISION CONTINUES TO STRESS QUALITY.

In fact, we list herewith some recent improvements which make our rotating components more expensive to build. But they give you a better product.

As pioneers in the synchro and rotating components field, we think our years of experience in building a QUALITY product continue to give buyers a plus factor that they will not want to overlook.

QUALITY FEATURES

1. Improved high temperature resistant magnet wire is used in all synchro construction. Standard units now withstand in excess of 125°C.

2. Improved high temperature resistant epoxy impregnation of rotors and stators is used in all synchros and servo motors.

3. Higher temperature resistant silicon lubricants are used in all bearings.

4. High temperature resistant slot insulation in all synchros and servo motors permits repeated high potential testing with no deterioration of insulating characteristics.

5. Completely solderless brush construction eliminates cold solder joints.

6. Improved interlaminar insulation techniques give our synchros and servo motors lower power consumption due to core losses thereby giving same or better electrical performance with a cooler design.

7. Increased usage of gold alloys in critical areas of slip ring construction (including increased thickness) improves reliability and permits versatility of slip ring design.

CLIFTON PRECISION PRODUCTS CO., INC. cppc Clifton Heights, Pa.
Colorado Springs, Colo.

National Electronics Conference

McCormick Place, Chicago, Ill., October 8–10, 1962

The 1962 National Electronics Conference, sponsored by the IRE, AIEE, and other societies, will be held at McCormick Place, Chicago, Ill., on October 8–10, 1962. The program is as follows:

Monday, October 8

Session 1—Energy Beams as Working Tools I

Chairman: *P. B. Myers, Martin Company.*

"Electron, Ion, and Light Beams as Present and Future Materials Working Tools," *Helmut Schwarz and A. J. DeMaria, United Aircraft Corporation, Research Labs.*

"Recent Electron-Optical Developments in the Recording Field," *P. H. Gleichauf, General Electric Company.*

"Ion Beam Formation and Control for Reflected Ion Beam Deposition," *L. R. Bittman and J. Litton, Jr., Martin Company.*

"Application of Electron Beam Techniques to Electronics," *J. W. Meier, Hamilton Standard Div., United Aircraft Corporation.*

"Electron Beam Welding Applied to Today's Technology," *Robert Bakish, Electronics and Alloys, Inc.*

Session 2—University Presentations I

Chairman: *W. B. Boast, Iowa State University.*

Presentations delineating university-industry cooperation, current interdisciplinary programs of the universities, in-depth discussions of university research programs, and discussion of intensive short-course opportunities for practicing engineers. Each university will be allotted approximately 45 minutes for its presentation.

Session 3—Digital Computer Applications and Components

Chairman: *R. O. Sather, Wayne State University.*

"Basic Principles of Some Pattern Recognition Systems," *L. Kanal, F. Slaymaker, D. Smith, and W. Walker, General Dynamics/Electronics.*

"Electronics of a Precision Comparator," *W. Heacock, Fairchild Camera and Instrument Corporation.*

"On the Digital Computer Classification of Geometric Line Patterns," *Herbert Freeman, New York University.*

"Statistical Techniques in Circuit Optimization," *C. Clunies-Ross and S. S. Husson, Data Systems Division, International Business Machines Corporation.*

"Design of Digital Control Systems," *K. S. Fu and R. M. Kline, Purdue University.*

Session 4—Solid State Applications—Varactors

Chairman: *A. L. Aden, Motorola, Inc.*

"Parametric Amplification by Phase Modulation," *D. K. Adams, Cooley Electronics Laboratory, University of Michigan.*

"Varactor Fabrication for Microwave Applications," *W. F. Epperly, American Electronics Laboratories.*

"Solid-State Microwave Signal Sources Using Varactor Harmonic Generation," *M. E. Hines, Microwave Associates, Inc.*

"Cascading Low-Gain Parametric Amplifier Stages," *Carl Blake, Lincoln Laboratory, Massachusetts Institute of Technology.*

"High Order Broad-Band Varactor Multiplier," *R. J. Bauer, Aircraft Armaments, Inc.*

Session 5—Reliability—Are We Spending Our Money Wisely?

Chairman: *George Rappaport, Warnecke Electronic Tubes, Inc.*

Six to eight panelists, speaking 10 minutes apiece, followed by a question and answer session under the guidance of a moderator (chairman). Panelists represent a variety of informed positions questioning the complacency with which reliability attainment is related to increased dollar spending. It is expected that the panel will be a controversial airing of the key question of how reliability efforts can be evaluated and justified economically.

Monday Afternoon

Session 6—Energy Beams as Working Tools II

Chairman: *L. R. Bittman, Martin Company.*

"Electron Beam Phenomena Associated with Perforated Wall Hollow Cathode Discharges," *H. L. L. vanPassen, E. C. Muly, and R. J. Allen, Martin Company.*

"Cold Hollow Cathode Discharge Welding," *E. C. Muly, H. L. L. vanPassen, and R. J. Allen, Martin Company.*

"Some Aspects of Laser Beam Welding," *R. L. Martin, Technical Research Group, Inc.*

"High Power Laser for Welding Applications," *G. W. Dunlap and David Williams, General Electric Company.*

"Metallurgical Applications of Lasers," *R. D. Engquist and C. J. Bahun, Hughes Aircraft Company.*

Session 7—University Presentations II

Chairman: *H. W. Farris, University of Michigan.*

Presentations delineating university-industry cooperation, current interdisciplinary programs of the universities, in-depth discussions of university research programs, and discussion of intensive short-course opportunities for practicing engineers. Each university will be allotted approximately 45 minutes for its presentation.

Session 8—Programed Education

Chairman: *J. J. Gershon, DeVry Technical Institute.*

"Teaching Machines in Programed Instruction," (tutorial) *L. M. Stolurow, University of Illinois.*

"Undergraduate EE Via Video Tape and Closed Circuit TV," *W. H. Hayt, Jr., Purdue University.*

"Use of Closed Circuit Television in Graduate Teaching," *D. L. Dietmayer, R. A. Greiner, V. C. Rideout, and W. B. Swift, University of Wisconsin.*

"Description and Use of a Computer-Controlled Teaching System," *D. L. Bitzer and P. Braunfeld, Coordinated Science Laboratory, University of Illinois.*

Session 9—Microwave Applications

Chairman: *J. A. Boyd, Radiation Incorporated.*

"A Sum and Difference Interferometer System for HF Radio Direction Finding," *A. D. Bailey and W. C. McClurg, University of Illinois.*

"Analysis of the Miniaturization of Resonant and Nonresonant Antennas Utilizing High 'Q' Materials," *J. A. M. Lyon, A. T. Adams, and R. M. Kalafus, Cooley Electronics Laboratory, University of Michigan.*

"Pattern Gain of Phased Arrays," *R. K. Thomas, Martin Company.*

"Efficiency, Phase Shift and Power Limiting in Variable-Pitch Traveling-Wave Amplifiers," *J. E. Rowe and C. A. Brackett, Electron Physic Laboratory, University of Michigan.*

"Microwave Modulation of Light with ADP," *M. C. Watkins, Aircraft Armaments, Inc.*

Session 10—Solid State Applications—General

Chairman: *H. W. Katz, General Electric Company.*

"Transient Response of Forward Biased Diffused *P-N* Junctions," *H. K. Cooper, Pacific Semiconductors, Inc.*

"A General Synthesis of Tunnel Diode Amplifiers and Sensitivity Minimization," *B. A. Shenoi, University of Minnesota.*

"A New Feedback Broad-Banding Technique for Transistor Amplifiers," *M. S. Ghausi, New York University, and D. O. Pederson, University of California.*

"Asymmetrical Scattering from a Ferrite Cylinder," *J. C. Palais, Cooley Electronics Laboratory, University of Michigan.*

Session 11—The Consultant's Role in Research and Development

Chairman: *D. C. Strain, Electro-Scientific Industries.*

A panel of executive representatives of leading consulting firms.

Session 12—EUROMART

Chairman: *W. C. Kotteman, Illinois Bell Telephone Company.*

A session sponsored by the International Activities Committee on the subject of EUROMART and its effect on electronics in the United States will be presented.

Tuesday, October 9

Session 13—Modulation Theory I

Chairman: *W. R. Bennett, Bell Telephone Laboratories.*

"On Comparing the Modulation Systems," *J. C. Hancock, Purdue University.*

"Hilbert Transforms and Modulation

Theory," *F. F. Kuo and S. L. Freeny, Bell Telephone Laboratories.*

"Signal-to-Noise Effects and Threshold Effects in FM," *Mischa Schwartz, Polytechnic Institute of Brooklyn.*

"What Does Circuit Theory Have to Do with PCM?" *M. R. Aaron, Bell Telephone Laboratories.*

Session 14—Digital Computer Workshop I

Chairman: *W. J. Eccles, Purdue University.*

"Organization of a Digital Computer," *D. T. Herrman, Jr., General Precision, Inc.*

"Programming of PINT," (tutorial) *W. J. Eccles, Purdue University.*

"Problem-Oriented Languages," *C. H. Davidson, University of Wisconsin.*

Session 15—Adaptive Systems

Chairman: *G. E. Karres, Martin Company.*

"Design Capabilities of Model Reference Adaptive Systems," *H. P. Whitaker, Massachusetts Institute of Technology.*

"Convergence Properties of a Model Reference Adaptive Control System from a Simple Stability Criterion," *J. Bongiorno, Polytechnic Institute of Brooklyn.*

"Aerospace Vehicles and Adaptive Flight Control" *Milton Reed, Minneapolis-Honeywell Regulator Company.*

"Pulse Frequency Modulation for Adaptive Control," *Gordon Murphy, Northwestern University, and R. L. West, McDonnell Aircraft Company.*

"Adaptive Learning Systems," *J. E. Gibson, Purdue University.*

Session 16—Medical Electronics

Chairman: *George Zacharopolous, Lafayette Clinic.*

"Design Toward a Chronic Artificial Heart," *V. W. Bolie and Jacob Kline, Iowa State University.*

"Medical Electronics at the Mayo Clinic," *R. J. Hansen, Mayo Clinic.*

"Comments Upon Problems Encountered in Electronic Monitoring of Uncooperative Patients," *J. T. Martin, M.D., Mayo Clinic.*

"A Review of Air Ionization and Its Effects on Living Systems," *H. F. Schulte, Jr., University of Michigan.*

"A Computer System for Hospital Medical Record Data," *V. N. Slee, M.D., Professional Activities Study.*

Session 17—Microelectronics I

Chairman: *A. P. Stern, Martin Company.*

"Design of Integrated Radio Frequency Amplifiers," *Glen Madland, Motorola, Inc.*

"Using Decision Theory Techniques for Optimum Selection of Thin Film Packaging Concepts," *D. L. Brisendine, Martin Company.*

"Design and Performance of Tantalum Thin Film Circuits," *P. Thomas, J. S. Ekiss, J. Roschen, and M. Walker, Philco Corporation.*

"Thin Film Active Devices," *W. Tantraporn and K. K. Reinhartz, General Electric Company.*

"Integrated Circuitry Embodying Thin Film Passive and Active Components," *T. E. Harr, Martin Company.*

Session 18—University Presentations III—Comments from Industry

Chairman: *To be announced.*

Presentations delineating university-industry cooperation, current interdisciplinary programs of the universities, in-depth discussions of university research programs, and discussion of intensive short-course opportunities for practicing engineers. Each university will be allotted approximately 45 minutes for its presentation. In this session an opportunity will be afforded an industry representative to picture the needs of industry in these cooperative endeavors.

Tuesday Afternoon

Session 19—Modulation Theory II

Chairman: *J. B. Cruz, Jr., University of Illinois.*

"Transient Response of Narrow-Band Networks to Angle Modulated Signals," *J. J. Hupert, DePaul University.*

"Transmission of FM Signals Through Linear Filters," *D. T. Hess, Polytechnic Institute of Brooklyn.*

"Frequency Feedback Demodulators," *L. H. Enloe, Bell Telephone Laboratories.*

"Optimum Coherence Demodulation for Continuous Modulation Systems," *A. J. Viterbi, Jet Propulsion Laboratory, California Institute of Technology.*

"Phase-Lock Demodulators," *B. J. Miller and L. L. Kocsis, Zenith Radio Corporation.*

Session 20—Digital Computer Workshop II

Chairman: *W. J. Eccles, Purdue University.*

"Organization of a Digital Computer," *D. T. Herrman, Jr., General Precision, Inc.*

"Programming of PINT," *W. J. Eccles, Purdue University.*

"Problem-Oriented Languages," *C. H. Davidson, University of Wisconsin.*

Session 21—Timely Aspects of Space Science

Chairman: *F. B. Llewellyn, University of Michigan.*

"Overseas Telephone and Television Transmission," *author to be announced.*

"Navigation Satellite Progress," *A. B. Moody, National Aeronautics and Space Administration.*

"Weather Satellites," *S. Fritz, U. S. Weather Bureau.*

"Astronomy with the Aid of Satellites," *F. T. Haddock, Radio-Astronomy Laboratory, University of Michigan.*

Session 22—The Role of R and D in Future Profits

Chairman: *T. F. Jones, Jr., University of South Carolina.*

This session will review the successful techniques employed by members of the electronics industry in achieving a good return on their investment in R and D.

Session 23—Microelectronics II

Chairman: *J. R. Black, Motorola, Inc.*

"Thin Film Technologies for Electronic Components," *W. D. Fuller, Lockheed Missiles and Space Company.*

"The Impact of Microelectronics and Solid State Technology on Electro-Mechanical Control Systems," *T. Mitsutomi and W. F. DeBoice, Autonetics.*

"Semiconductor Networks for Use in Electro-Mechanical Control Systems," *C. Abbott, L. Bohan, L. Housey, and L. Regis, Texas Instruments, Inc.*

"A Novel Solution to the Interconnection Problem in Microsystem Circuits," *T. L. Robinson, Cornell Aeronautical Laboratory, Inc.*

"Thin Film Integrated Components for Telemetry Subsystems," *A. J. Nichols and W. D. Fuller, Lockheed Missiles and Space Company.*

Session 24—The Role of the Universities in Industrial Assistance

Chairman: *F. R. Bacon, Jr., University of Michigan.*

The present activities of Midwest universities to stimulate industrial growth in the Midwest.

Session 25—High School Student Program

Chairman: *To be announced.*

Wednesday, October 10

Session 26—Advanced Computer Technology

Chairman: *F. P. Diemer, Martin Company.*

"Research Input-Output Equipments for General Purpose Computers," *W. S. Holmes and H. M. Maynard, Cornell Aeronautical Laboratory, Inc.*

"A Bit Oriented Sequential Access Memory," *C. H. Fischer, Hughes Aircraft Company.*

"Two-Dimensional Spatial Filtering Research in General Purpose Computers," *W. D. Fryer and G. E. Richmond, Cornell Aeronautical Laboratory, Inc.*

"A Zero Loss Capacitor Storage Device," *C. H. Fischer, Hughes Aircraft Company.*

"HCM-202 Thin Film Computer," *M. M. Dalton, Hughes Aircraft Company.*

Session 27—Research Previews I

Chairman: *D. S. Gage, Northwestern University.*

National Electronics Conference is including in its 1962 program a type of session new to the Electronics field—sessions of short (10 minute) papers. The purpose of these sessions is to include information concerning the most recent developments in industrial and university laboratories.

Session 28—Infrared Applications I

Chairman: *F. J. Kocsis, Jr., Servo Corporation of America.*

"A Rugged, Low-Noise Solid-State Infrared Detection System," *M. C. Baum, Mnemotron Corporation.*

"Terrain Mapping by Use of Infrared Radiation," *D. E. Harris and C. L. Woodbridge, HRB-Singer, Inc.*

"An Infrared Signal Generator," *Arthur Glaser and Allan Ross, Telewave Laboratories, Inc.*

"Infrared Hot Box Detectors," *W. M. Pelino, Servo Corporation of America.*

Session 29—Circuit and System Theory

Chairman: *S. Louis Nakimi, Northwestern University.*

NUMBER 24—NEW PRODUCT SERIES

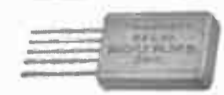
New Bourns Subminiature Relay— Its Reliability Is as High as Its Size Is Small

You can see that it's little, and you can bet that it's reliable. Only .2" x .4" x .6", but a steady performer even at 40 G, 55-2000 cps, this subminiature SPDT relay is designed to meet all environmental requirements of MIL-R-5757D. Its features include single-coil design, rotary balanced armature, hermetically sealed case and self-cleaning long-life contacts. Efficient coil design and packaging improve sensitivity to just 100 mw maximum.

By subjecting every unit to a 5000-operation run-in, Bourns precludes the possibility of relay "infant mortality." To further ensure consistent quality, Bourns conducts 100% final inspection for all important relay characteristics including mass spectrometer leak testing. The last and most punishing test of quality is the trip taken by monthly samples through the Bourns Reliability Assurance Program. This is one of the most extensive series of electrical and environmental tests in the electronics industry, and has long been the reliability double-check for the famous Trimpot® potentiometer. With Bourns relays, as with Bourns potentiometers, every possible step is taken to see that the quality you specify is the quality you get.

Units are available now from the factory, and will soon be available through Bourns distributors as well. Write for complete technical data.

Size: .2" x .4" x .6"
Maximum operating temperature: 125°C
Contacts: SPDT; Rating: 1.0 amp
resistive, 26.5 VDC
Coil resistances: 50Ω to 2000Ω
Pick-up sensitivity: 100 milliwatts
Vibration: 40 G standard, 60 G special
Shock; 150 G


ACTUAL SIZE

Manufacturer: Trimpot® potentiometers; transducers for position, pressure, acceleration. Plants: Riverside, California; Ames, Iowa; and Toronto, Canada

"Theoretical Basis and Practical Implications of Band-Pass Sampling," *C. L. Ackerman, C. S. Miller, and J. L. Brown, Jr. Pennsylvania State University.*

"Discrete Orthonormal Exponentials" *T. Y. Young and W. H. Huggins, Johns Hopkins University.*

"Adaptive System Identification by State Variable Operations," *H. M. Estes, U. S. Air Force Academy.*

"Analysis and Design of Sampled-Data Systems Via State Transition Flow Graphs," *B. C. Kuo, University of Illinois.*

"Miyata's Method Applied to Active Network Synthesis," *R. E. Thomas, U. S. Air Force Academy.*

Session 30—Aerospace Control Systems

Chairman: *Norman Winter, Motorola, Inc.*

"Air Space System Design and the Resolution of Conflicts," *D. F. Babcock, Stanford Research Institute.*

"Air Traffic Control System Plans," *Neal Blake, Federal Aviation Agency, Aviation Research Service, System Design Team.*

"Flight Line of the Future," *Harold Johnson, Directorate of Telecommunications, Air Staff, USAF.*

Panel discussion.

Session 31—Electronics in Hydrospace

Chairman: *R. L. Miller, University of Chicago and Woods Hole Oceanographic Institute.*

"New Concepts for Intense Sound Transducers," *R. R. Whymark, Armour Research Foundation.*

"Airborne Bathythermograph System," *G. Gruener and A. Leumpert, Sparton Electonics.*

"Electronic Instrumentation for the Great Lakes Water Quality Study," *J. L. Verber, U. S. Public Hea th Service.*

"The Hall Compass" *E. A. Keller, Motorola, Inc.*

"Radio Communication Within the Hydrosphere," *R. C. Becker, Amphenol-Borg Electronics Corporation.*

Session 32—Communication Systems

Chairman: *Ira Jacobs, Bell Telephone Laboratories.*

"Decision Theory," *J. C. Hancock, Purdue University.*

"Signal Design," *R. M. Lerner, Lincoln Laboratory, Massachusetts Institute of Technology.*

"Adaptive Systems," *L. S. Schwartz, New York University.*

"Coding," *I. M. Jacobs, Massachusetts Institute of Technology.*

Wednesday Afternoon

Session 33—Infrared Applications II

Chairman: *T. E. Harr, Martin Company.*

"Some Electronics Problems in Infrared Systems Design," *F. G. Whelan, Martin Company.*

"Optical Techniques for Target Enhancement and Background Rejection," *R. S. Neiswander, The TE Company.*

"Recent Advances in Infrared Detectors for the 8.5–13.5 Micron Spectral Band," *J. K. Lennard, Martin Company.*

"Status Report on Thermistor Bolometers," *I. J. Melman and I. M. Meltzer, Servo Corporation of America.*

Session 34—Signal Theory

Chairman: *Herbert Sherman, Lincoln Laboratory, Massachusetts Institute of Technology.*

"The Space of Essentially Time and Band-Limited Signals," *H. O. Pollak, Bell Telephone Laboratories.*

"Orthonormal Exponentials," *D. C. Ross, IBM Federal Systems Division.*

"A Note on Orthogonal Digit Coding," *L. Kurz, New York University.*

"Two Dimensional Signal Representation Using Prolate Spheroidal Functions," *D. A. Landgrebe and G. R. Cooper, Purdue University.*

Session 35—Trends in Aerospace Ground Equipment

Chairman: *Harold Flowers, Avco Corporation.*

"Analog Checkout of Large Systems—The Digital Solution," *Leo Kaye, Martin Company.*

"Digital Computer Checkout of Inertial Navigation Systems," *J. L. Henry, Autonetics.*

"Use of Integrated Circuitry in Digital System," *Lloyd Thayne, Martin Company.*

"A Simulation—Calibration System for Space Flight, Landing and Rendezvous Control Systems," *W. J. Hollandswirth, Missouri Research Laboratories, Inc.*

Session 36—Engineering Writing and Speech

Chairman: *A. A. Canfield, Bendix Corporation.*

"The Conditions of Communication," *J. D. Chapline, Philco Computer Division.*

"Reports as a Measure of Competence," *W. B. Dennen, Radio Corporation of America.*

"Better Report Writing Pays," *R. M. Woelfle, Bendix Mishawaka Division.*

"The Poor Writer," *R. G. Marolf, Research Laboratories, General Motors Corporation.*

Session 37—Research Previews II

Chairman: *D. S. Gage, Northwestern University*

National Electronics Conference is including in its 1962 program a type of session new to the Electronics field—sessions of short (10 minute) papers. The purpose of these sessions is to include information concerning the most recent developments in industrial and university laboratories.

Session 38—New Components, New Techniques

Chairman: *D. G. Dow, Varian Associates.*

"The Avalanche Injection Diode and Its Application as a Switch for High Frequency Signals," *J. R. Szedon and A. G. Jordan, Carnegie Institute of Technology.*

"An Anti-Storage Clamp and a Method of Increasing I_p/I_v Ratio of Tunnel Diodes," *W. T. Rhoades, Hughes Aircraft Company.*

"Design of a High Performance *S*-Band Varactor Frequency Multiplier," *L. K. Staley, The Bendix Corporation.*

"Ceramic Band-Pass Filter with Unsymmetric, Tuned Hybrid Lattice Structure," *F. L. Sauerland, Clevite Corporation.*

"A Study of Optimum Switching of On-Off Control Systems Through Logic," *O. I. Elgerd, University of Florida, and L. B. Scheiber, AC Spark Plug Division, General Motors Corporation.*

Session 39—Microwatt Control of Megawatt Systems

Chairman: *To be announced.*

Information regarding opportunities for electronics engineers in the power systems field will be presented.

4 NEW ALFRED MODULATORS

...Especially designed to increase the versatility of your microwave oscillators and amplifiers.

SQUARE WAVE GENERATOR
(MODEL 305A)

Provides 0 to 60 volts peak for modulating microwave amplifiers, signal generators and TWT amplifiers, and for checking response of wide band amplifiers.

BRIEF SPECIFICATIONS: Frequency, 850 to 1150 cps; rise and fall time, 0.2 μsec rise, 2 μsec fall; output impedance, less than 2500 Ω; price, $90.

PULSE GENERATOR
(MODEL 5-6826P)

Designed for use with microwave amplifiers requiring 300 to 500 volts pulse signal for "ON-OFF" gating or modulation. Provides adequate pulse voltage and current for modulating high voltage grids and control electrodes.

BRIEF SPECIFICATIONS: Pulse output, 300 to 500 v positive, adjustable by panel control; pulse duration, 1 to 12 μsec, adjustable by panel control; rise and fall time, 0.5 μsec or less when load capacitance less than 100 μμf; repetition rate, internal, 10 to 10,000 pps; external, 0 to 10,000 pps by application of any positive going signal 20 v peak or greater; price, $490.

SAWTOOTH GENERATOR
(MODEL 325A)

A compact, general purpose source for sawtooth modulating helices of traveling wave tubes for generating doppler shift. Minimum side band energy content. Both positive and negative going sawtooth voltage signals are simultaneously available, allowing simultaneous operation of two microwave amplifiers.

BRIEF SPECIFICATIONS: Output, negative and positive going sawtooth; amplitude, 0 to 60 v peak-to-peak independently adjustable by panel control; frequency, 0.1 cps to 1 mc; flyback, less than 1% of period or 70 nanoseconds, whichever is greater; minimum output load, 5 K, 100 μμf maximum; price, $990.

PULSED SAWTOOTH GENERATOR
(MODEL 328B)

When used with Alfred Microwave Oscillators produces frequency scanned pulsed RF signals for simulating pulse compression radar signals. Produces 3 video signals, two pulsed sawtooth voltages differing only in dc value, and a coincident gate pulse. Sawtooth signals may be applied to cathode and anode of the BWO in Alfred Microwave Oscillators to develop change in frequency. Gate pulse may be applied to a crystal switch or pulse modulatable microwave amplifier to key RF signal "ON" during frequency sweep.

BRIEF SPECIFICATIONS: Output sweep width, continuous linear adjustment from 1 to 30 μsec; sweep amplitude, continuously variable from 0 to 50 v; sweep slope, negative or positive; sweep output, two simultaneous sawtooth waves of equal amplitude during pulsed sweep interval. Output #1: (Cathode of BWO), 0 v (ground potential), 15 ma peak. Output #2: (Anode of BWO), 0 to +250 v relative to output #1, 50 μa peak; gate, positive pulse during sweep. Gate rise and fall time 0.2 μsec max. "OFF" voltage adjustable —100, —50, —25 v; "ON" voltage 0 to +4 v, @ 0 to 60 ma. Price, $1,290.

NEW 1962 CATALOG AVAILABLE

For your copy of our new 16 page full line catalog and complete technical bulletins on these modulators, call your Alfred engineering representative or write us directly.

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12

International Symposium on Space Phenomena and Measurement

STATLER-HILTON, HOTEL, DETROIT, MICH., OCTOBER 15–18, 1962

The ninth annual meeting of the IRE Professional Group on Nuclear Science, the International Symposium on Space Phenomena and Measurement, will be held at the Statler-Hilton Hotel, Detroit, Mich., on October 15–18, 1962. It is co-sponsored by NASA and AEC. The program of the Symposium is as follows:

Session I—Review Session—Recent Satellite Research on Energetic Particles and Magnetic Fields

"Solar Cosmic Radiation," *Dr. John Winckler, University of Minnesota.*

"A Van Allen Radiation Belt Progress Report," *Dr. W. N. Hess, Goddard Space Flight Center.*

"A Study of the Outer Geomagnetic Field," *Dr. L. J. Cahill, University of New Hampshire.*

"Galactor Cosmic Radiation," *Professor Peter Meyer.*

Session II—Review Session—Recent Results in Astronomy and Interplanatary Plasma Physics Research

"Results of the Ionospheric Physics Program and its Impact on Future Program," *Dr. R. E. Bordeaux, Goddard Space Flight Center.*

"Interplanatary Plasma," *Dr. Herbert Bridge, Massachusetts Institute of Technology.*

"Gamma Ray Astronomy," *Dr. G. G. Fazio.*

Speaker to be announced.

Session III—Effects of Space Radiation Environment on Semiconductor Components

Chairman: *W. C. Scott, NASA.*

Session IV A—Scientific Instrumentation and Measurements for Space Exploration

Chairman: *George Ludwig, NASA.*

"A Combination Scintillator and Solid Radiation Detector," *Mr. Cline.*

"A 2-Parameter 256-Channel Pulse Height Analyzer for Space Use," *Mr. Way, Ludwig Kampko.*

"A Logarithmic Digital Amplifier," *S. Jones.*

Session IV B—Nuclear Instrumentation

Chairman: *Alex Stripeika.*

"Latest Developments in Solid-State Detectors for Particle Counting," *S. Friedland.*

"Summary of Work in Circuits for Picosecond Range," *Q. Kerns.*

"Summary on Characteristics and Stability of Logarithmic Devices and Amplifiers," *E. Sikorski.*

"Instrumentation for Pluto Test Reactor and its Engineering Problems," *G. S. T. Leger Barter.*

Session V A—Space Environment

Chairman: *J. E. Kupperian, NASA.*

Session V B—Plasma Diagnostics

Chairman: *William Kerr, University of Michigan.*

Session VI A—Detection of Space and High Altitude Atmospheric Tests

Chairman: *Dr. Alois Schardt, ARPA.*

"Satellite Based Detection Systems for Space Tests," *Dr. Richard Paschek, Los Alamos.*

"Near Earth Satellite Radiation Environment," *Dr. S. Bloom, Lawrence Radiation Laboratory.*

"Ground Based High Altitude Detection System," *author be announced.*

"Radio Wave Phase Shift as a Detection Method," *Dr. Glenn Gean, NBS, Boulder, Colo.*

Session VI B—Detection of Underground Tests and Debris from Atmospheric Tests

Chairman: *Dr. Alois Schardt ARPA.*

"Detection of Debris From Atmospheric Tests," *Dr. Lester Machta, USW Bureau, Washington.*

"Sersmic Detection Methods for Underground Tests," *Dr. Karl Romney, AF Technology Application Center, Washington.*

"On Site Inspection for Underground Tests," *Dr. Charles Bates, ARPA.*

Third New York Conference on Electronic Reliability

STEVENS INSTITUTE OF TECHNOLOGY, HOBOKEN, N. J., OCTOBER 19, 1962

The Third New York Conference on Electronic Reliability, sponsored by PGRQC, PEP and CP, will be held on October 19, 1962 at Stevens Institute of Technology, Stevens Center, Hoboken, N. J. The theme of this year's Conference is "Reliability Engineering in Practice." The program is as follows:

Session IA

"The Practical Significance of Reliability Prediction," *C. R. Thomas, V. Selman, and B. Ellison, IEC.*

"Quality Control—A Small Plant's Philosophy," *L. Kuferman, Filtors, Inc.*

"SARADA, IDEP, GMIDEP—Part Reliability Data Exchange Programs Operated by NOL, Corona," *S. Pollack, Naval Ordnance Labs.*

Session IB

"Role of the Reliability Organization in the Value Engineering Program," *M. Tall, RCA.*

"Personnel and Reliability," *F. Beachler, Bendix Corp.*

"Microminiaturization Vs Standard Packaging—Comparative Reliability," *J. P. Marone, Jr., P. R. Mallory and Co., Inc.*

Session IIA

"Calculating the Reliability of Multimodal Systems," *K. N. Curtin, RCA.*

"Logical Analysis of Redundant Reliability Flow Networks," *S. W. Leibholz, Auerbach Corp.*

"Graphical Methods of Solving Reliability Logic Configuration," *W. E. Marshall, Minneapolis-Honeywell.*

"Application of Flow Graphs to the Solution of Reliability Problems," *Dr. W. W. Happ, Lockheed Missiles and Space Co.*

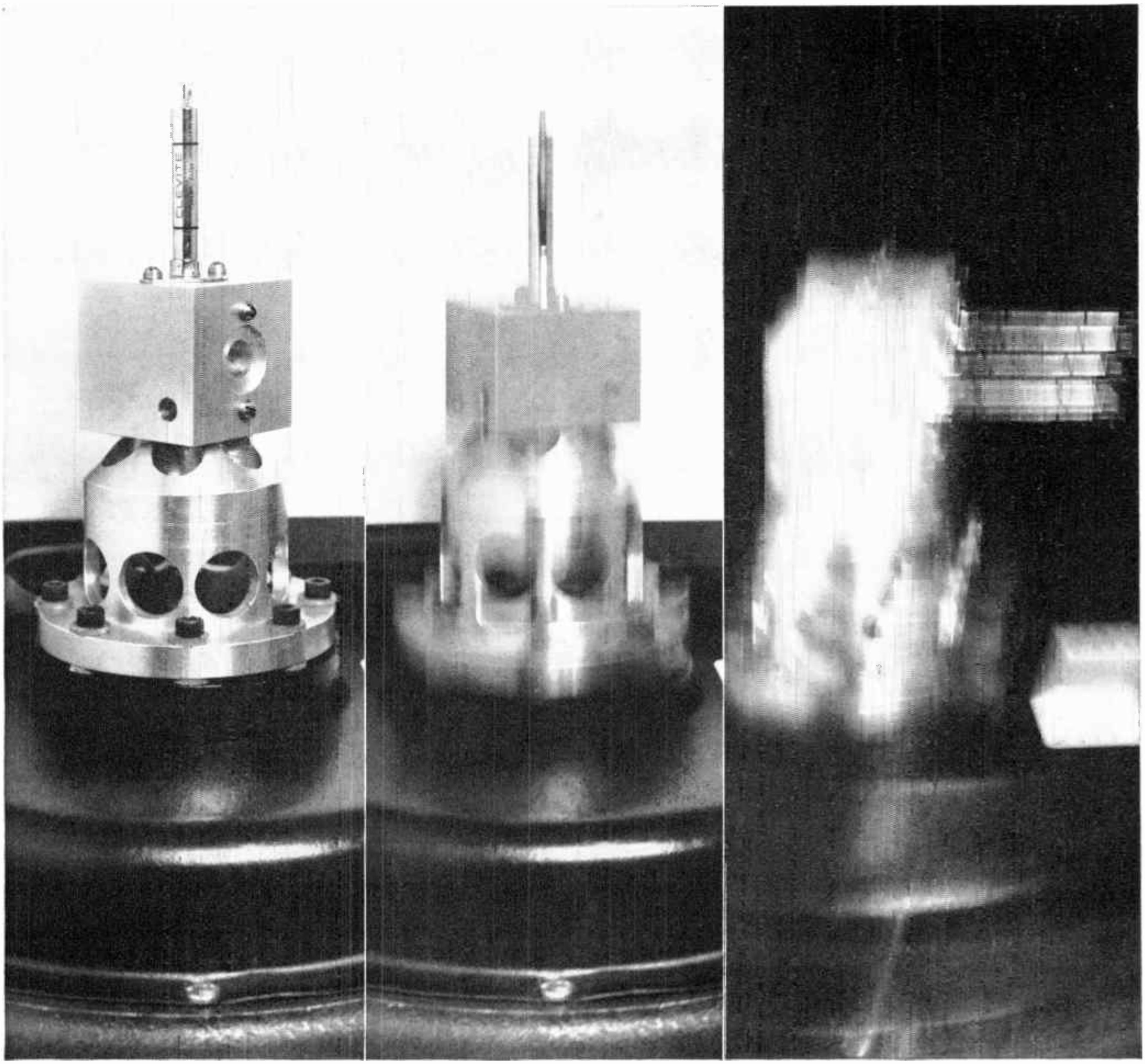
Session IIB

"Application of Darnell Report in Component Part Specifications," *G. Wyler, Battelle Memorial Institute.*

"In-Process Control for High Reliability Component Parts," *W. R. Arnold, Vitramon.*

"Application of Component Parts in Military Equipment," *author to be announced.*

"Reliability Vs Delivery Time in Component Parts," *author to be announced.*



Ninth Annual East Coast Conference on Aerospace and Navigational Electronics

EMERSON HOTEL, BALTIMORE, MD., OCTOBER 22–24, 1962

The Ninth Annual East Coast Conference on Aerospace and Navigational Electronics will be held at the Emerson Hotel, Baltimore, Md., on October 22–24, 1962. The Conference is sponsored by the Baltimore Section of the IRE Professional Group on Aerospace and Navigational Electronics. The program is as follows:

Monday, October 22

Session 1.1—Microminiaturization I

Moderator: *Dr. R. E. Thun, IBM Labs., Neighborhood Road, Kingston, N. Y.*

"Microminiaturization 1940—?, A Survey," *T. E. Harr, Electronic Systems and Products Div., The Martin Company, Baltimore, Md.*

"Distributed RC Networks for Approximately Maximally Flat Narrow-Band Amplifiers," *R. W. Wyndrum, Jr., Instructor, and G. J. Herskowitz, Instructor, Dept. of Electrical Engineering, New York University, New York, N. Y.*

"Band-Pass Amplifiers Using Distributed Parameter Networks," *G. C. Riddle, Electronic Sciences Lab., Microsystems Electronics, Lockheed Missiles and Space Company, Palo Alto, Calif.*

"A Thin-Film Frequency Discriminator," *P. S. Castro, Microsystems Electronics Dept., Electronic Sciences Lab., Lockheed Missiles and Space Company, Palo Alto, Calif.*

"Effect of Cathode Geometry on the Novel Electron-Beam Mode Discharge," *H. L. L. van Paassen and E. C. Muly, Electronic Systems and Products Div., The Martin Company, Baltimore, Md.*

Session 1.2—Navigation

Moderator: *Dr. I. Kanter, RCA, Moorestown, N. J.*

"A Second Look at Intertial Navigation for Aircraft," *E. R. Dayton, Assistant Engineering Division Manager, Instrument Div., Lear-Siegler, Inc., Grand Rapids, Mich.*

"Considerations of Airborne Transit Navigation," *J. D. Campbell and R. L. Hovious, Light Military Electronics Dept., General Electric Company, Utica, N. Y.*

"A Continuous Satellite Navigation and Guidance System," *T. W. Godbey and A. W. Roeder, Light Military Electronics Dept., General Electric Company, Utica, N. Y.*

"Spaceborne Radar for Rendezvous Guidance and Lunar Landings," *D. B. Newman, Assistant Director of Advanced Design, and H. E. Prew, Senior Staff Engineer, Electronic Systems Div., Fairchild Stratos Corporation, Wyandanch, N. Y.*

"Star Field Recognition for Space Vehicle Orientation," *S. S. Viglione and H. F. Wolf, Astropower, Inc., Douglas Aircraft Company, Costa Mesa, Calif.*

Monday Afternoon

Session 2.1—Lasers

Moderator: *Dr. D. F. Nelson, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"A Review of Principles and Developments in the Field of Optical Masers," *Dr. P. P. Sorokin, IBM Research Center, Yorktown Heights, N. Y.*

"The Use of Glass in Lasers of Very High Output Energy," *Dr. E. Snitzer, American Optical Company, Southbridge, Mass.*

"The Communications Potential of the Optical Maser," *Dr. J. P. Gordon, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"New Developments in Gaseous Optical Masers," *Dr. W. R. Bennett, Jr., Physics Dept., Yale University, New Haven, Conn.*

"Solid-State Optical Maser Amplifiers of High Gain," *Dr. J. E. Geusic, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

Session 2.2—Air Traffic Control

Moderator: *R. P. Pringle, Data Processing and Display Branch Sstems, Research and Development Service, FAA, Washington, D. C.*

"A Digital System for Air Navigation Simulation," *S. K. Chao, J. E. Kearns, F. W. Wright, Sylvania Electronic Systems, Sylvania Electric Products, Inc., Needham, Mass.*

"A Computer Generated Traffic Sample," *S. P. E. Price, Experimentation Div., FAA, National Aviation Facilities Experimental Center, Pomona, N. J., and F. Cristofano, Control Technology, Inc., ATC Systems Branch, NAFEC, Pomona, N. J.*

"Digital Read-In Equipment for Air Traffic Control Research," *M. J. Criswell and J. L. Redifer, Aircraft Armaments, Inc., Cockeysville, Md.*

"Time Compression as an Aid to Air Traffic Control," *B. E. Potter, Surface Armament Div., Sperry Gyroscope Company, Sperry Rand Corporation, Great Neck, N. Y.*

"A New Solid-State Air Traffic Control Transponder," by *E. O. Kirner, Bendix Radio Div., The Bendix Corporation, Towson, Md.*

Tuesday, October 23

Session 3.1—Microminiaturization II—A Look at a New Technology

Panel Discussion

Moderator: *Dr. W. M. Spurgeon, Research Laboratories Div., The Bendix Corporation, Southfield, Mich.*

Members of Panel: N. F. Barnes, Light Military Electronics Dept., General Electric Company Utica, N. Y., Dr. P. K. Weimer, Radio Corporation of America, Princeton, N. J., R. H. Norman, Fairchild Semiconductor Corp., Palo Alto, Calif., Dr. G. Landsman, Associate Director of Research, Motorola Inc., Scottsdale, Ariz., Dr. A. E. Lessor, IBM Lab., Kingston, N. Y.

Session 3.2—Systems

Moderator: *Dr. H. Schutz, Manager, Control and Computing Systems Development, Westinghouse Electric Corp., Air Arm Div., Baltimore, Md.*

"A Simulator Investigation of a Self-Adaptive Pitch Damper, Utilizing a Digital Computer in the Adaptation Loop, for a High Performance Fighter Aircraft," *D. T. Boslaugh, Professor of Electrical Engineering, and G. J. Thaler, Professor of Electrical Engineering, U. S. Naval Postgraduate School, Monterey, Calif.*

"Another Version of a Channel Combiner," *M. G. Kaufman, U. S. Naval Research Labs., Washington, D. C.*

"A Passive Automatic Direction Finder," *J. A. Kaiser, NASA, Goddard Space Flight Center, Greenbelt, Md.; H. B. Smith, Jr., U. S. Naval Torpedo Station, Keyport, Wash.; W. H. Pepper and J. H. Little, Diamond Ordnance Fuze Labs., Washington, D. C.*

"Design Considerations for Space to Earth Data Link," *R. Hauptman, Research Engineer, Armour Research Foundation, Illinois Institute of Technology, U. S. Naval Engineering Experiment Station, Annapolis, Md.*

"A Study of Communications for Manned Space Flight," *A. M. McCalmont, Scientific Analysis Corporation, Concord, Mass.*

"Data Compaction with High-Speed Circuitry," *G. Glen and J. W. Land, Aero Geo Astro Corporation, Alexandria, Va.*

Tuesday Afternoon

Session 4.1—Satellites

Moderator: *Dr. A. J. Kelley, Director, Electronics and Control, NASA, Washington, D. C.*

"Satellite Attitude Determination" *A. R. Mowlem, Space Systems Ctr., Federal Systems Div., IBM Corporation, Rockville, Md.*

"Optimal Attitude Control of Space Vehicles," *W. H. Foy, Jr., and E. J. Lefferts, Electronic Systems and Products Div., The Martin Company, Baltimore, Md.*

"Two-Thrust Rendezvous," *D. H. Winfield, Defense Electronic Products, Radio Corporation of America, Burlington, Mass.*

"A Commercial Doppler Satellite Navigation System," *R. Lieber, Missile and Surface Radar Div., Radio Corporation of America, Moorestown, N. J.*

"Satellite Assisted Tracking Network," *D. B. Davis, Astro Technology Corporation, Palo Alto, Calif.*

"Radar Applications to Geodesy," *Dr. S. Shucker and A. Reich, Missile and Surface Radar Div., Radio Corporation of America, Moorestown, N. J.*

Session 4.2—Circuits

Moderator: *Dr. W. Gore, Dept. of Electrical Engineering, The Johns Hopkins University, Baltimore, Md.*

"Analysis of Automatic Frequency (AFC) and Phase (APC) Control," *E. M. Robinson, Defense Systems Dept., General Electric Company, Syracuse, N. Y.*

"A New AFC System for Rapidly Tuned Radar Pulses," *J. E. Devlin and D. W. C.*

Shen, The Moore School of Electrical Engineering, University of Pennsylvania, Philadelphia, Pa.

"Application of the Delay Lock Discriminator to the Satellite Rendezvous Problem," *W. G. Weis and M. Evans, Research Lab., Lockheed Missiles and Space Company, Sunnyvale, Calif.*

"A Phase-Locked Detector for FM Multiplex Applications," *D. W. Ford, Engineering Dept., Philco Corporation, Philadelphia Pa.*

"Principles of Self-Test Circuits for Airborne Radio Navigation Equipment," *J. M. Tewksbury, Bendix Radio Div., The Bendix Corporation, Towson, Md.*

"Quadded NOR Logic," *P. A. Jensen, Advanced Development, Westinghouse Electric Corporation, Baltimore, Md.*

Wednesday, October 24

Session 5.1—Microwave Components

Moderator: *N. Lipetz, Chief, Transmission Lines Section, USASRDL, Fort Monmouth, N. J.*

"The Backward Wave Converter," *R. W. Wilmarth and R. J. Blanchard, Electron Tube Lab., ITT Components Division, Nutley, N. J.*

"Frequency Multiplication with Varactors," *R. J. Bauer, Aircraft Armaments, Inc., Cockeysville, Md.*

"A Dual-Diplexer Microwave Front-End," *W. J. Engle, S. Rossen, and A. W. Smith, ACF Electronics Division, Paramus, N. J.*

"Strip Line Dual Channel Receiver," *R. J. Bauer and J. C. Tankersley, Aircraft Armaments, Inc., Cockeysville, Md.*

"Electromagnetic Radiation in Biological Research," *Dr. V. Tomberg, Biophysical Research Laboratory, New York, N. Y.*

Session 5.2—New Trends in Control Theory

Panel Discussion

Moderator: *G. S. Axelby, Advisory Engineer, Space Guidance and Control, Westinghouse Air Arm Division, Baltimore, Md.*

Discussion by authorities in the new field of adaptive control. Subjects to be discussed are: Stability—Methods of using Lyapunov's method for stability analysis; Optimal Control—Definition of optimal and examples of the application of optimal control theory; Computational and Physical Realization.

Wednesday Afternoon

Session 6.1—Devices

Moderator: *J. Houston, Manager, Electronics Dept., Aircraft Armaments, Inc., Cockeysville, Md.*

"Low Direct Current-to-Frequency Converter," *E. A. Chilton, Project Engineer, and G. V. Zito, Senior Project Engineer, Eclipse-Pioneer Div., The Bendix Corporation, Teterboro, N. J.*

"Ultra Stable Transistorized D-C Preamplifier with Thermo-Electric Temperature Stabilization," *B. Kruger, Electronic Systems and Products Div., The Martin Company, Baltimore, Md.*

"Design Considerations in Reflected Ion Beam Recording," *D. Forrest, S. Hildum, and L. Bittman, Electronic Systems and Products Div., The Martin Company, Baltimore, Md.*

"Applications of Electroluminescence to Ground Support Equipment," *W. Brooks, Electronic Sciences Lab., Microsystems Electronics, Lockheed Missiles and Space Company, Palo Alto, Calif.*

"A Precise, Fast Earth-Rate Direction Detector," *E. Ohlberg, Nortronics, Northrop Corporation, Palos Verdes Estates, Calif.*

Session 6.2—Antennas and Propagation

Moderator: *Dr. T. J. Carroll, Advanced Research Dept., Bendix Radio Div., The Bendix Corporation, Towson, Md.*

"Experimental Studies of the Correlation Bandwidth of the Tropospheric Scatter Medium at 5 Gc," *W. S. Patrick and M. J. Wiggins, Martin Marietta Corporation, Orlando, Fla.*

"Atmospheric Attenuation in the Millimeter and Submillimeter Wavelength Region," *A. P. Sheppard, The Martin Company, Orlando, Fla.*

"Microwave Exhaust Characteristics of Electrical Space Propulsion Engines," *G. Levy, Plasma Propulsion Lab., Republic Aviation Corporation, Farmingdale, N. Y.*

"Adaptive Antenna Systems for Communications," *R. G. Roush, Research Div., Electronic Communications, Inc., Timonium, Md.*

"Wide Angle Circularly Polarized Antenna Techniques for Spacecraft," *C. E. Ermatinger, V. L. Harrington, W. G. Scott, and C. W. Westerman, Aeronutronic Div., Ford Motor Company, Newport Beach, Calif.*

"A Retro-Directive Antenna for Satellite Data Transmission," *C. Rothenberg, C. Belfi, L. Schwartzman, Surface Armament Div., Sperry, Gyroscope Company, Sperry Rand Corp., Great Neck, N. Y.*

Spaceborne Computer Engineering Conference

DISNEYLAND HOTEL, ANAHEIM, CALIF., OCTOBER 30–31, 1962

The Spaceborne Computer Engineering Conference, sponsored by the IRE Professional Group on Electronic Computers, will be held at the Disneyland Hotel, Anaheim, Calif., on October 30–31, 1962. The program of the Conference is as follows:

Tuesday, October 30

Session I

Introductory Remarks: *W. F. Gunning, Conference Chairman.*

"Design Problems for Space Environment," *J. F. Shea, Manned Space Flight Office, NASA.*

"Space Science Data Devices," *W. E. Brown, Jr., Jet Propulsion Laboratory, California Institute of Technology.*

Tuesday Afternoon

Session II

"Design of the Orbiting Geophysical Observatory Data Handling System," *L. B. Kleiger, Space Technology Laboratories.*

"Two Approaches to the Design of Spacecraft Data Handling Systems," *R. C. Baron, R. W. Waller, Computer Control Company.*

"The Nimbus System," *D. W. Gade, California Computer Products.*

"Starfield Recognition for Space Vechicle Orientation," *S. S. Viglione, H. F. Wolf, Astropower, Inc.*

Wednesday, October 31

Session III

"The ADD-1000 Aerospace Computer," *D. C. Morse, B. J. Jansen, R. P. Blixt, Remington Rand.*

"MAGIC, An Advanced Computer for Spaceborne Guidance Systems," *E. L. Hughes, F. Gurzi, A. H. Faulkner, AC Spark Plug Division, General Motors Corporation.*

"Subminiature Computer Design for Space Environments," *W. A. England, Minneapolis Honeywell Regulator Company.*

"A Micro Computer for Space Navigation," *E. Keonjian, J. Marx, Arma Division, American Bosch Arma. Corporation.*

Wednesday Afternoon

Session IV

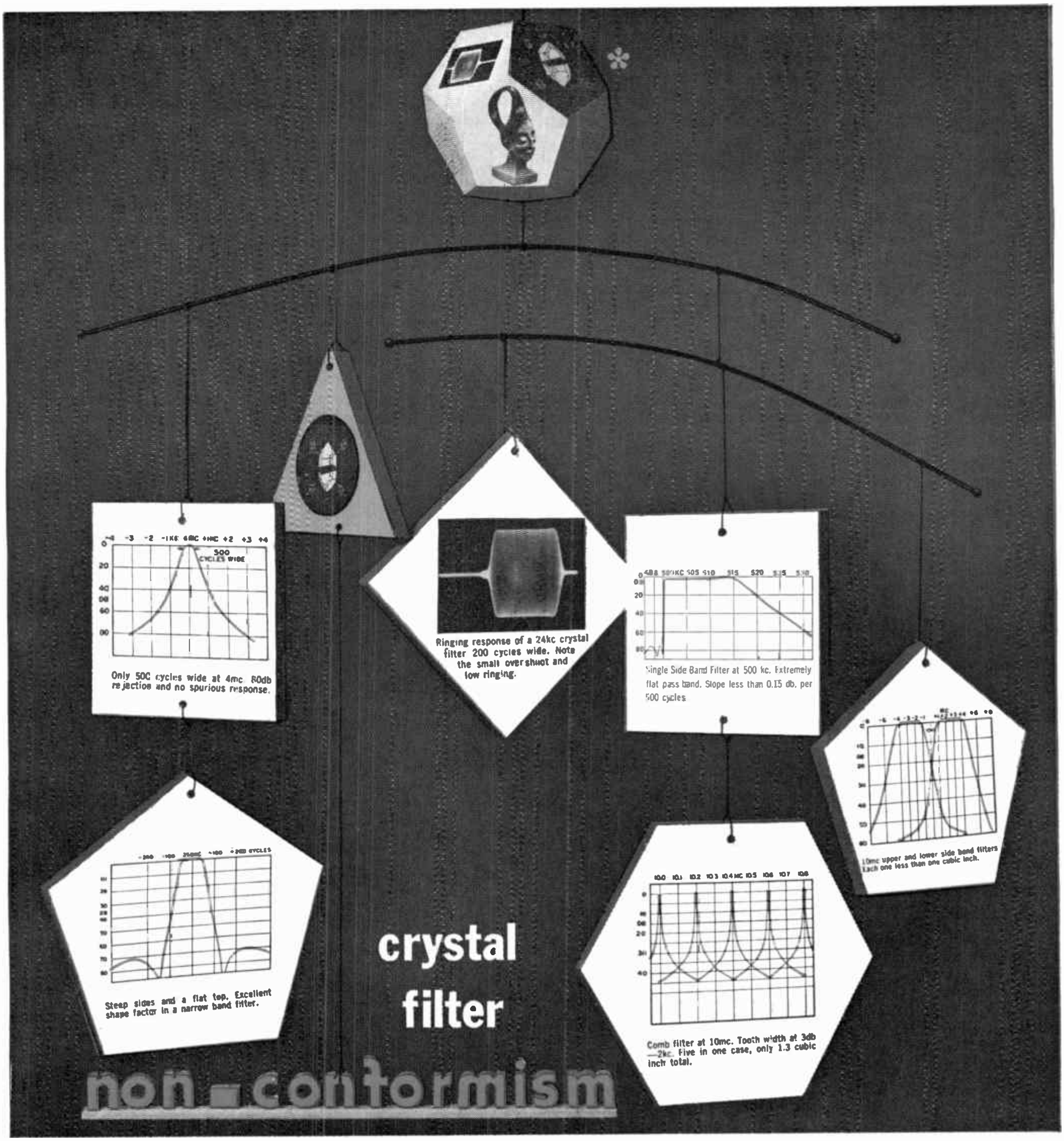
"The D-210 Magnetic Computer," *E.T. Walendziewicz, Burroughs Laboratories.*

"HCM-202 Thin Film Computer," *M. M. Dalton, Hughes Aircraft Company.*

"On the Analysis of Reliability Improvement Through Redundancy," *H. L. Ergott, D. P. Rozenberg, Space Guidance Center, IBM Corporation.*

"Microelectron, Micropower Digital Circuits and Low-Level Amplifiers for Space Applications," *W. W. Gaertner, C. Heizman, C. Levy, M. Schuller, CBS Laboratories.*

Keynote Speaker—*To be announced.*



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Professional Groups*

Aerospace and Navigational Electronics (G-11)—S. H. Dodington, ITT Labs., 500 Washington Ave., Nutley 10, N. J.; H. R. Mimno, Cruft Lab., Harvard Univ., Cambridge 38, Mass.

Antennas and Propagation (G-3)—S. A. Bowhill, 222 Electrical Engineering Bldg., University Park, Pa., Chairman and Editor.

Audio (G-1)—R. W. Benson, Elec. Engrg. Dept., Vanderbilt University, Nashville 5, Tenn.; M. Camras, Armour Res. Foundation, Tech. Ctr., Chicago 16, Ill.

Automatic Control (G-23)—L. B. Wadel, 3546 Caruth Blvd., Dallas 25, Tex.; G. S. Axelby, Westinghouse Air Arm Div., Friendship Airport, Baltimore 3, Md.

Bio-Medical Electronics (G-18)—O. H. Schmitt, Dept. of Physics, Univ. of Minnesota, Minneapolis, Minn.; E. F. MacNichol, Jr., Dept. of Biophysics, Johns Hopkins Univ., Baltimore 18, Md.

Broadcast and Television Receivers (G-8)—J. F. Bell, Zenith Radio Corp., 6001 W. Dickins Ave., Chicago 39, Ill.; C. W. Sall, RCA, Princeton, N. J.

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

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

Communications Systems (G-19)—R. L. Marks, Rome Air Dev. Ctr., Griffiss AFB, N. Y.; E. J. Baghdady, Elec. Engrg. Dept., M.I.T., Cambridge 39, Mass.

Components Parts (G-21)—L. Kahn, Astron Corp., 255 Grant Ave., E. Newark, N. J.; G. Shapiro, Engineering Electronics Sec., Div. 1.6, NBS, Connecticut Ave., & Van Ness St., Washington 25, D. C.

* Names listed are Group Chairmen and Transactions Editors.

Education (G-25)—G. E. Moore, Westinghouse Elec. Corp., East Pittsburgh, Pa., S. S. Shamis, Dept. of E.E., New York University, University Heights, New York 53, N. Y.

Electron Devices (G-15)—R. W. Sears, Bell Telephone Labs., Rm 2A-229, Murray Hill, N. J.; G. Wade, Spencer Lab., Raytheon Co., Burlington, Mass.

Electronic Computers (G-16)—W. L. Anderson, 1408 Lexington Rd., Falls Church, Va.; Prof. N. R. Scott, Dept. of Elec. Engrg., University of Michigan, Ann Arbor, Mich.

Engineering Management (G-14)—E. S. White, Warwick Mfg., Corp., 7300 N. Lehigh, Chicago 48, Ill.; A. H. Rubenstein, Dept. of Indus. Engrg., Northwestern Univ., Evanston, Ill.

Engineering Writing and Speech (G-26)—J. M. Kinn, Jr., IBM Research Ctr., P.O. Box 218, Yorktown Heights, N. Y.; F. T. Van Veen, General Radio Co., West Concord, Mass.

Geoscience Electronics (G-29)—R. W. Olson (Acting Chairman) Texas Instruments, Inc., Dallas, Tex.; Editor to be advised.

Human Factors in Electronics (G-28)—A. S. Jackson, 111 S. Pittsburgh Ave., Ventnor, N. J.; J. I. Elkind, Bolt Beranek and Newman, Inc., 50 Moulton St., Cambridge, Mass.

Industrial Electronics (G-13)—J. E. Eiselein, RCA Victor Div., Camden, N. J.; R. W. Bull, Coleman Instruments, Inc., 42 Madison St., Maywood Ill.

Information Theory (G-12)—D. Van Meter, 1716 Cambridge St., Cambridge 38, Mass.; A. Kohlenberg, Melpar Inc., 11 Galen St., Watertown, Mass.

Instrumentation (G-9)—L. D. McClellan, 420 Via Mesa Grande, Redondo, Calif.; G. B. Hoadley, Dept. of E.E., North Carolina State College, Raleigh, N. C.

Microwave Theory and Techniques (G-17)—S. B. Cohn, Rantec Corp., 2399 Ventura Blvd., Calabasas, Calif.; D. D. King, Electronic Commun., Inc., 1830 York Rd., Timonium, Md.

Military Electronics (G-24)—E. N. Dingley, Jr., Electronic Commun. Inc., 1501-72 St., N., St. Petersburg 10, Fla.; D. R. Rhodes, Radiation Inc., Box 37, Melbourne, Fla.

Nuclear Science (G-5)—Dr. W. A. Higinbotham, Brookhaven Nat'l. Lab., Upton, N. Y.; R. F. Shea, Dig Power Plant Engrg., Knolls Atomic Power Lab., General Electric Co., Schenectady, N. Y.

Product Engineering and Production (G-22)—J. W. Trinkaus, Sperry Gyroscope Co., Mail Station 1A36, Great Neck, N. Y.; D. B. Ehrenpreis, 325 Spring St., New York, N. Y.

Radio Frequency Interference (G-27)—H. Garlan, Office of Chief Engineer, F.C.C., Washington, D. C.; O. P. Schreiber, Technical Wire Prods., Inc., 129 Dermondy St., Cranford, N. J.

Reliability and Quality Control (G-7)—L. J. Paddison, Sandia Corp., Sandia Base, Albuquerque, N. Mex.; W. X. Lamb, Jr. 22124 Dumetz Rd., Woodland Hills, Calif.

Space Electronics and Telemetry (G-10)—H. B. Riblet, 3300 Plyrrs Mill Rd., Kensington, Md.; F. T. Sinnott, Mail Zone 549-30, Convair-Astronautics, San Diego 12, Calif.

Ultrasonics Engineering (G-20)—A. H. Meitzler, 12 Harding Rd., Morristown, N. J.; O. E. Mattiat Res. Div. Acoustica Associates Inc., 415 E. Montecito St., Santa Barbara, Calif.

Vehicular Communications (G-6)—S. C. Bartlett, American Elec. Power Service Corp., 2 Broadway, New York 8, N. Y.; W. G. Chaney, AT&T Co., 195 Broadway, New York 7, N. Y.

Sections*

Akron (4)—R. C. Weyrick, 629 Yerrick Rd. Akron 12, Ohio; J. R. Shoemaker, 1353 Shanabrook Dr., Akron 13, Ohio.

Alamogordo-Holloman (7)—H. T. Castillo, 2611 Pontiac Ave., Alamogordo, N. Mex.; J. T. Bradford, 2704 Highland Dr., Alamogordo, N. Mex.

Albuquerque-Los Alamos (7)—A. B. Church, 2900 San Joaquin, S.E., Albuquerque, N. Mex.; T. L. Pace, 1101 Clancy Dr., N.E., Albuquerque, N. Mex.

Anchorage (7)—H. D. DeVoe, Box 644, Anchorage, Alaska; (To be advised.)

Atlanta (3)—W. B. Jones, Jr., School of Elec. Engrg., Georgia Inst. of Technology, Atlanta 13, Ga.; R. S. Johnson, 1325 Scott Blvd., Decatur, Ga.

Baltimore (3)—W. C. Gore, 1400 Glendale Rd., Baltimore 12, Md.; D. D. King, Elec. Communications, Inc., 1830 York Rd., Timonium, Md.

Bay of Quinte (8)—A. S. Robb, 246 Dufferin Ave., Belleville, Ont., Canada; E. E. Pascal, 558 Victoria Ave., Belleville, Ont., Canada.

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

Beaumont-Port Arthur (6)—C. H. Burgess, Sun Oil Co., Box 2831, Beaumont, Tex.; A. B. Long, Box 3265, Beaumont, Tex.

Benelux—C. B. Broersma, 149 De Lairessestraat, Amsterdam 2, Netherlands; G. J. Siezen, 3 Konig Willem III Laan, Blaricum, Holland.

Binghamton (1)—D. R. Weisser, Meeker Rd., R.D. 3, Vestal, N. Y.; E. L. Blossom, 1304 Seymour Rd., Vestal, N. Y.

Boston (1)—C. E. Faflick, 28 Moon Hill Rd., Lexington 73, Mass.; R. L. Blanchard, 1310 Massachusetts Ave., Lexington 73, Mass.

Buenos Aires—A. DiMarco, Avda. Rivadia 5639-48, Buenos Aires, Argentina; I. C. Grant, Paseo Colon 995, Buenos Aires, Argentina.

Buffalo-Niagara (1)—E. W. Whyman, 375 Mt. Vernon Rd., Snyder 26, N. Y.; R. W. Ames, 132 Evergreen Dr., Tonawanda, N. Y.

Canaveral (3)—W. J. Haberhern, Jr., 123 Atlantic Blvd., Eau Gallie, Fla.; P. H. Betts, Box 609, Cocoa, Fla.

Cedar Rapids (5)—S. M. Morrison, 2034 Fourth Ave., S.E., Cedar Rapids, Iowa; A. E. Anderson, 1054 34 St., N.E., Cedar Rapids, Iowa.

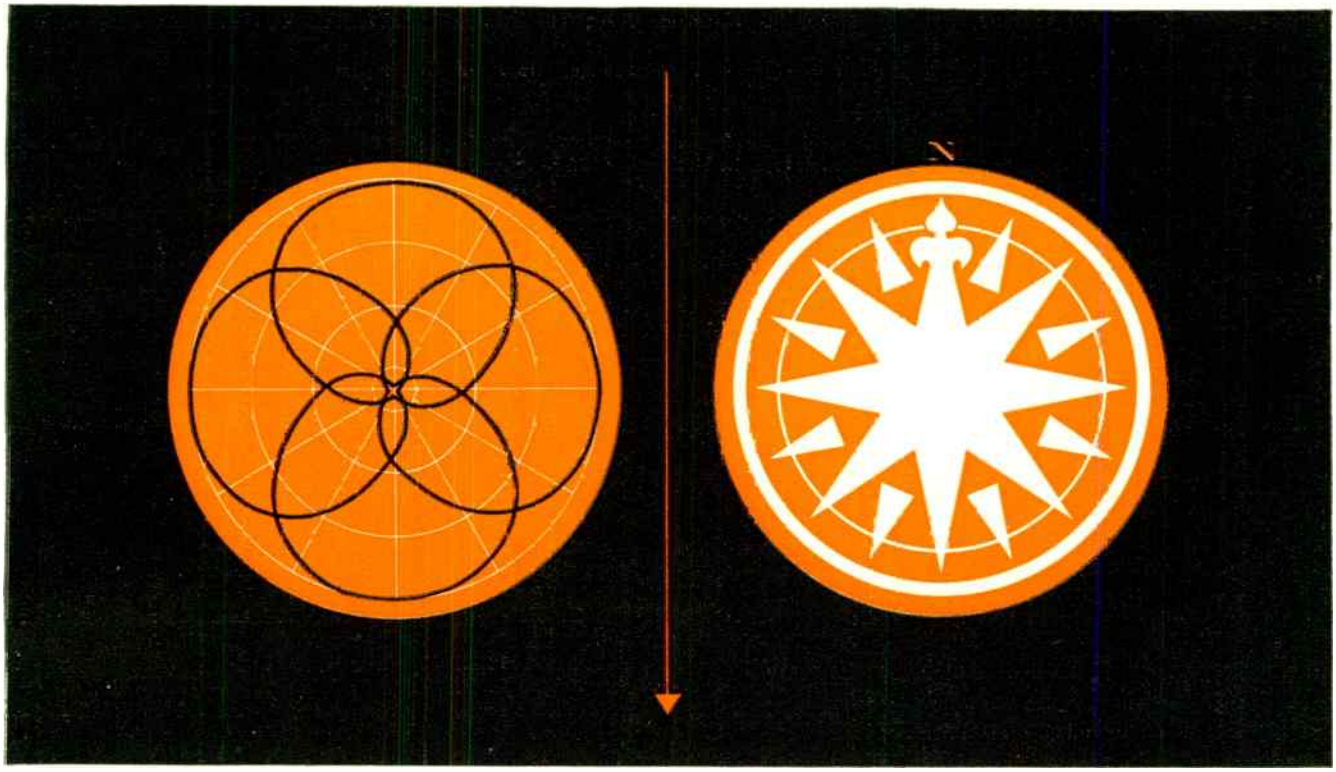
Central Pennsylvania (4)—E. J. Oelbermann, Science Park, Box 60, HRB, Inc., State College, Pa; P. Ebaugh, 222 Hammond Bldg., Penn. State Univ., University Park, Pa.

Chicago (5)—B. G. Griffith, 2379 Oak Tree Lane, Park Ridge, Ill.; D. L. Anderson, Allen-Bradley Co., 838 S. Desplaines, Chicago 7, Ill.

Chile—J. del Rio, Corp. de Radio de Chile, Casilla 1407, Santiago, Chile; J. de la Quintana, IBM World Trade Corp., Cassilla 3630, Santiago, Chile.

China Lake (7)—R. G. S. Sewell, 75-B Rowe, China Lake, Calif.; A. S. Brookman, 205-A N. Dibb Rd., China Lake, Calif.

Cincinnati (4)—A. J. Bissonette, 119 Robinwood Dr., Terrace Park, Ohio; J. W.



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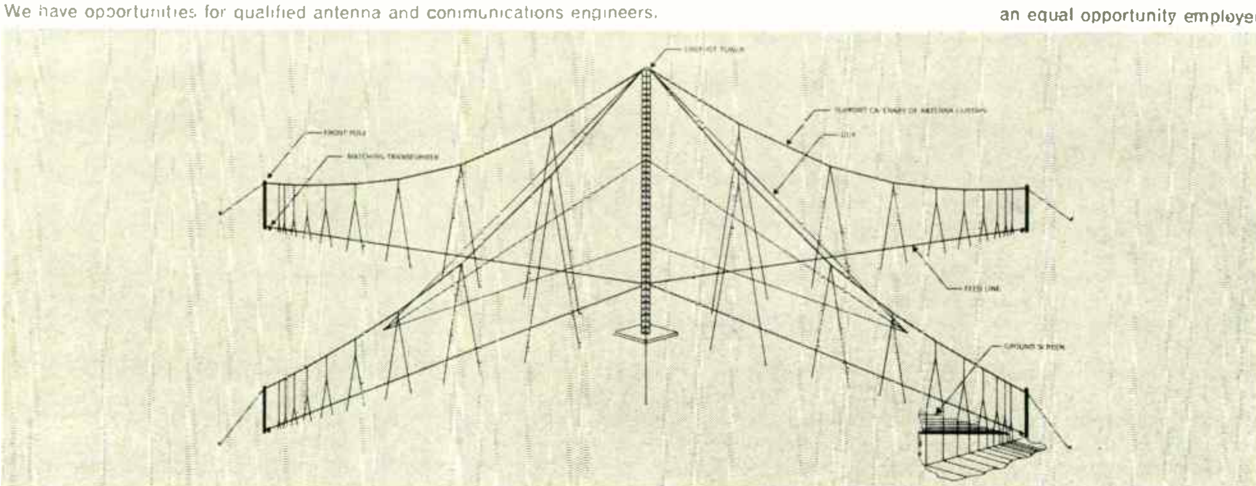

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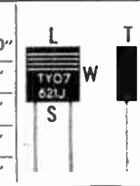
TY06	1	560	.300″	.200″	.115″	.200″
TY07	560	1000	.300″	.300″	.115″	.200″
TY08	1001	2700	.500″	.300″	.115″	.400″
TY09	2701	10000	.900″	.700″	.195″	.800″



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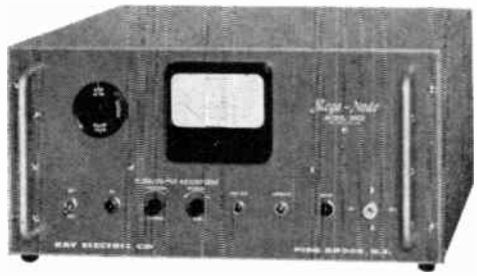

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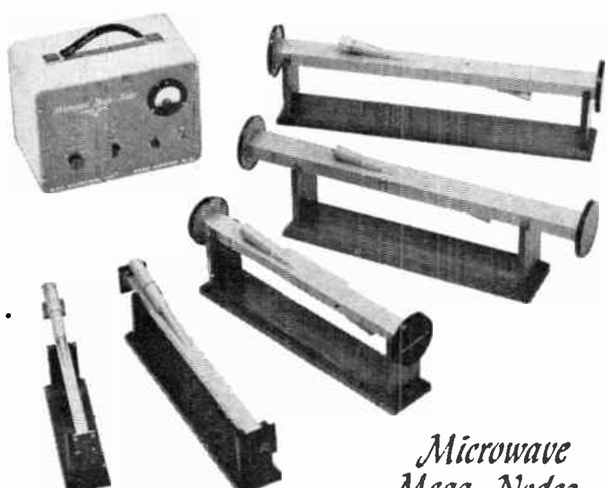

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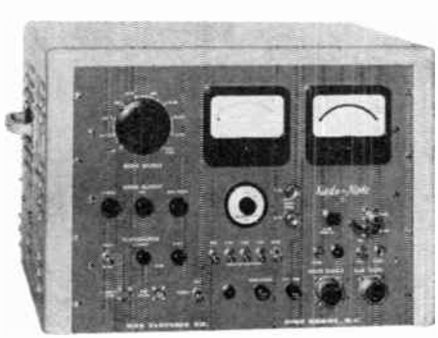

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Catalog No. 600-A

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- **COMPLETE FREQUENCY COVERAGE**

The Rada-Node complements the Kay line of noise sources in providing a complete, precision, radar noise figure measuring set, designed to cover the entire range from 5 mc to 400 mc. With optional higher-frequency noise sources available separately when required, this range may be extended to 26,500 mc. It can also be adapted for TV and other applications.

Features: The Rada-Node consists of two broad band IF amplifiers, an IF detector probe, a single noise diode, an electronically-regulated IF power supply, and noise diode and gas tube power supplies. With optional equipment, it may be used as an IF noise source operating in the 5-3000 mc range or as an RF noise source in the 1120-26,500 mc range. The IF detector probe may be adapted to fit various size tube sockets.

The IF input is fed through switched attenuators, totaling 21 db, and a single, fixed 3-db (2x power) step to the IF amplifiers. These amplifiers, centered at 30 mc and 60 mc, are 14 mc wide and have a gain of 70 db. Each contains a 2-db "interpolation" gain control, as well as "course" and "fine" gain controls common to both amplifiers. Each amplifier consists of six successive stages of IF amplification and a detector. The detector output is monitored by a calibrated db meter and is fed to the video output.

Price: $1595.00 f.o.b. factory ($1755.00 F.A.S., N. Y.)

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Industrial Engineering Notes*

Association Activities

The 39th Annual EIA Convention will be held June 18–20, 1963, at the Pick-Congress Hotel in Chicago, Executive Vice President James D. Secrest has announced. A majority of the Committee on Arrangements, appointed by Past President L. Berkley Davis, favored this date after the Board of Directors had indicated it no longer favored holding the convention during the week of the Parts Show. Mr. Secrest pointed out that the Association will be returning to a schedule followed for a number of years before tying in with the Parts Show. The later date also will permit a wider separation between the Spring Conference and the Annual Convention. . . . **EIA's annual statistical and textual review of the state of the electronics industries, previously called the EIA Fact Book, will be published in early August bearing a new name, a new format, and containing a vast amount of new information never before consolidated into a single volume.** The EIA Marketing Services Department, which prepares the traditional best-seller, changed the name of the publication this year to EIA Year Book. Unpublished last year because of a reorganization within the department, it will be continued as an annual publication. Department Director Frank W. Mansfield (Sylvania Electric Products, Inc.) said the 85-page Year Book will review the shape of the industry in 1961 in greater detail than the old Fact Book. A dozen charts and 70 tables support the text. A limited number of copies will be sent to EIA member-companies. Additional copies may be obtained by check or money order from the Office of Information, EIA Headquarters, at $1 a copy for member-company representatives. The price is $2 for non-members. . . . **The EIA Marketing Services Department announced it is soliciting participation of microwave component manufacturers in its forthcoming quarterly statistical program.** Factory dollar sales data will be compiled on 16 line items including isolators, circulators, attenuators, switches, harmonic generators, mixers, and similar microwave components by material types such as ferrite and semiconductor. Research and development figures will also be collected. All U. S. manufacturers of microwave components are eligible to participate by reporting their dollar sales data each quarter to EIA for which they will recieve the consolidated industry figures in return. Data will be compiled beginning with the first quarter 1961. Manufacturers who wish to participate are asked to contact the Marketing Services Department, EIA Headquarters. The reporting program is sponsored by the Microwave Components Section, EIA Industrial Electronics Division, chaired by Dr. Howard Sharfman (Raytheon Co.).

Governmental and Legislative

Publication of a new magazine for businessmen was announced by Secretary of Commerce Luther H. Hodges. The new weekly, entitled "International Commerce," succeeds the old "Foreign Commerce Weekly," published under various names and titles since 1880. The first issue of "International Commerce" describes more than 300 specific opportunities for U. S. trade and investment abroad. It also lists upcoming construction projects around the world, contains a market survey of prospects in Cameroon, a full-scale review of the business outlook in oil-rich Kuwait, how-to-do-it stories for exporters, in-depth economic reviews, a calendar of events for international traders, and other features, Mr. Hodges said. Subscription for "International Commerce at $16 a year ($5 additional for foreign mailing) may be placed with the Superintendent of documents, U. S. Government Printing Office, Washington 25, D.C., or with Commerce Field Offices. Air mail service is available to domestic subscribers at an additional charge of $21.85 to cover postage. The price of a single copy is 35 cents.

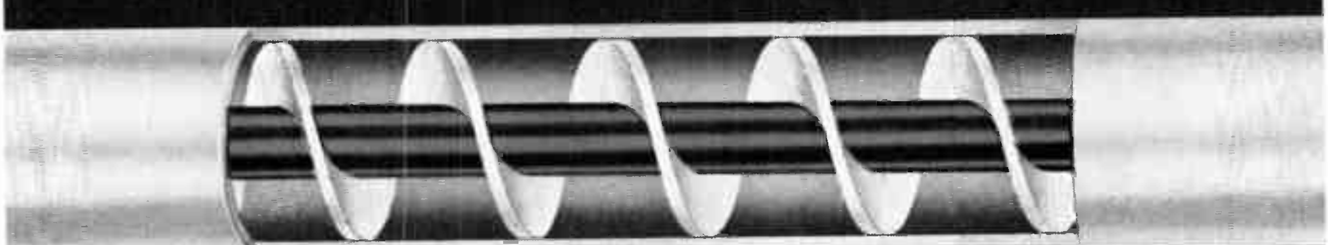
Military and Space

A set of management principles for common use by contractors of the Department of Defense and the National Aeronautics and Space Administration in managing complex development projects has been adopted by the two agencies. These principles are incorporated in a document entitled "DOD and NASA Guide, PERT COST System Design," which is expected to be published July 1. PERT was explained as a management system useful in planning and managing complex development and construction programs and for forecasting the ultimate ability of contractors to meet scheduled target dates. PERT COST is the application of cost to the system which enables both contractor and program manager to forecast possible overrun or underrun of time and cost in sufficient time to take preventive action. The basic PERT (program evaluation and review technique) was first introduced to Government-industry use in 1958. Since then it has been applied largely to evaluation of schedules and time-related problems where it has met with "such wide success" that it is being extended to prediction and control of cost, the agencies said. Since NASA and Defense use virtually the same industrial base, the several approaches to the development of PERT systems could have required a con-

(Continued on page 40A)

* The data on which these Notes are based were selected by permission from *Weekly Report* issues of June 11 and 25 and July 2 and 9, 1962, published by the Electronic Industries Association, whose helpfulness is gratefully acknowledged.

NEW PD HELICAL MEMBRANE AIR DIELECTRIC COAXIAL CABLE



Fast propagation, low signal loss and high temperature resistance—all in one efficient, lightweight cable!

PD Helical Membrane cable marks a new step in the state of coaxial cable art for missile, missile launching and atomic energy instrumentation applications.

It combines all the outstanding advantages of PD air dielectric coaxial cables—low attenuation, excellent frequency response, uniform electrical properties over wide temperature variations and unlimited operating life—with even greater speed of propagation and, when used with a Teflon® helix, higher heat resistance.

The inner conductor is coaxially supported by a polyethylene helix within a commercially pure, seamless aluminum outer conductor. For applications involving high temperatures (100° C-250° C), PD Helical Membrane cable with Teflon® substituted for polyethylene is ideal.

PD Helical Membrane cable of 50, 75 and 100 ohm impedance is fabricated in 1000-foot continuous lengths and in standard sizes of ½", ¾" and 1" diameters; other sizes from ⅜" to 1⅝" on order. Complete cable systems, including attachments and connectors, are available. Your Phelps Dodge representative will be glad to give you additional information. PD Helical Membrane cable is made by Phelps Dodge Copper Products Corporation at Yonkers, N. Y.

PHELPS DODGE ELECTRONIC PRODUCTS
CORPORATION • 300 PARK AVENUE, NEW YORK 22, N.Y

Industrial Engineering Notes

(Continued from page 38A)

tractor to use different systems for different jobs, the agencies said. Adoption of a single approach by the DOD and NASA is expected to minimize such differences. Contractors may secure copies of the DOD and NASA guide from Government Printing Office, Washington 25, D.C. These are expected to be available after July 1.

Engineering

Organic semiconductors are still in their infancy. How will they affect our technology when they grow up? That question is explored in a report published by the Commerce Department's Office of Technical Services. The report is one of a series being specially prepared under the direction of the Office of Technical Services as a new service to science and industry. The report emphasizes that "industrial managers, with the spectacular history of silicon and germanium semiconductors fresh in mind and with thermoelectrics looming as a technological bonanza, are keeping a watchful eye on research related to the electrical properties of organic materials." Research into organic semiconductors and their applications has been spurred by alleged Russian advances in the field, according to the report. The Russians claim to have produced stable, electrically conductive materials by subjecting acrylic resins to ionizing radiation. Both the polyacrylonitrile and silicon-acrylonitrile types of polymers are said to be under test at the Institute of Semiconductors in Leningrad. Although literature dealing with the industrial applications of organic semiconductors is relatively sparse, the report notes that the most significant documents dealing with the subject are those resulting from government-sponsored research projects. The report is "Organic Semiconductors—Their Technological Promise." Clyde Williams and Company, for OTS. 19 pages. (Order PB 181 037 from OTS, U. S. Department of Commerce, Washington 25, D.C., price 50 cents.)

IRE People

Charles F. Horne (A'35–SM'53–F'58) has been elected President of the Electronic Industries Association for 1962–1963. He is President of General Dynamics/Pomona, Pomona, Calif., and of General Dynamics/Electronics, Rochester, N. Y., and a Senior Vice President of General Dynamics.

He is both a veteran naval officer and electronics engineer, who had a distinguished career before retiring from Government service to enter the electronics industry. In 1926, he was graduated from the Naval Academy. He did postgraduate work at the Navy's graduate school in communications and electronics, and in 1935 he received the M.S. degree in communications and electronics from Harvard University. During World War II, he was a communications and radar officer in the Pacific, receiving several combat citations and campaign ribbons. From 1946 to 1948, he was Deputy Chief of Naval Communications. On loan from the Navy to the Civil Aeronautics Administration in 1949, he was acting director of the Federal Airways Division. Upon his retirement from the Navy in 1951, with the rank of Rear Admiral, he was appointed Civil Aeronautics Administrator, serving in that capacity until 1953. In that same year, he joined General Dynamics' former Convair Division as manager of its Pomona Operating Division.

During 1960–1961, Mr. Horne was a member of the IRE Board of Directors. He has been a member of the EIA Military Products Division, Policy Committee, and Chairman of its Military Systems and Management Relations Committee. He is also West Coast advisor to the Radio Technical Commission for Aeronautics, member of the Board of Directors of the Armed Forces Communications and Electronics Association. He served as Chairman of the Board of Directors of the Southern California Industry Education Council and is a member of the Advisory Board of the Los Angeles International Science Center and Space Museum, and the U. S. Chamber of Commerce Education Committee.

❖

Robert Munk (M'56) has joined the technical staff of Electro-Optical Systems, Inc., as Chief Scientist of the company's Advanced Electronics and Information Systems Division. In this position he will be responsible for systems research and analysis activities for space vehicles and weapon systems.

Prior to joining Electro-Optical Systems, he was Chief Scientist and Advanced Systems Engineering Manager for Ryan Aeronautical Corp., San Diego, Calif., responsible for new business planning, the preparation of proposals and the management of study programs. Before that he was manager of the Special Equipment Development Department of Litton Systems, Inc. Before joining Litton, he was head of the Pershing missile guidance and control section for The Martin Company.

Mr. Munk received his B.E.E. and M.E.E. degrees from the Polytechnic Institute of Brooklyn. He is a member of the National Society of Professional Engineers.

❖

The appointment of **Dr. John P. Nash** (M'53–SM'54) as a Vice President of the Lockheed Missiles and Space Co. was announced recently. He has been director of the company's research activities since 1959, and in March, 1962 became Director of the Research and Engineering Laboratories. In his capacity as Vice President, he will continue to provide executive direction over the company's Research and Engineering organization.

He has been with Lockheed since 1957. He was first manager of the Information Processing Division, and in 1958 became associate director, Communications and Controls Research. From 1940 to 1946, he was an assistant professor of mathematics at the University of Notre Dame. From 1942 to 1946, he was a staff member of the MIT Radiation Lab., and from 1946 to 1950, a research physicist with Kimberly-Clark Corp. From 1950 to 1957, he was, successively, assistant professor, associate professor and full research professor in applied mathematics at the University of Illinois.

Dr. Nash received the B.A. degree in mathematics from the University of California, Berkeley, in 1936, the M.A. degree from Rice Institute in 1938, and the Ph.D. degree from Rice in 1940.

❖

Appointment of **H. Malcolm Ogle** (S'40–A'41–SM'50) as assistant to the president of Applied Systems, Palo Alto, Calif., has been announced. Before joining Applied Systems, he worked for twenty-one years for General Electric Company, managing such projects as the development of electronic instruments and systems for nuclear

(Continued on page 16A)

New Nanosecond* Pulse Transformers for Ultra-miniature, Ultra-high Speed Applications

Digital circuit designers will find the new Sprague Type 43Z Nanosecond Pulse Transformers of considerable interest. These tiny transformers have been carefully designed for the all-important parameter of minimum rise time at high repetition rates up to 10 mc.

The new Type 43Z series is comprised of a broad line of 72 pulse transformers in 10 popular turns ratios. They are Sprague's latest addition to the most complete listing of pulse transformers offered by any manufacturer for use in digital computers and other low-level electronic circuitry.

Type 43Z Pulse Transformers are designed so that the product of leakage inductance and distributed capacitance is at a minimum. They are particularly well suited for transformer coupling in transistor circuits since transformers and transistors are very compatible low impedance devices. Nanosecond transformers are equally suitable for transmission line mode of operation, in twisted-pair transmission line coupling, and in regenerative circuits.

The epoxy-encapsulated "pancake" package is excellent for both etched wire board or conventional chassis mounting. To simplify etched-board design, these ultra-miniature pulse transformers are available with leads terminating at the side or the bottom of each unit.

For complete technical information on Type 43Z Nanosecond Pulse Transformers, write for Engineering Data Sheet 40235 to Technical Literature Section, Sprague Electric Co., 235 Marshall St., North Adams, Mass.

*millimicrosecond

NEW . . . for "Bread-Boarding" Your Circuit Designs . . .

Contains 12 specially-selected Sprague Type 32Z miniature pulse transformers in clear, hinged-lid plastic case, complete with simple instructions.

SPRAGUE 100Z41 EXPERIMENTAL PULSE TRANSFORMER ASSORTMENT

- **Helps you choose the right pulse transformer for your specific application.**
- **Puts at your disposal 58 turns-ratio/primary-inductance combinations, providing the parameters required in most electron tube or transistorized circuits.**
- **Primary inductances from 160 microhenries to 43 millihenries.**
- **Turns ratios from 1:5 step-up to 6:1 step-down.**
- **Potted, pre-molded case construction facilitates bread-board wiring, permits frequent re-use.**

Once you determine needed transformer characteristics, it's easy to get production quantities to your exact requirements from Sprague's broad line of hermetically sealed or encapsulated pulse transformers.

For fast delivery or additional information on the 100Z41 Pulse Transformer Assortment, see your Sprague Products Co. Industrial Distributor, or write Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

19-432

SPRAGUE®
THE MARK OF RELIABILITY

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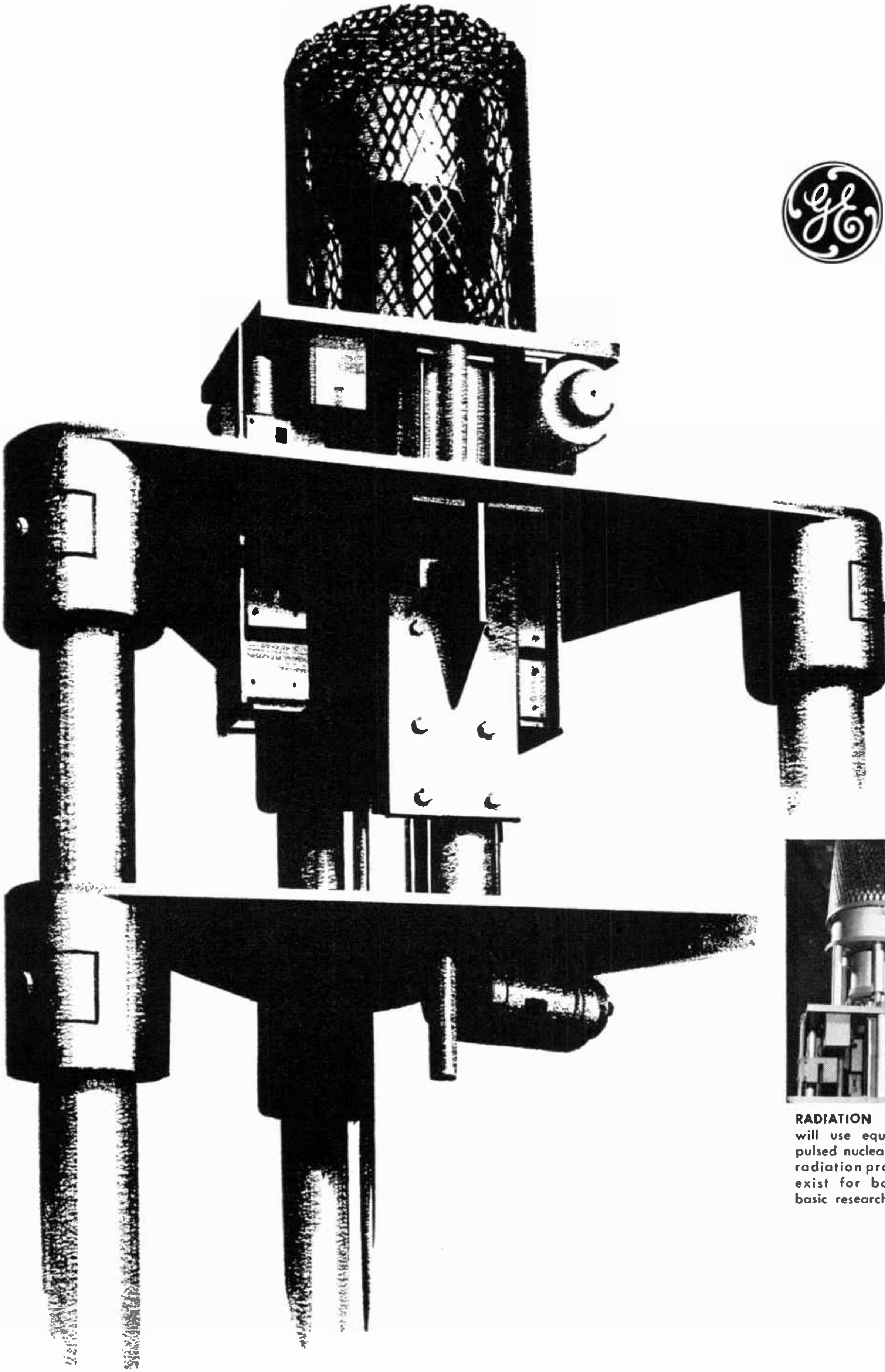


NEWS.

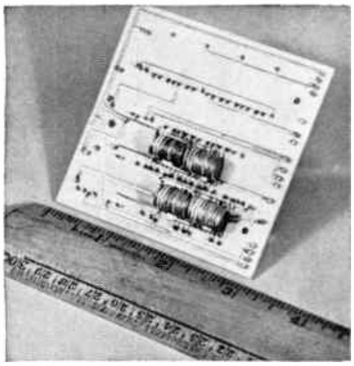
RADIATION EFFECTS OPERATION will use equipment such as this pulsed nuclear test reactor to solve radiation problems. Capabilities exist for both simulation and basic research in radiation effects.

...OF DEFENSE TECHNOLOGIES

NUCLEONICS

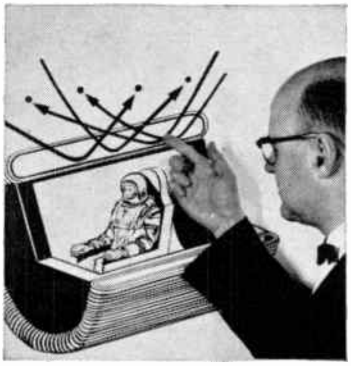
The damaging effects of radiation from nuclear weapons and space have introduced a whole new class of problems in the design of electronic systems and support equipment. For example, gamma radiation ionizes air to provide leakage paths for stray currents. Conventional insulating materials become partial conductors. The performance of transistors and ferrites is altered and voltages are induced in coils, wires, and cables. Van Allen radiation darkens the windows of space vehicles and causes deterioration of semi-conductor materials, such as solar cells.

The creation of systems and equipment to function reliably in these environments requires special test facilities, skills tools, and knowledge that have been developing at General Electric for more than twenty years. A newly organized Radiation Effects Operation is now integrating nuclear and electronic disciplines (nucleonics) and further developing the capabilities of the Company in this new field.



HARDENED ELECTRONICS, such as this Thermionic Integrated Micro Module board, are being developed. TIMM circuits operate above 500°C. and can tolerate 1000 times more radiation than conventional circuits.

NUCLEAR DETECTION SYSTEM (NUDETS-477L) is the first defense system combining nuclear and electronic technologies. It is being developed for U. S. Air Force to locate and measure any nuclear explosion in U. S.



SHIELDING OF SPACE VEHICLES from radiation may be possible by surrounding them with magnetic energy. This new concept, under study for NASA, may alleviate the need for heavier solid shielding.

ELECTRO-HYDRAULIC HANDYMAN is an example of remotely operated servo systems being produced to manipulate radioactive materials for routine tasks as well as highly specialized nuclear research projects.

Progress Is Our Most Important Product

GENERAL ELECTRIC

DEFENSE ELECTRONICS DIVISION

IRE People

(*Continued from page 42A*)

reactors and jet engines. He holds twenty-five U. S. Patents on control systems and devices. Most recently he was manager of General Electric Plasma Diode Project at Vallecitos Atomic Labs., Pleasanton, Calif. At Applied Systems, Ogle will supervise the development of instrumentation in the thermoelectric and thermionic energy fields.

Mr. Ogle holds a B.S in electrical engineering from Johns Hopkins University. In 1946 he received the Charles A. Coffin Award for developing gas turbine controls. He is a member of the American Rocket Society and the American Institute of Electrical Engineers.

❖

William E. Seaman (M'55) has been appointed to the position of Manager of Engineering of Radiation Counter Laboratories, Inc., Skokie, Ill. He will be responsible for all engineering, research and development work within RCL.

Previously, he headed his own engineering consulting firm, William E. Seaman and Associates of Woodside, Calif. He has also held key posts with Ampex as Chief Engineer, Instrumentation Division, and with the Ernest O. Lawrence Radiation Laboratory of the University of California at Berkeley. In the latter position he was responsible for the development of a wide range of nuclear instrumentation.

Mr. Seaman is a graduate of the University of California (1949), a member of ISA, American Management Association, and other professional organizations.

❖

Whitney M. Silhavy (A'40) has been named Technical Assistant to the President of Varian Associates. Succeeding Mr. Silhavy as Manager, Field Engineering, Palo Alto Tube Division is **Clifton Rockwood** (S'48–A'50–M'56), formerly Manager, Applications Engineering.

Manager of Tube Field Engineering since 1954, Mr. Silhavy joined Varian in 1952 as Assistant Director, Applications Engineering. Prior to that he was employed by the Sperry Gyroscope Co. in a number of technical sales and engineering positions. He attended Pacific State University, UCLA, and Oregon State College.

Mr. Rockwood joined Varian in 1951 after holding positions with the Intermountain Broadcasting Co. and as an electronics technician for the U. S. Army during World War II. He received his B.S.E.E. in 1950 from the University of Utah. He is a member of Tau Beta Pi and the Electronic Sales Managers Association.

❖

(*Continued on page 50A*)

ALITE®

CERAMIC-TO-METAL SEALS

Standard Bushings or Special Designs

FROM ONE COMPLETELY INTEGRATED SOURCE

ALITE — with its completely equipped facilities for producing high quality, vacuum-tight, ceramic-to-metal seals — is geared to meet all your requirements for high alumina ceramic-metal components. From design to finished assembly, every manufacturing step — including formulating, firing, metalizing and testing—is carefully supervised in our own plant. Result: effective quality control and utmost reliability.

Hermetic seals and bushings made of high alumina Alite are recommended for electromechanical applications where service conditions are extremely severe or critical. Alite has high mechanical strength and thermal shock resistance. It maintains low-loss characteristics through a wide frequency and temperature range. It resists corrosion, abrasion and nuclear radiation. Its extra-smooth, hard, high-fired glaze assures high surface resistivity.

To simplify design problems and speed delivery, Alite high voltage terminals, feed-throughs and cable end seals are available in over 100 standard sizes. However, when specifications call for special units for unusual applications, you can rely on expert assistance from Alite engineers to help you take full advantage of Alite's superior properties.

Write us about your specific requirements today.

WRITE FOR HELPFUL FREE BULLETINS

Bulletin A-8 gives useful comparative data. Bulletin A-40-R describes Alite facilities and complete line of Alite Standard Bushings.

ALITE DIVISION

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102-G

NEWS New Products

Pulse Generator

Charles E. Rutherford, president of **Rutherford Electronics Co.,** P.O. Box 472, Culver City, Calif., announces a new pulse generator, the Model B7D. This new model features variable rise and fall time functioning completely independent of each other. The rise or fall time may be degraded continuously to approximately one microsecond without affecting the other. Other features of the B7D include simultaneous positive and negative 50 volt output and the ability to continuously vary the dc level of the output pulses from −10 to +10 volts, by front panel control.

Other specifications, such as: repetition rate to 2 mc, delay and width to 10,000 μs, rise and fall time of 15 μs, and 50 volts into 50 ohms to 30% duty factor are the same as the present Rutherford Model B7B.

The new Model B7D is priced at $1,200.00 each, F.O.B. Culver City.

Itek Appoints Kelly

Dr. Donald H. Kelly has been named a Senior Optical Physicist in the **Itek Laboratories** Graphic Information Research Division, Lexington, Mass.

Before joining Itek, Dr. Kelly was a member of the Senior Research Staff at Technicolor Corporation, where he directed the development of a new optical component which made possible color motion-picture photography at low light levels previously suitable only for black-and-white photography. He has also conducted important research work in the areas of photographic image enhancement and spatial filtering.

While a Technicolor employee, Dr. Kelly also completed his graduate studies in vision at UCLA, designed photographic research instruments, and was involved in projects on color film measurements and television bandwidth compression. Previously, as a U. S. Navy officer, he had participated in projects concerned with target visibility, atmospheric optics, and aircraft camouflage.

Dr. Kelly is specially qualified in applying the techniques of systems engineering and information theory to studies on the performance of human vision. At Itek Laboratories, he is organizing a visual research program to evaluate the human eye as a component of information systems.

Dr. Kelly received a B.S. degree in optics from the University of Rochester, and in 1960 was awarded a Ph.D. degree in engineering at the University of California, Los Angeles. His honorary affiliations include Phi Beta Kappa and Sigma Xi.

The author of numerous technical articles on vision and image evaluation, Dr. Kelly holds 12 U. S. patents and is a member of the Optical Society of America, and the Society of Motion Picture and Television Engineers

Liquid Laser Material

Semi-Elements, Inc., Saxonburg Blvd., Saxonburg, Pa., announces the newest and latest exploratory laser material, rare earths doped into liquids. The new liquid lasers present the latest in laser technology in that by utilizing liquid lasers the cooling problem is solved. There are many advantages to liquid lasers. First of all the frequency of the output of the liquid laser can be varied by the introduction of a second impurity. The liquid laser concentrate is put into a chamber similar to the chambers used for gas lasers. If the liquid laser is to be used for a pulsed operation, cooling is unnecessary. But, if continuous operation is desired, then the liquid is circulated through a cooling system into the laser chambers and, again, through the cooling chamber. By maintaining a constant temperature, the index of refraction of the liquid is not changed. This is also an advantage to the system. From the economy standpoint one can change to another frequency output by merely changing to another rare earth liquid. So far, Semi-Elements has achieved success in putting Gadolinium, Neodymium and Samarium into a liquid concentrate, the concentrations being 5% of Gadolinium, 2% Samarium and 2% Neodymium. The emissions of rare earths in liquid concentrate are still being investigated, as well as other types of liquids used in conjunction with rare earths to find the most efficient liquid. Another advantage of the liquid laser is that it may be possible to frequency modulate the liquid by introducing an additional conductive liquid, which will vary the angstrom output by means of varying the voltage potential. Further research work is being carried on currently on this new phenomenon. The Gadolinium, Samarium and Neodymium concentrates are $250.00 per pound.

(Continued on page 150A)

High voltage

High gain

Low leakage

Silicon Planar

DIFFERENTIAL AMPLIFIERS

Matched h_{FE}

Matched V_{BE}

Thermally matched

Fairchild Planar process alone makes this matching practicable in volume production

Matching Characteristics	2N2060 Min.	2N2060 Max.	2N2060 Units	2N2223 Min.	2N2223 Max.	2N2223 Units	2N2223A Min.	2N2223A Max.	2N2223A Units	Test Conditions
Beta Ratio	0.9	1.0								$I_C = 1.0$ mA $V_{CE} = 5.0$ V
Beta Ratio	0.9	1.0		0.8	1.0		0.9	1.0		$I_C = 0.1$ mA $V_{CE} = 5.0$ V
V_{BE} Differential		0.005	Volts							$I_C = 1.0$ mA $V_{CE} = 5.0$ V
V_{BE} Differential		0.005	Volts		0.015	Volts		0.005	Volts	$I_C = 0.1$ mA $V_{CE} = 5.0$ V
ΔV_{BE} Tracking		10	μV/°C		25	μV/°C		25	μV/°C	$I_C = 0.1$ mA $V_{CE} = 5.0$ V

IRE People

(*Continued from page 46A*)

The board of directors of Granger Associates elected **E. W. Pappenfus** (M'45–SM'50–F'62) Vice President, engineering. He will assume wide responsibilities for development of the company's programs in radio communications systems and devices.

He spent 19 years with Collins Radio Co., where he was at the last director of development of the firm's largest design division. Among his past responsibilities was design and technical management of several large communications and electronics programs, including the Air Force "Bird Call" single sideband system, the radio frequency portion of the Naval tactical data system, Signal Corps vehicular and fixed radio devices and Dew Line scatter communications.

Mr. Pappenfus graduated with honors in electrical engineering from the University of Minnesota. He is co-author of a book in the single sideband field to be published later this year by McGraw-Hill Book Co. He holds three issued U. S. patents and is an enthusiastic radio amateur, holding the call letters W0SYF.

❖

Delmer C. Ports (A'38–S'45–F'61), Vice President of Jansky and Bailey, has been selected as the recipient of the George Washington University Distinguished Engineer Alumni Award for 1962. The award is granted to an alumnus of the University in recognition of notable contributions in the engineering profession. As recipient of this award, he was invited as Guest Lecturer for the Engineer Alumni Association's Frank A. Howard Lecture on May 8, 1962, in the Lisner Auditorium of the University.

Mr. Ports holds a B.S. degree in electrical engineering from George Washington University, and a M.S. degree from Ohio State University. He has been associated with Jansky and Bailey since 1936.

❖

Edgar A. Post (A'37–SM'47–F'61) manager of the radio and weather sciences laboratory at Stanford Research Institute, has been appointed to the Institute's European office in Zurich, Switzerland. His appointment reflects the Institute's increasing engineering activity in Europe and the need for additional staff there.

A specialist in systems engineering, he has a long background in the aircraft communications and navigation field. He was with United Air Lines before joining the Institute in 1956.

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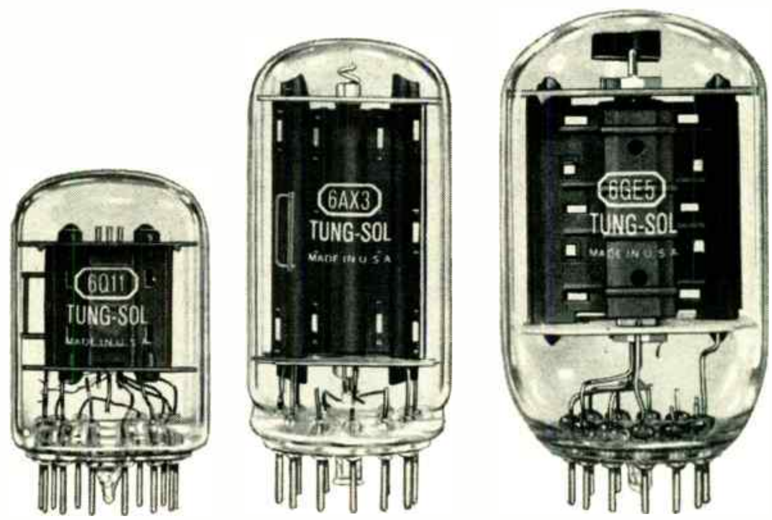
(*Continued on page 52A*)

Higher Performance Standards With Improved Reliability... Tung-Sol compactrons provide several advantages that can contribute to lower costs and improved performance. For example, the increased number of pins permit greater heat dissipation. As a result, compactrons run cooler with higher reliability than conventional tubes. The exhaust tubulation is situated between the pins so that broken tips rarely occur. This also permits the use of top caps for very high voltage designs. In addition, the compactron design readily lends itself to combining multiple tube elements within a single envelope.

Compactrons require less space on the chassis or printed circuit boards, less height than conventional tubes, less air cooling volume per function. More space between pins improves element isolation, allows higher voltage ratings, simplifies printed circuit and chassis design.

Tung-Sol compactrons are available in production volume for numerous circuit requirements, including radio, tv, hi-fi and stereo, controls and instrumentation equipment. Write for Tung-Sol compactron data file which includes the following types: 6AX3, 6GE5, 6Q11, 12AX3, 12GE5, 8149, 8150 and 1AJ2. Other types will soon be available and special designs will be considered. Tung-Sol Electric Inc., Newark 4, N. J. TWX: NK193.



TUNG-SOL COMPACTRONS

TUNG-SOL®

IRE People

(Continued from page 50A)

Charles K. Raynsford (A'53–M'59) has been appointed engineering and research manager for the Paramus plant of ACF Electronics, a division of ACF Industries, Inc., it was announced recently. He comes to ACF from Vitro Laboratories in West Orange, N. J. where he was chief engineer. He will be responsible for technical and administrative direction at the plant of five engineering laboratories, product design and technical services. Mr. Raynsford is a graduate of the Massachusetts Institute of Technology.

❖

Mischa Schwartz (S'46–A'49–M'54–SM'54) Acting Head of the Electrical Engineering Department of the Polytechnic Institute of Brooklyn, was elected to the Board of Directors of Burmac Electronics Co., Inc. of Farmingdale, N. Y.

He joined the teaching staff of Brooklyn Polytechnic in 1952, soon became Professor of Electrical Engineering and has been Acting Head of the department since January of this year. From 1947 to 1952 he was a project engineer with the Sperry Gyroscope Co. assigned to the Basic Systems Study Group of the Radar Engineering Department. His work included microwave systems analysis and evaluation, systems engineering and basic studies in noise theory.

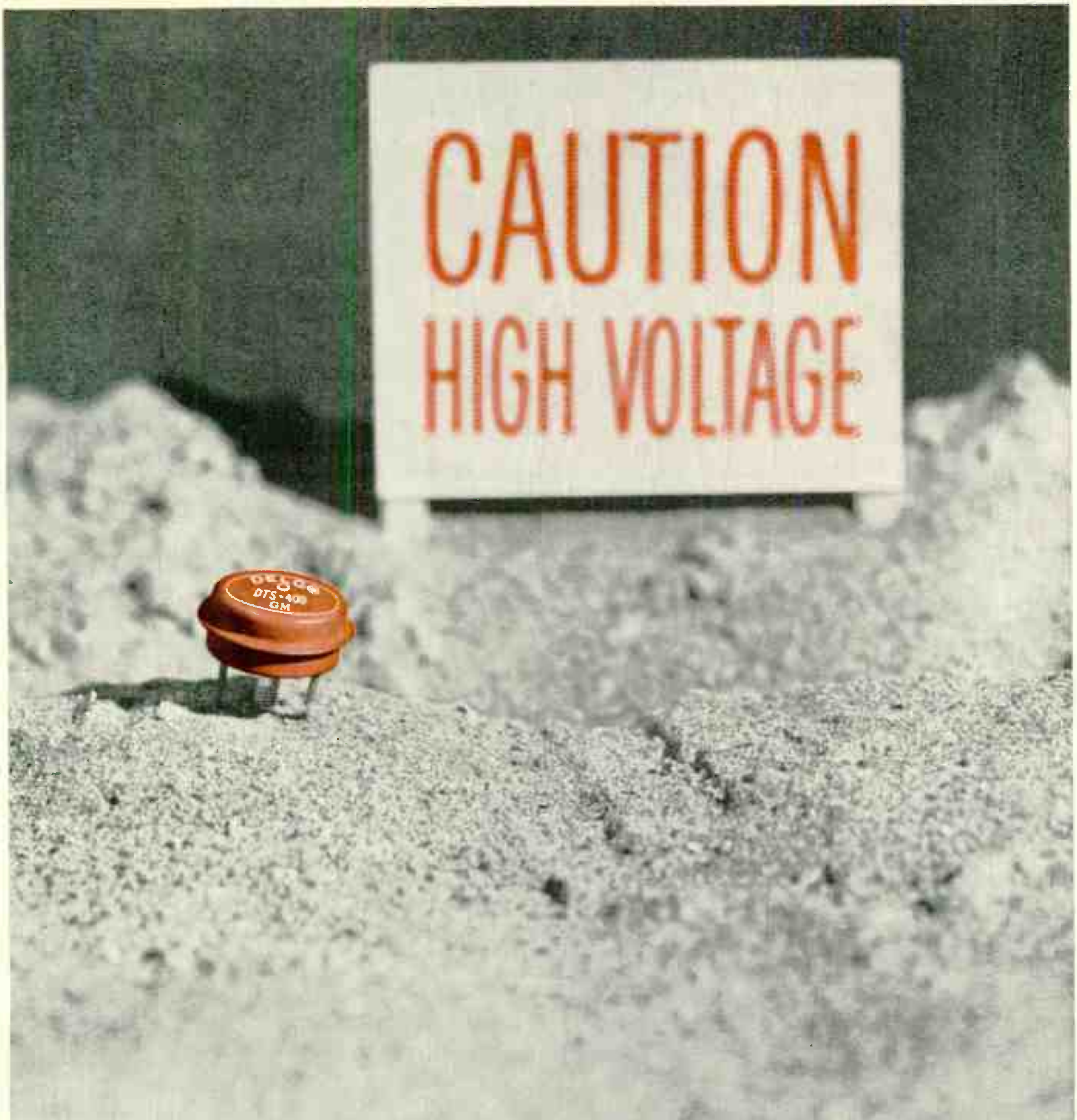
Dr. Schwartz received his Bachelor's degree in electrical engineering from Cooper Union, his Master's degree from Brooklyn Polytechnic and his Ph.D. degree in applied physics from Harvard University. He is a member of the American Society of Electrical Engineers, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

❖

Peter N. Sherrill (M'60) has resigned as Publications Manager of Hewlett-Packard Company to accept appointment as a senior associate and member of the Plans Board of West Associates, a Los Angeles and Palo Alto communications firm. He will assume direction of West Associates' Palo Alto facility. He will also undertake a full graduate program in mass communications research at Stanford University.

He joined H-P in 1954 following extended Navy service. He has directed the

(Continued on page 54A)



This is the new DTS-400 from Delco Radio . . . one of the highest voltage silicon power transistors available. The DTS-400 offers Vceo, Vcbo and Vces of 400 volts. Because of its high voltage capabilities and its ability to withstand high temperatures, this transistor offers a significant advancement in the art of power conversion.

The Delco DTS-400's capabilities make possible "direct to line" voltage hook-ups eliminating the need for transformers or other devices in between . . . and their related space and weight requirements. Production samples of the new DTS-400 silicon power transistor are available now to help you reduce the size, weight and cost of your power package. For complete engineering data, write or call our nearest sales office.

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DELCO RADIO

General Sales Office: 700 E. Firmin, Kokomo, Ind., Gladstone 2-8211—Ext. 500 • Area Code 317 • Division of General Motors, Kokomo, Ind.

IRE People

(Continued from page 52A)

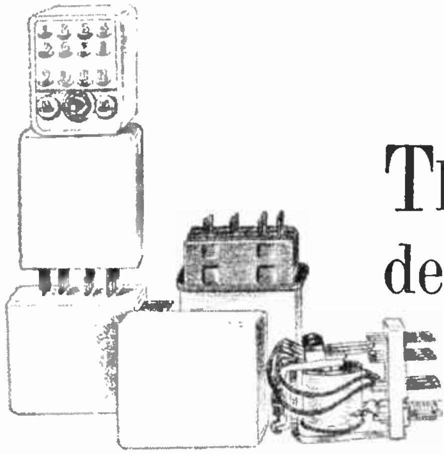
highly regarded technical literature program on behalf of Hewlett-Packard instruments and systems for eight years.

Mr. Sherrill is an engineering graduate of the U. S. Naval Academy. He is very active in Peninsula engineering and electronic industry affairs, and is a current executive committee member of the San Francisco section of the IRE. In 1961, he served as public relations chairman of the Western Electronic Show and Convention in San Francisco. He is also a past director of the San Francisco Chapter of the American Astronautical Society.

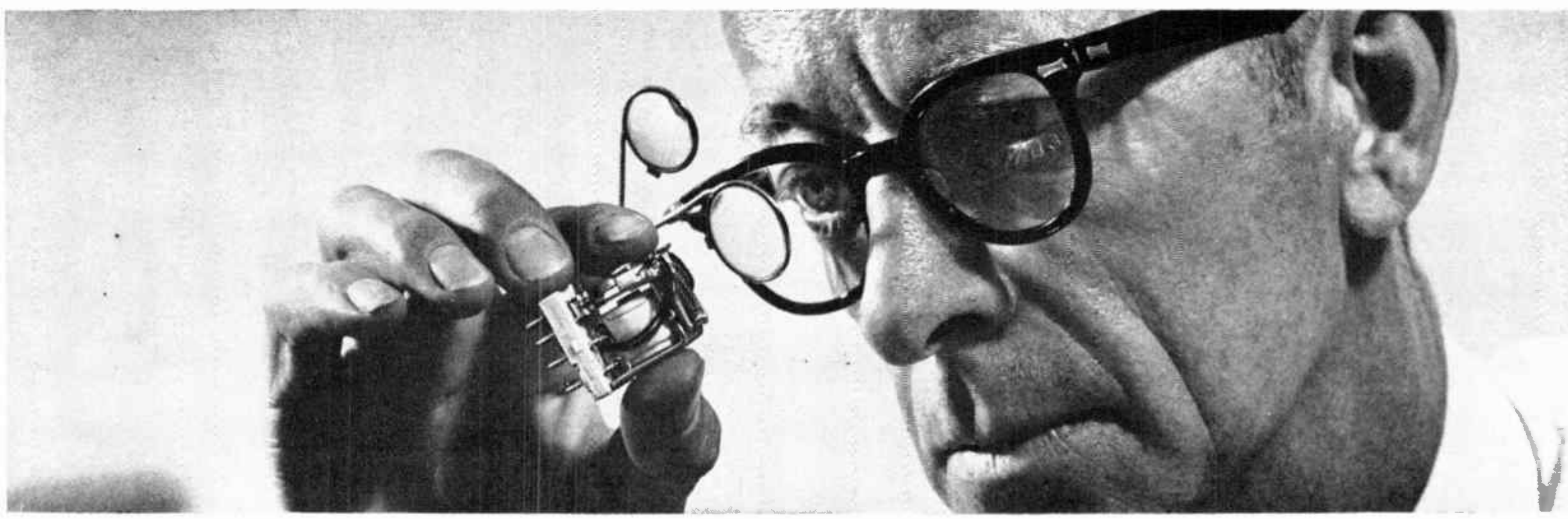
❖

Dr. Malcolm L. Stitch (SM'58) has been named manager of the newly created laser development department of Hughes Aircraft Co. He will work on systems applications of the laser to range finders and surveillance devices, and study effects of laser radiation on materials and biological specimens.

(Continued on page 56A)

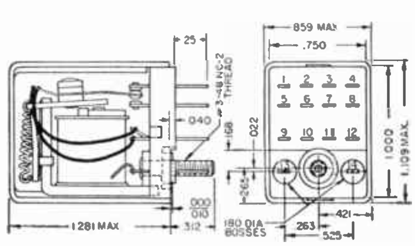


This new P&B a.c. relay deserves your critical appraisal



This remarkable new a.c. relay gives you these advantages:

SMALL SIZE. Only slightly larger than a cubic inch, yet has **MULTI-POLE CAPACITY.** Available in 4 form C or 2 form Z contact arrangements. Contacts are rated at 3 AMPS at 115 volts a.c., or 30 volts d.c. resistive. **LONG LIFE.** Engineered for millions of operations. **LOW POWER REQUIREMENTS.** Consumes only 1.2 volt-amps at nominal voltage. **SPECIFY** P&B's KHP17A.



KHP Series, Dust Cover

Available Hermetically Sealed a.c. or d.c. Specify KHS Series

PROVEN DESIGN

Thousands of these relays designed for d.c. are successfully being used in such diverse applications as telephone carrier equipment, citizens band transceivers and business machines. If you use d.c., specify KHP17D. For a.c. applications, ask for KHP17A.

WRITE FOR DATA SHEET

P&B STANDARD RELAYS ARE AVAILABLE AT YOUR LOCAL ELECTRONIC PARTS DISTRIBUTOR

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(Continued from page 54A)

From 1949 to 1951 he was an instructor in physics at Sarah Lawrence, Bronxville, N. Y., and the following year at Cooper Union College of Engineering, New York, N. Y. From 1953 to 1956, he was a research physicist at Varian Associates, Palo Alto, Calif., and since 1956 he has been with Hughes.

Dr. Stitch received the B.A. degree in French and the B.S. degree in physics from the Southern Methodist University, Dallas, Tex., both in 1947, and the Ph.D. degree in physics from Columbia University in 1953. He is a member of the Physical Society, the American Association of Physics Teachers, AAAS, Physical Society of Japan, New York Academy of Sciences, and Sigma Xi.

❖

Donald M. Stuart (A'39–SM'46–F'62), Vice President and General Manager of Hazeltine Technical Development Center, Inc., of Hazeltine Corp., has received the 1962 Pioneer Award of the IRE Professional Group on Aerospace and Navigational Electronics. He was presented with the award at the 14th Annual National Aerospace Electronics Conference in Dayton, Ohio, May 15, 1962, in recognition of his contributions to electronic aids to air navigation.

He was graduated from the University of Minnesota with a degree in electrical engineering. In 1929, he joined the National Bureau of Standards to help develop radio aids to air navigation and continued in this field with the Bureau of Air Commerce, the Civil Aeronautics Administration (CAA) and the Federal Aviation Agency (FAA). He was responsible for the development of the VHF marker beacon system, and also designed the modulation system for the VHF visual-aural range (VAR) and the instrument landing system (ILS) localizer. As Director of Technical Development for CAA, he was responsible for the development of the VHF omnidirectional range (VOR) and was instrumental in bringing about its acceptance as the international standard en route navigation aid. In 1959, he was elected Vice President and General Manager of Hazeltine Technical Development Center, Inc.

A former chairman of the Indianapolis Section of the IRE, Mr. Stuart has been honored by the National Business Aircraft Association, the National Civil Service League and the U. S. Department of Commerce in the past.

❖

(Continued on page 58A)

Coming soon!
IRE DIRECTORY!

Sanborn data amplifiers

(Three types available now — more on the way)

(Specifications and prices subject to change without notice; prices are FOB Waltham, Mass.)

Match amplifier characteristics much more closely to your over-all *system* requirements — and pay for only the performance you need — by choosing from these newly-developed, all solid-state DC data amplifiers now available from Sanborn. Ask your local Sanborn Sales-Engineering Representative for complete specifications, application help and a copy of the Industrial Division Catalog — or write the Main Office in Waltham.

• • • • •

Wide Band, Floating Input—Floating Output "FIFO"

Bandwidth DC to 3 db down at 10 KC • Input isolated from output • Max. gain 1000, smooth gain covers intermediate ranges or switch out for calibrated gains of 1000 to 50 • Input impedance 100 meg. min. at DC, output impedance 60 ohms • Output capability ±10 V at 10 ma • Common mode rejection (1000 ohms in either input lead) 160 db at DC, 120 db at 60 cps • Linearity ±0.1% of 10 V full-scale at DC • Recovery from 500% overload is 300μsec to 1% of f.s. output • Recovery from 20 V overload is 1 millisecond to 1% of f.s. output • Model 860-4000 "FIFO", $825. Model 860-4000P (grounded output ±5 V at ±100 ma, impedance less than 1 ohm), $900.

DC—50 KC, 3-Terminal Floating Amplifier

Gain 1000 to 10 in 1, 2, 5 ratios; does not phase invert • Input impedance 100 meg. at DC • Output ±10 V ±100 ma, impedance less than 0.2 ohm • Linearity ±0.01% of 10 V output • Gain stability ±0.01% at DC at constant ambient for 40 hours • Model 860-4200, including internal power supply, $650.

Narrow Band, Floating Input — Floating Output

Bandwidth DC to 3 db down at 100 cps • Optional plug-in output filters to limit bandwidth • Floating input isolated from floating output • Gain 1000 to 10; fixed step attenuator, gain trim and zero trim controls • Input impedance 300,000 ohms min., output impedance 75 ohms • Output ±5 V, ±2.5 ma • Linearity ±0.05% of 5 V output • Recovers from ±10 V overload in 200 ms • Common mode rejection (1000 ohms in either input lead) 130 db at 60 cps • Model 860-4300, $425.

INDUSTRIAL DIVISION
SANBORN COMPANY
175 Wyman Street, Waltham 54, Massachusetts
A SUBSIDIARY OF HEWLETT-PACKARD COMPANY

IRE People

(Continued from page 56A)

The appointment of **James L. West** (A'47–M'55–SM'55) as Vice President of Intercontinental Instruments Inc., Farmingdale, N. Y. was recently announced.

He was previously associated with the General Applied Science Laboratories, Inc. where, as Supervisor of Advanced Development, he contributed significantly to studies in the fields of low frequency electromagnetic wave propagation and sampled data systems. He came to GASL following broad experience at Link Aviation, Inc. as Manager of Analog Computer Development and, subsequently as Director of their Binghamton Laboratories. He is a former Member of the Staff of both the Bell Telephone Laboratories and Columbia University's Electronics Research Laboratory, where he also served as an Instructor in the Department of Electrical Engineering.

A Graduate of the College of the City of New York and Columbia University, Mr. West is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi. He holds numerous patents and is the author of several papers on computing devices.

❖

Ralph E. Willison (M'60) has been named Manager, Engineering and Development, of the Electronics Division of Research-Cottrell, Inc., Bound Brook, N. J. He was formerly Chief Electrical Engineer for Research-Cottrell. The firm's new Electronics Division has been formed to design and manufacture custom high-voltage, high-energy power supplies, controls, and related equipment, for such applications as radar, electron-beam processing, plasma and ion propulsion research, and missiles and space vehicles.

Before joining Research-Cottrell in 1949, he was associated for several years with the Naval Research Laboratory, Field Station, Boston, Mass., working especially on pulse techniques and special radar systems. During his wartime Navy service he was assigned to the NRL and the Massachusetts Institute of Technology's Radiation Laboratory. At Research-Cottrell, Willison's work has included development and design of high-voltage, high-energy power supplies and controls. He holds several patents in these fields. A graduate of Case Institute of Technology, Mr. Willison is a member of Eta Kappa Nu.

❖

Melvin N. Abramovich (S'36–A'46–M'47–SM'50) has recently retired from active service with the U. S. Air Force and has joined Frederick Research Corp., Wheaton, Md., as Director of Test Equipment Engineering. In this capacity he will be responsible for the planning, coordination, and direction of all electronic test equipments, techniques, and standardization programs handled by the company.

He is well-known for his work in the test equipment field. His efforts led to the establishment by the Secretary of Defense of the Department of Defense Electronic Test Equipment Coordination Group, which helped formulate a well-defined and coordinated program of research, development, and standardization in that field. He served as the Group's first Executive Secretary and later as its Chairman. He was also instrumental in establishing the Joint Army-Navy-Air Force PROJECT SETE at New York University.

Mr. Abramovich is a graduate of the Institute of Technology of the University of Minnesota and has received additional training at Massachusetts Institute of Technology and George Washington University. He also attended the Army's Signal Corps School at Fort Monmouth, N. J. He later joined the Curtiss-Wright Corp.

In 1940 he was called to active duty and served in various electronic and communications assignments, including the Communications Department of the

(Continued on page 60A)

From Raytheon / Rheem:

1N3730/RD750 subminiature power diodes feature high conductance—1 Amp @ 1 volt—at nanosecond switching speeds

High-speed, high-current core driving This Raytheon/Rheem .107"-diameter silicon diode combines high power dissipation plus ultra-high conductance — typically 1 Amp @ 1 volt. Result: cooler junctions and more reliable operation. Switching speed is 15 nanoseconds max. at 10 mA ($V_R = -6V_L$, $R_L = 75$ ohms); power dissipation rating is 750 mW — three times that of conventional diodes. The high current capabilities of the 1N3730 provide substantial operating stability for greater reliability.

Miniaturized power supplies The 1N3730/RD750 marks a big step forward for low frequency power supply design. Extremely fast turn-on time — V_{fr} (peak) is typically 1.0 volt at $I_F = 750$ mA — prevents impulse distortion and undesirable voltage feedback. Superior turn-off results from a low stored charge (typically less than 20 picocoulombs per mA).

Computer switching This subminiature diode is ideal for a large number of computer applications because of its wide switching current range — 1.0 mA to 5 Amps. Actual specification at several current levels is your assurance of ultra-fast reverse recovery over a wide range. Direct correlation to stored charge is also provided.

For complete details, please call or write the Raytheon Field Office nearest you, or write Raytheon Semiconductor Division, 900 Chelmsford Street, Lowell, Massachusetts.

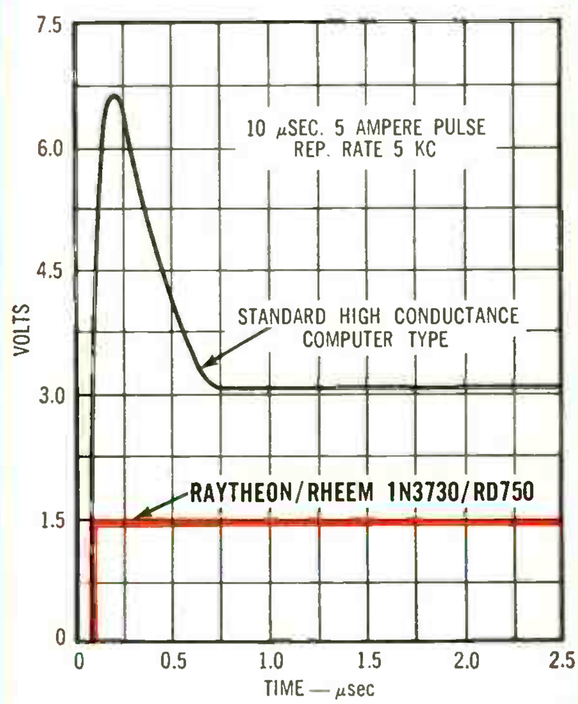
RAYTHEON/RHEEM 1N3730/RD750

KEY PARAMETERS

- Switching Speed...... 15 nanosecond max.
- Forward Conductance.. 1 Amp @ 1 volt typical
- Power Dissipation..... 750 mW max. @ 25°C.

FAST TURN-ON TIME

FIVE AMPERE PULSE SCOPE TEST — COMPARING TURN-ON TIME OF 1N3730/RD750 WITH STANDARD COMPUTER DIODE



SEMICONDUCTOR DIVISION

LOWELL, MASSACHUSETTS

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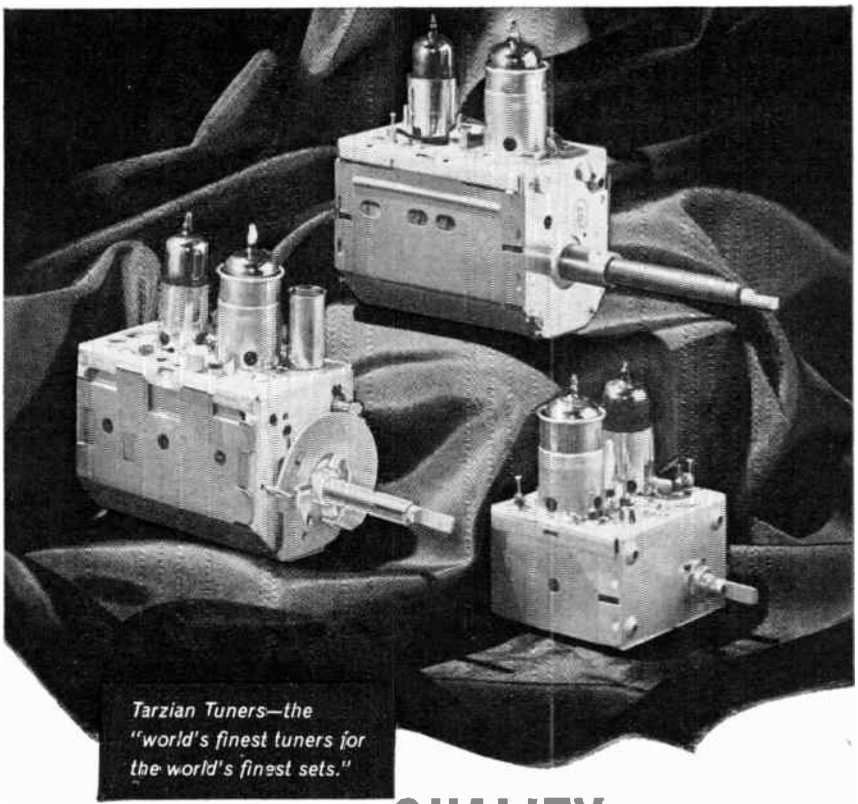

IRE People

(Continued from page 58A)

Armored Force School at Fort Knox, Research and Development Division of the Office of the Chief Signal Officer, Watson Labs. (now Rome Air Development Center), and AAF Liaison Office of the Navy's Bureau of Ships and Headquarters, USAF. While overseas during World War II, he served as Deputy Chief Radar Officer of the Mediterranean Allied Coastal Air Force. Upon returning to civilian status in 1946, he joined Cambridge Electronics Corp., Baltimore, Md. In 1947 he was again recalled to active duty to become the technical liaison officer for the Engineering Division of the Air Materiel Command with duty station at the Naval Research Laboratory. He later served in the Electronics Directorate at Headquarters, Air Research and Development Command. In 1954 he was selected to become a member of the staff of the Director of Electronics for the Assistant Secretary of Defense (R & D). From 1957 until his retirement from the Air Force, he was Electronics Officer and Chief of the ARDC (later Air Force Systems Command) Regional Offices, Washington, D. C.

Mr. Abramovich has been a member of the Panel on Tubes, Panel on Component Parts, Panel on Test Equipment, and Advisory Group on Reliability of Electronic Equipment (AGREE) of the former Research and Development Board of the Department of Defense. He was also a member of Joint Test Equipment Subpanel of the Joint Communications-Electronics Committee, Joint Chiefs of Staff, and of the RETMA Reliability Committee and the EIA Value Engineering Committee. He is a member of IRE Professional Groups on Engineering Management and Military Electronics, Armed Forces Communications-Electronics Association, American Association for the Advancement of Science, Institute of Navigation, and is a Charter Member of the Engineers Club of Washington and of the Society of American Value Engineers. He is a Registered Professional Engineer of the District of Columbia.

❖

J. Chuan Chu (A'48–M'55) has been appointed Associate Director of the Product Planning Division of Honeywell Electronic Data Processing. He was former Director of Engineering, Univac Division, Sperry Rand Corp. He will be responsible for formulating new equipment programs and broadening the Honeywell EDP product line.

He had been with Univac since 1956. For several years he was Chief Engineer on the LARC Project. He also served as Manager of Product Planning, Manager of Commercial Engineering, and Manager

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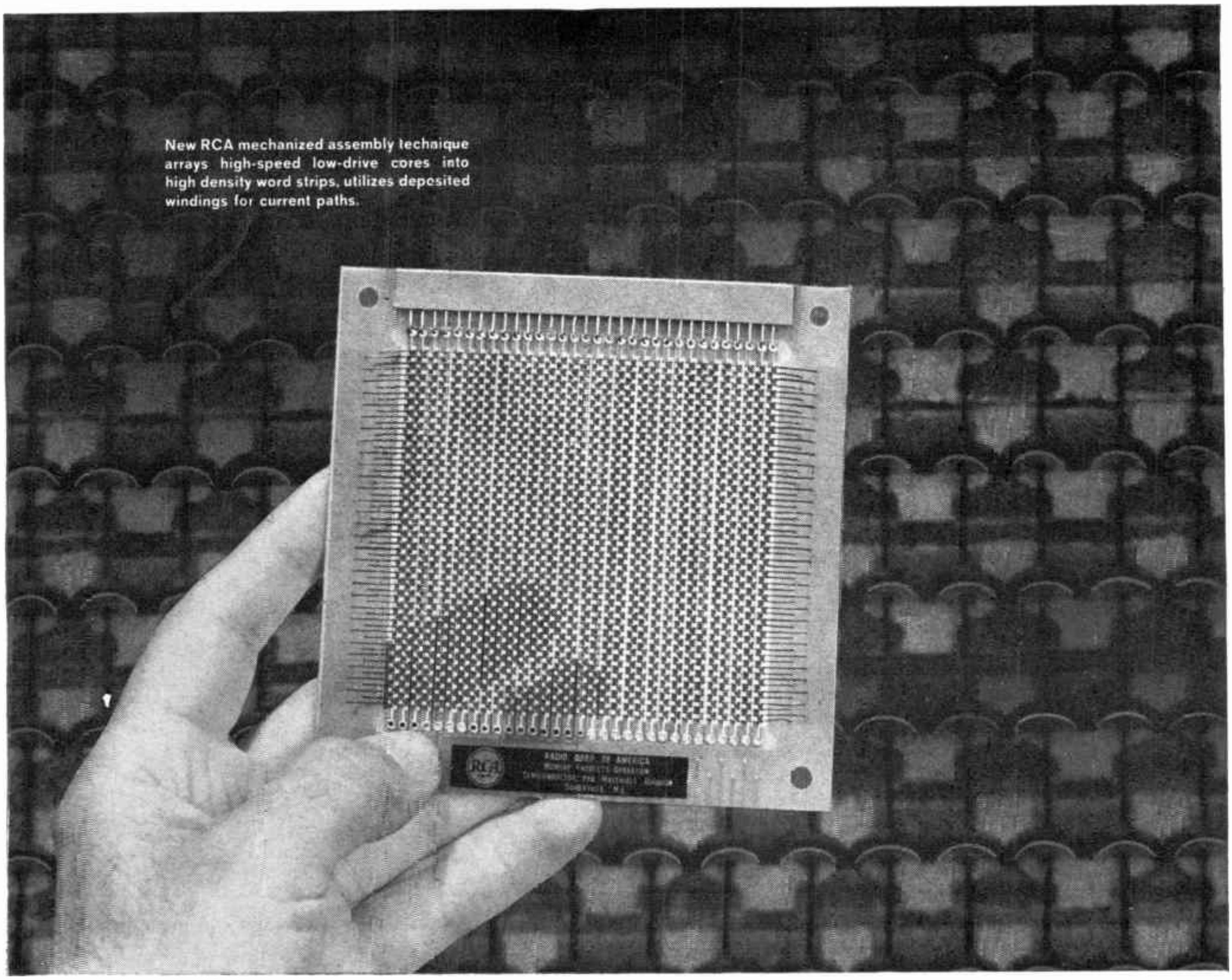

This Revolutionary New RCA Memory Stack

Completes A Full Cycle In 300 Nanoseconds With Only 350 ma Drive

Now, a major advance in Ferrite Stack Design and Construction by RCA makes 65-Nanosecond Switching a reality.

Here is the industry's first commercially available Microferrite Memory Stack with complete read/write cycle time of 300 nanoseconds at drive current levels below 350 ma—bit outputs of 50 mv.

This revolutionary two-core-per-bit word-address system bypasses today's experimental memory techniques by using proved, reliable ferrite cores in a high-density array of advanced design. Check these important benefits:

- **High Packing Density**... 1,000 to 2,000 bits per cubic inch.
- **Superior Stability and Ruggedness**... Printed wiring assures positive, rigid contact to each core. Planes designed to meet Military Mechanical and Environmental Specifications.
- **Precision Uniformity**... Mechanized fabrication eliminates many hand-assembly variables.
- **Outstanding Reliability**... Mechanized production techniques permit more precise control of each fabrication step—produce a rugged, high-reliability structure.
- **Broad Capacity Range**... Available in 32 word x 30 bit size, and in any multiple of this size.
- **Plug-In-Convenience**... Each stack incorporates standard plug connections for fast, easy installation.
- **Complete Service**... Whatever your requirements, custom or RCA standard, your local RCA Semiconductor and Materials Division Field Representative is prepared to provide a completely coordinated application service for all RCA Computer-Memory Products. Call him today at your nearby RCA Field Office.

For complete technical information on new RCA Microferrite Memory Stacks, write RCA Semiconductor and Materials Division, Commercial Engineering, Section FT-9, Somerville, N. J.

TENTATIVE DATA
TYPICAL DRIVE REQUIREMENTS AT 25°C

	Amplitude (ma)	Rise Time (nsec)	Duration (nsec)
Read Pulse	350	30	100
Partial Write Pulse	250	20	45
Digit Pulse	70	15	85

BIT OUTPUT (Two-Core/Bit Word-Address)

		Undisturbed '1' (mv)	Undisturbed '0' (mv)
Bit Out-Puts	Amplitude Sensing	60	12
	BiPolar Sensing	+50	−50

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IRE People

(Continued from page 60A)

of the Division's Philadelphia operations. At the time he joined Honeywell, he was responsible for research, product development and systems planning and was Chairman of Univac's Standards Council. He had previously been a Senior Scientist and Manager of the Computer Department, Argonne National Laboratory, University of Chicago; Professor of Electrical Engineering and Acting Department Head, Fournier Institute; Senior Engineer, Reeves Instrument Company; Research Associate and Instructor, University of Pennsylvania; and Television Engineer, Philco Corp.

Mr. Chu is a member of the Research Society of America, Society for the Advancement of Management, American Management Association, Eta Kappa Nu and Sigma Xi.

❖

Election of **Christopher Buff** (A'45–M'48–SM'55) as Vice President and Chief Engineer of American Cable and Radio Corporation has been announced. He will be responsible for all engineering activities of the Corporation's four operating companies. He joined AC and R in 1946, and has served in various engineering capacities since that time. He holds two patents and has three others pending in the radio telegraphy field.

❖

John M. Clayton (S'16-A'23-M'26-SM'43) will retire this month after 30 years with General Radio Company. Born in Arkansas, he completed his secondary education at Little Rock High School in 1917, and went on to the Student Army Training Corps program at Cornell University during World War I. After the War, he became active in the American Relay League and was appointed Technical Editor of *QST* the radio amateur trade magazine. He was subsequently asked to serve as Secretary of the IRE, where he was instrumental in a substantial increase in its membership.

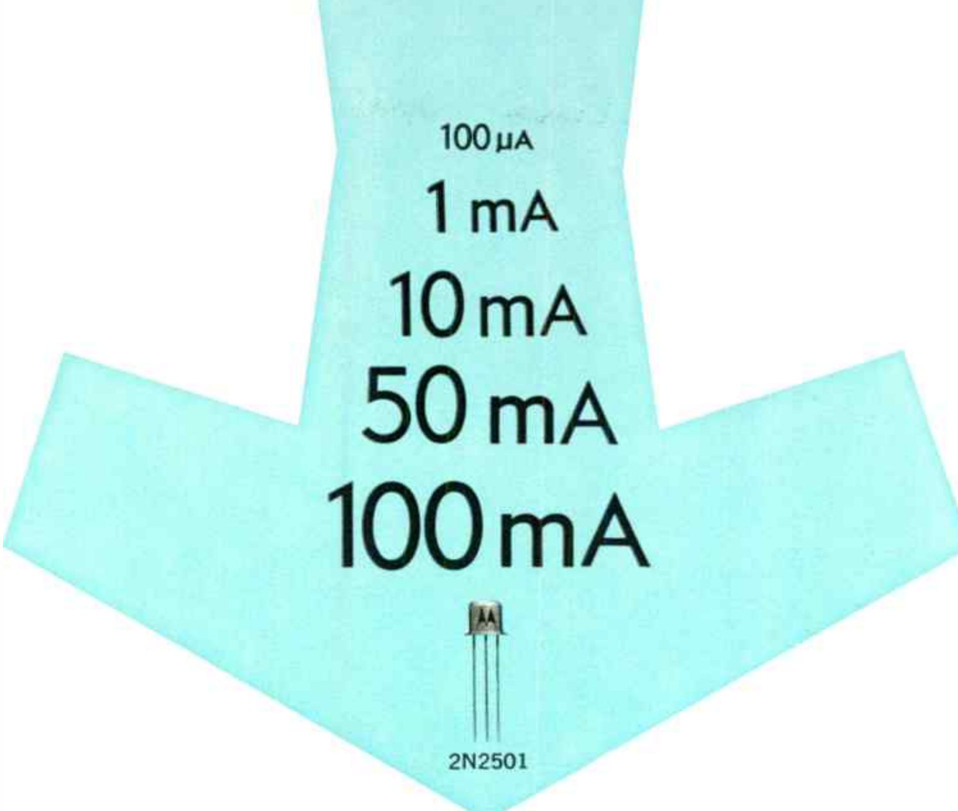
In 1930 he accepted the position of Manager of Operations of the Globe Wireless Limited, a subsidiary of the Dollar Steamship Lines (presently known as the American President Lines).

Mr. Clayton began his career with General Radio as a designer of amateur radio equipment and parts in 1932. In 1935 he was appointed Manager of the Advertising Department. The only break in his service occurred when he left temporarily from 1942–1944 to help in the operations of the Naval Research Laboratories. His work for the Navy won him the Meritorious Civilian Service Award.

❖

(Continued on page 66A)

another Si series from **MOTOROLA**



the most *completely specified* silicon epitaxial planar logic switch!

Whether you're designing switching circuits at 100 µA or as high as 100 mA, you can design with confidence using the new Motorola 2N2501 NPN silicon epitaxial planar logic switch.

This new high-gain transistor is characterized over its optimum usable current range, with beta specified from 100 µA to 100 mA, including measurements at 1, 10, and 50 mA.

And, with the specified active region time constant and total control charge parameters, you can more closely predict performance at various operating conditions (using a standard formula) than ever before.

In addition, saturation voltage is specified at 10, 50, and 100 mA, with extremely low values for these critical ratings.

The Motorola 2N2501 (TO-18 package) is specifically designed for low-level logic switching in the 100 µA to 100 mA region, and is supported by fuller, more definitive specifications than available in any present device.

Units are immediately available to meet your production requirements, or if you have a present application in which you would like to evaluate this new type, contact your nearest Motorola District Office. An engineering representative will advise you how you may obtain free samples.

Boston / Chicago / Cleveland / Dallas / Dayton / Detroit / Garden City
Los Angeles / Minneapolis / New York / Orlando / Philadelphia / Phoenix / San Diego / San Francisco / Syracuse / Washington / Toronto

MOTOROLA 2N2501* PERFORMANCE SPECIFICATIONS

BV_{CBO}	40 volts (min)					
BV_{CEO}	20 volts (min)					
BV_{EBO}	6 volts (min)					
h_{FE} @ $V_{CE} = 1$ V	$I_C = 100\ \mu A$ 20 (min)	$I_C = 1$ mA 30 (min)	$I_C = 10$ mA 50 (min) 150 (max)	$I_C = 10$ mA @ $-55°C$ 20 (min)	$I_C = 50$ mA 40 (min)	$I_C = 100$ mA 30 (min)
$V_{CE(SAT)}$ @ $I_C = 10 I_B$	$I_C = 10$ mA 0.2 V (max)	$I_C = 50$ mA 0.3 V (max)	$I_C = 100$ mA 0.4 V (max)			
τ_S @ $I_C = I_{B1} = I_{B2} = 10$ mA	15 nsec (max)					
τ_A (Active Region Time Constant)	2.5 nsec (max)					

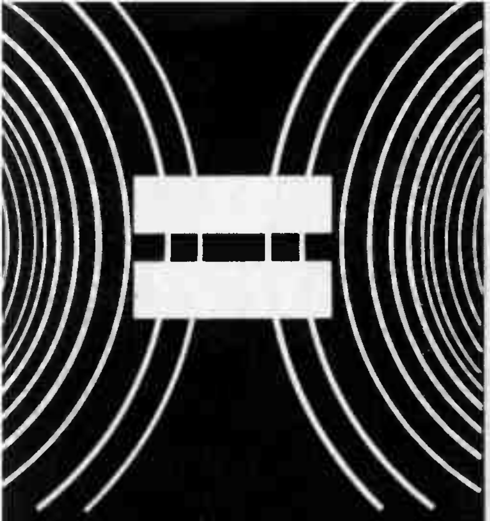
*TO-18 Package

The following Motorola silicon epitaxial logic transistor types are also available from your nearest Motorola Industrial Distributor or District Office:

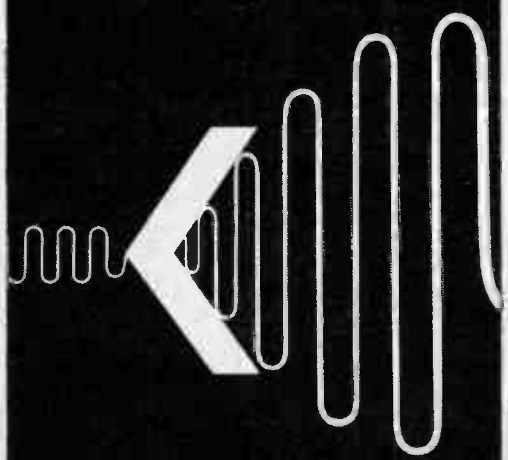
2N834	2N835	2N744
2N914	2N706	2N753
	2N708	

For your copy of the complete electrical specifications on the new Motorola 2N2501 transistor, call or write Motorola Semiconductor Products Inc., Technical Information Department, 5005 East McDowell Road, Phoenix, Arizona.

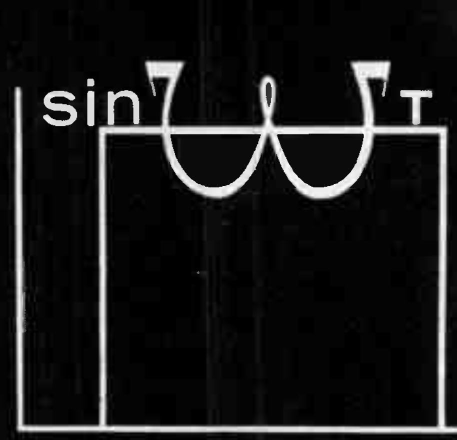
MOTOROLA
Semiconductor Products Inc.
A SUBSIDIARY OF MOTOROLA, INC. 2007



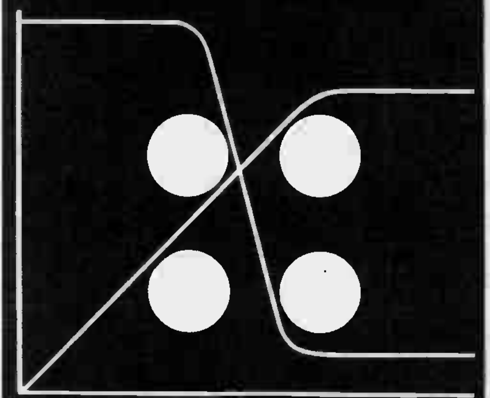
BULOVA
FREQUENCY STANDARDS



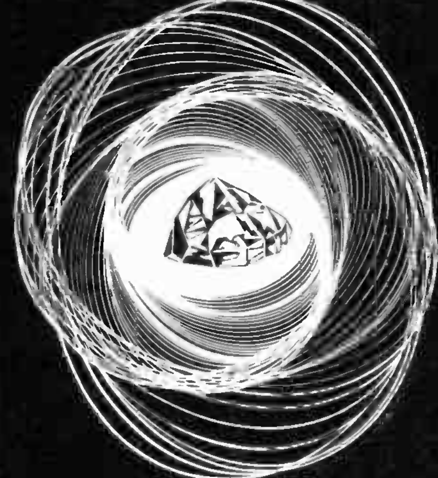
BULOVA
SERVO AMPLIFIERS



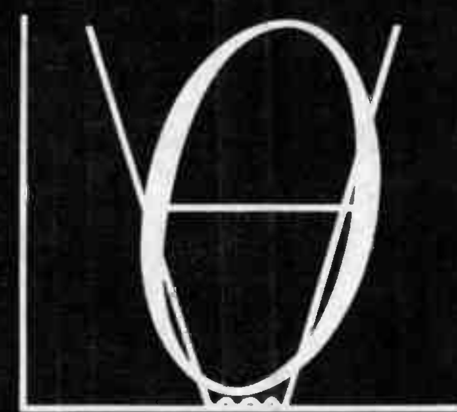
BULOVA
OSCILLATORS



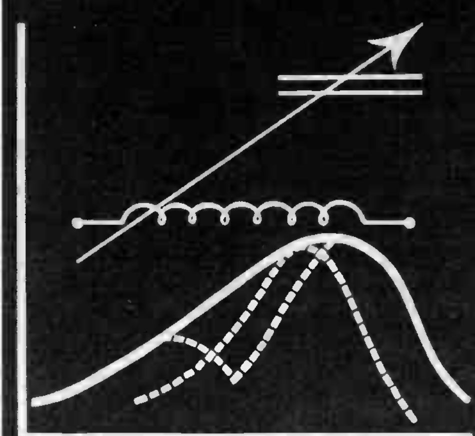
BULOVA
OVENS



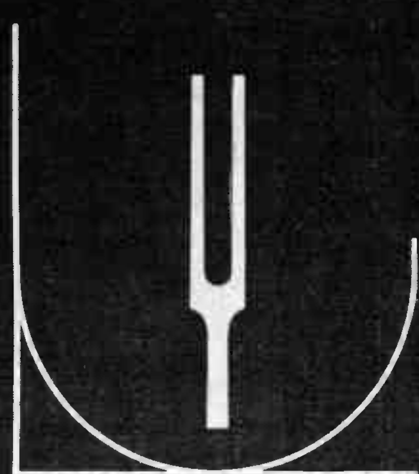
BULOVA
CRYSTALS



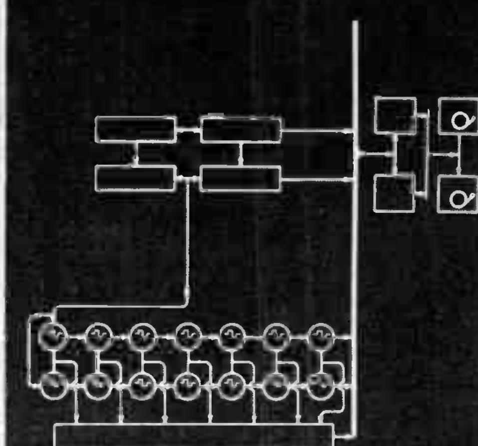
BULOVA
CRYSTAL FILTERS



BULOVA
COILS



BULOVA
TUNING FORKS



BULOVA
SYSTEMS & SUB SYSTEMS

(Continued from page 62A)

Harold B. Coleman (A'50–M'58) has been designated Manager, Satellite Interception, in the office of Defense Systems Studies of Aerospace Corporation's Systems Research and Planning Division.

Harold B. Coleman came to Aerospace Corp. in March, 1962 from the Lockheed California Company. While at Lockheed, he was the Guidance and Controls Staff Engineer in the Spacecraft Division. Prior to this, he was Project Manager for engine controls on PLUTO (nuclear Ramjet) Engine Controls at the Marquardt Corp.

He received a B.S. degree in 1946 and a M.S. degree in 1947, both in mathematics and physics, from the University of Michigan. He is a member of the American Rocket Society, the Association for Computing Machinery and the National Management Association.

❖

E. Paul Cote (A'52–M'56) has been appointed Manager of Special Products Marketing for MELABS, Palo Alto, Calif. He will be responsible for sale of MELABS special microwave systems. He was formerly with Zenith Radio Research Corporation, Menlo Park, Calif., in their microwave marketing department. His experience in the microwave field is comprehensive and extends over an eight year period. He is a graduate of Valpariso Technical Institute, and is a member of the Air Force Association and the Navy League.

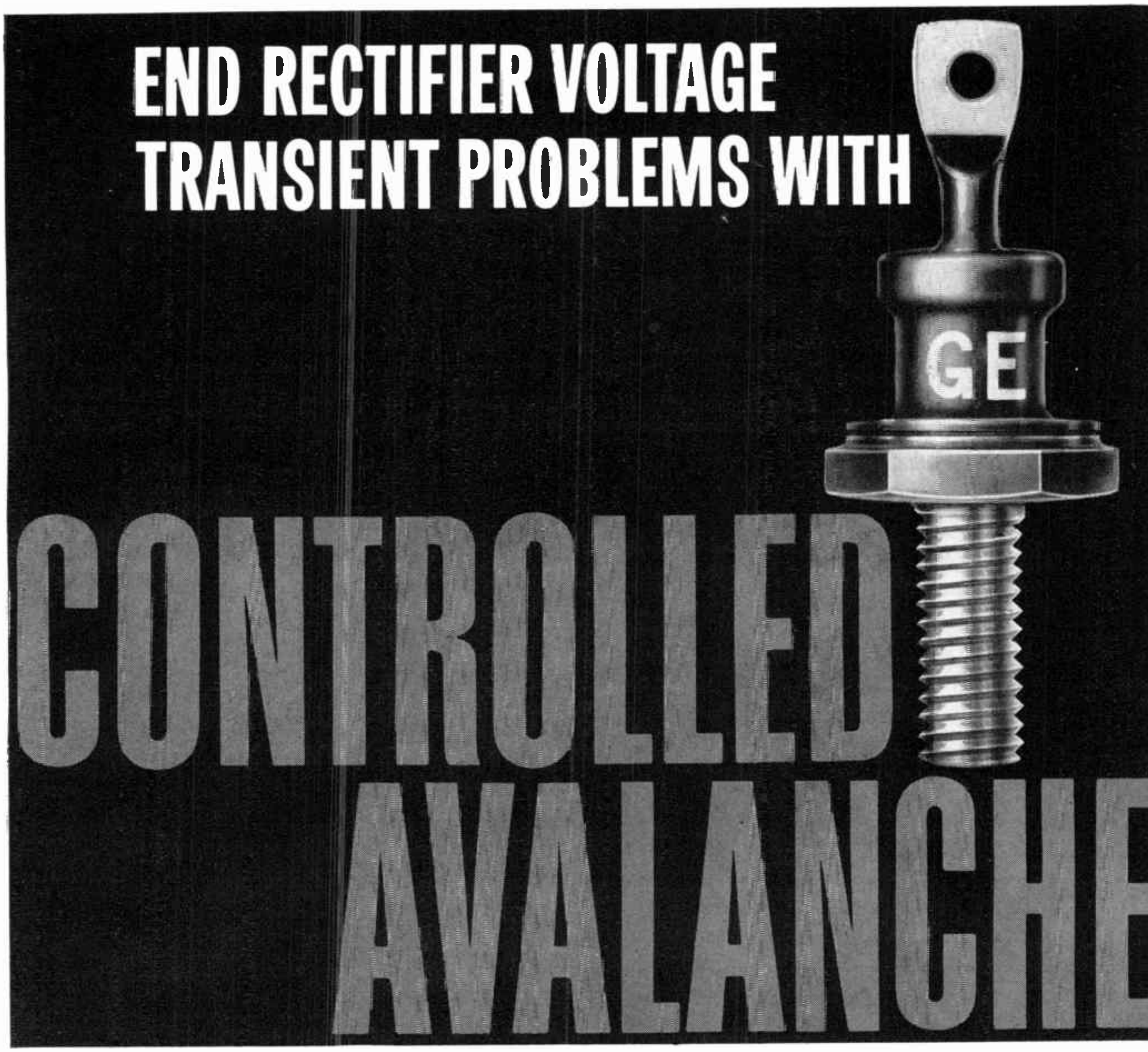
❖

Arthur F. Dickerson (SM'55) has been named Manager of Advanced Engineering for the Advanced Product Planning Operation (APPO) of General Electric's Electronic Components Division. He was formerly Advance Project Planning Manager for the operation. In his new post, he will be responsible for all engineering activities of APPO which serves as the master planning group in identifying and recommending new business opportunities for the departments of the Electronic Components Division.

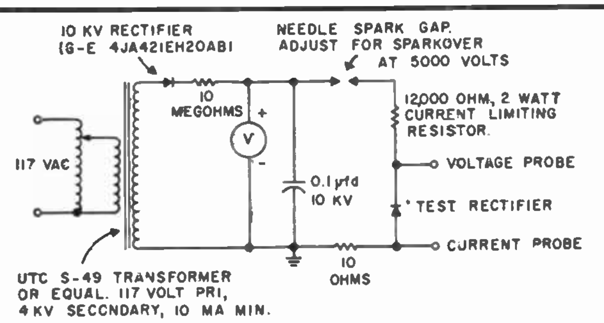
He has been with General Electric since his graduation from the University

(Continued on page 68A)

Coming soon!
IRE DIRECTORY!



The revolutionary 12 amp ZJ218 Controlled Avalanche Rectifier protects itself and the rest of the circuit up to 3900 watts peak power dissipation in the reverse direction. You get new high reliability standards up to 1200 PRV, protection of other circuit components, simplified rectifier series operation in high voltage applications, continuous operation in avalanche breakdown region at high voltage, and built-in "zener" diode protection even well beyond 1200 volts.



This reverse impulse test will prove how the Controlled Avalanche Rectifier withstands typical transient circuit voltages as high as 5000 volts, dissipates high levels of peak power in the reverse direction. Peak reverse power for rectifiers with avalanche voltages above 800 volts is over 250 watts in this circuit. (Connect a scope between the indicated voltage and current taps and ground to view impulse voltage and current.)

TEST IT YOURSELF . . . and prove beyond any doubt that your derating headaches are a thing of the past, that your transient voltage problems are solved more efficiently, more economically than ever before possible. The Reverse Impulse test circuit shown is all you need . . . along with the new G-E ZJ218 Controlled Avalanche Rectifier.

The ZJ218 Controlled Avalanche Rectifier is available in 600, 800, 1000 and 1200 PRV types. See your General Electric District Sales Manager and find out how to end *your* voltage transient problems with no derating. Or write Rectifier Components Department, Section 23174, General Electric Company, Auburn, New York. In Canada: Canadian General Electric, 189 Dufferin Street, Toronto, Ontario. Export: International General Electric, 159 Madison Avenue, New York 16, New York.

GENERAL ELECTRIC

IRE People

(Continued from page 66A)

of Texas in 1946 with a B.S. degree in electrical engineering. His experience includes service as engineer with the former Receiving Tube Sub-Department in Schenectady; commercial engineer in New York and Chicago; consulting engineer to the Air Froce under a G-E contract, at Wright-Patterson Field, Dayton, Ohio; application engineer, and Manager of Product Planning for the Receiving Tube Department at Owensboro, Ky. He was named Advance Project Planning Manager for AAPO in September, 1960.

Mr. Dickerson is co-author of an Air Force book, "Techniques for the Application of Electron Tubes in Military Equipment," published in 1952, and has written articles for publications in the electronics field.

William F. Eiseman (M'58) has been appointed Senior Project Engineer of Chesapeake Instrument Corporation's Underseas Systems Laboratory, where he will be in charge of advanced sonar systems development. He joined Chesapeake after six years with Westinghouse Electric Corporation as a computer engineer in all phases of computer development work. His responsibilities included memory design of the first airborne digital computer system and design of the Molecular Computer. Prior experience includes positions in computer engineering with ACF Industries, IBM and the National Bureau of Standards.

He received the B.S.E.E. degree from the University of Maryland in 1953. He currently has several patents pending, including nondestructive readout memories and electrically alterable memory systems.

Harlan W. Frerking (S'46–A'49–M'55) has been appointed manager of engineering for Microwave Electronics Corp., Palo Alto, Calif. He formerly was associated with Sperry Electronic Tube Division at Gainesville, Fla., where he was Head of Microwave TWT Tube Product Engineering. With the Sperry organization since 1949, he held similar responsibilities at Sperry Gyroscope Co., Great Neck, N. Y. During 1942–46 he was a communications officer with the U. S. Air Force.

Mr. Frerking received the B.S. and M.S. degrees in electrical engineering from the University of Michigan, where he was active in Eta Kappa Nu, Tau Beta Pi and Sigma Xi.

(Continued on page 70A)

NEW from National

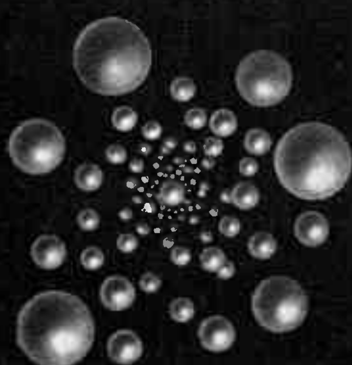
FOR THE FIRST TIME . . . AN ATOMIC FREQUENCY STANDARD FOR UNDER $10,000.

The most important breakthrough in frequency measurement since National's pioneering introduction of the Atomichron® in 1956 — the new NC-1601 — a primary cesium beam atomic frequency standard for under $10,000.!

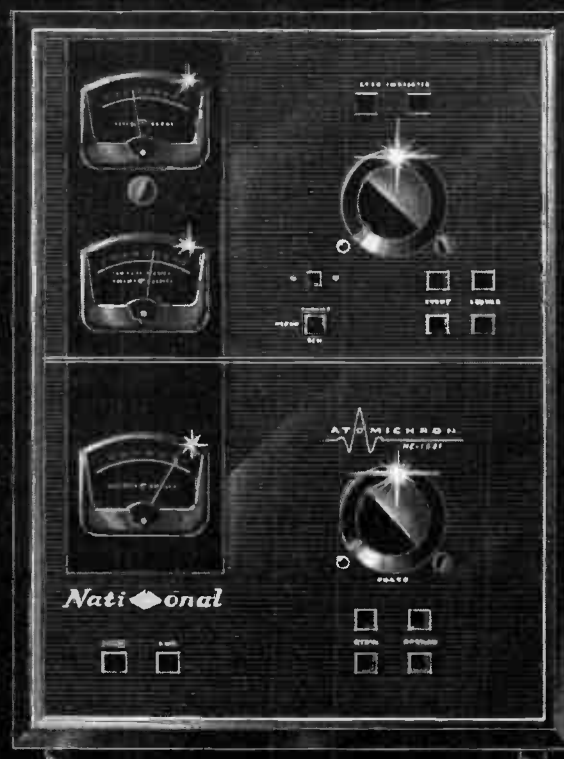
The introduction of the new NC-1601 Atomichron makes available for the first time at a price under $10,000. a frequency standard superior to any other commercially available equipment — with the exclusive operational advantages of National's higher-accuracy Atomichrons —fail-safe, self-calibrating, life-long, drift-free performance as a result of the use of National's proprietary Cesium Beam tube.

If you have a frequency standard application and would prefer a primary standard at less than the price of some secondary standards, we invite you to contact us on your letterhead for further information.

NC-1601 ATOMICHRON



Cesium Beam
Primary
Frequency
Standard



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NC-1501

Catalog Atomichron models for military, scientific, and industrial application with long-term stabilities of up to 5 parts in 10^{12} forever!

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National

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A Wholly Owned Subsidiary of National Company

IRE People

(Continued from page 68A)

William F. Garmon (M'60) has been appointed Director of Manufacturing of Interstate Electronics Corporation, Anaheim, Calif. He brings to Interstate an extensive background in administration, manufacturing and engineering.

Most recently he was General Manager—with full responsibility for manufacturing, sales, engineering and administration—of California Technical Industries, designers and manufacturers of aerospace instruments and test equipment. Immediately prior to this he was Vice President in charge of Manufacturing and Engineering for DIT-MCO, Inc., of Kansas City, Mo. He also has had a number of years experience in digital computer research with IBM Corporation, New York, N. Y. as a Technical Staff Engineer.

Mr. Garmon received the B.S. degree in electrical engineering and did graduate work in electronics and business. He is a member of the American Management Association, Western Electronic Manufacturers Association, and various other professional organizations.

❖

Anthony A. Guido (S'59–M'60) has been appointed Applications Engineer of PRD Electronics, Inc., Brooklyn, N. Y. Prior to joining PRD, he spent two years with the Digital Communications department of the RCA Surface Communications Division, New York, N. Y. as a project engineer. He was responsible for the supervision and planning of special digital communications test equipment for checking the "Minuteman" missile's communications system.

From 1955 to 1960, he was a member of the staff of the New York University's Electrical Engineering Research Department. At N.Y.U., he designed and developed new circuits, for which patents are pending, for use in countermeasures systems and character recognition circuits. In 1954, he joined the Bell Telephone Laboratories, Murray Hill, N. J., where he was technical aide in the Test Maintenance Department, responsible for the calibration and maintenance of frequency counters, oscilloscopes, and other electronic measuring equipment.

Mr. Guido received the B.E.E. degree from the New York University College of Engineering and a certificate from RCA Institutes, New York, N. Y. He is a member of Eta Kappa Nu and Tau Beta Pi.

❖

(Continued on page 72A)

Coming soon!
IRE DIRECTORY!

IRE People

(Continued from page 70A)

Leonard M. Jeffers, Jr. (S'33–A'35–SM'62) has been appointed Manager of Technical Publications for the Electronic Defense Laboratories (EDL) of Sylvania Electric Products Inc. He is responsible for all technical literature originated at EDL and for various phases of customer relations.

With EDL since he joined Sylvania in 1955, he has over 20 years of engineering and liaison experience. He previously was chief engineer for the City of San Francisco Utilities Engineering Bureau. During the Korean War, he served as a Captain in the U. S. Air Force at Wright-Patterson Air Force Base, performing duties as airborne early warning project officer. During World War II, he attended radar and electronics courses at Harvard University and Massachusetts Institute of Technology prior to assignments as a radar officer. He also served with the General Services Administration in San Francisco in electronics and aircraft work.

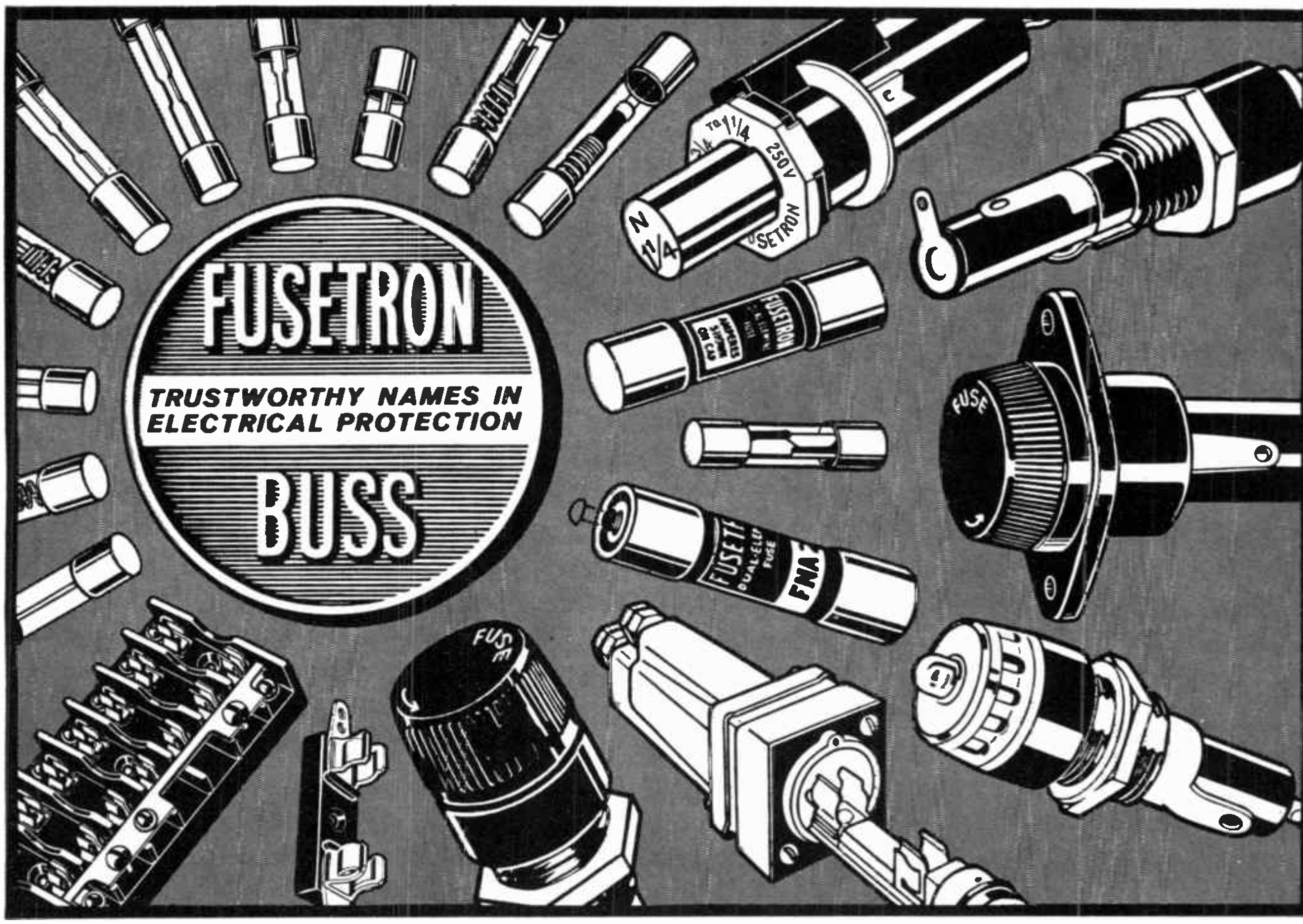
Mr. Jeffers received the B.S. and M.S. degrees in engineering from Stanford University. He is a member of the American Institute of Electrical Engineers.

❖

Robert L. Jones (S'50–A'52–M'54–SM'59) has been appointed Manufacturing Manager for Dickson Electronics Corporation, Scottsdale, Ariz. In his new capacity he will be responsible for the overall administration and direction of product managers and their operations. He was previously Production Manager for Dickson Electronics. His experience in the R and D, engineering, and production phases of the semiconductor industry was ideal preparation for his new position.

His experience includes development work on diodes and transistors at CBS-Hytron Semiconductor Division during 1952 and 1955. During the latter part of this period, he held the position of Chief Production Engineer. From 1955 to 1956, he was Production Manager of Hoffman Semiconductor Division, supervising manufacturing, quality control, material control, and production engineering. From 1956 to 1959 he was employed by the Semiconductor Division of General Electric Company as Project Engineer on switching, core driver, and medium power transistors. Immediately prior to joining

(Continued on page 74A)



Save Time and Trouble by standardizing on BUSS Fuses—You'll find the right fuse every time... in the Complete BUSS Line!

By using BUSS as your source for fuses, you can quickly find the type and size fuse you need. The complete BUSS line of fuses includes: dual-element "slow-blowing", single-element "quick-acting", and signal or visual indicating types . . . in sizes from 1/500 amp. up—plus a companion line of fuse clips, blocks and holders.

BUSS Trademark Is Your Assurance Of Fuses Of Unquestioned High Quality

For almost half a century, millions upon millions of BUSS fuses have operated properly under all service conditions.

To make sure this high standard of dependability is maintained . . . BUSS fuses are tested in a sensitive electronic device. Any fuse not correctly calibrated, properly constructed and right in all physical dimensions is automatically rejected.

Should You Have A Special Problem In Electrical Protection . . . BUSS fuse engineers are at your service—and in many cases can save you engineering time by helping you choose the right fuse for the job. Whenever possible, the fuse selected will be available in local wholesalers' stocks, so that your device can be serviced easily.

For more information on the complete line of BUSS and FUSETRON Small Dimension Fuses and Fuseholders, write for BUSS bulletin SFB.

BUSS: The complete line of fuses and fuse mountings of unquestioned high quality.

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IRE People

(Continued from page 72A)

Dickson Electronics Corporation in September, 1961, he was Production Manager for Silicon Alloy and Mesa Transistors at Hughes Semiconductor Division.

Mr. Jones is a member of AIIE, ASQC, and Tau Beta Phi, and was a past President of Alpha Pi Mu. He received the M.S. degree in industrial engineering, from Stanford University, the B.I.E. degree from Syracuse University (Cum Laude) and the B.E.E. degree from Rensselaer Polytechnic Institute.

❖

Leonard I. Kent (S'49–A'51–M'55–SM'58) has been appointed to the position of Vice President for Engineering of Antenna and Radome Research Associates, Westbury, N. Y. In this capacity he will be responsible for the company's design, research and development efforts which will include those in the fields of solid state and ferrite devices.

He was president of Consolidated Microwave Corporation whose assets were acquired by Antenna & Radome Research Associates. Among the previous positions he has held, have been of Vice President of Engineering of MSI Inc., Director of Engineering and Chief Engineer of Narda Microwave Corporation. Among his accomplishments have been contributions in the fields of directional couplers, ridged waveguides, millimeter wave techniques, bolometric and solid state devices. Several papers have been written by him on these subjects.

Mr. Kent is a graduate of The Cooper Union School of Engineering where he received a B.E.E. degree. He has also been awarded an M.E.E. and is completing requirements for the D.E.E. at Polytechnic Institute of Brooklyn. He is a member of the Committee on Waveguides and Fittings of the Electronic Industries Association, and of the Transducer Committee of the ISA.

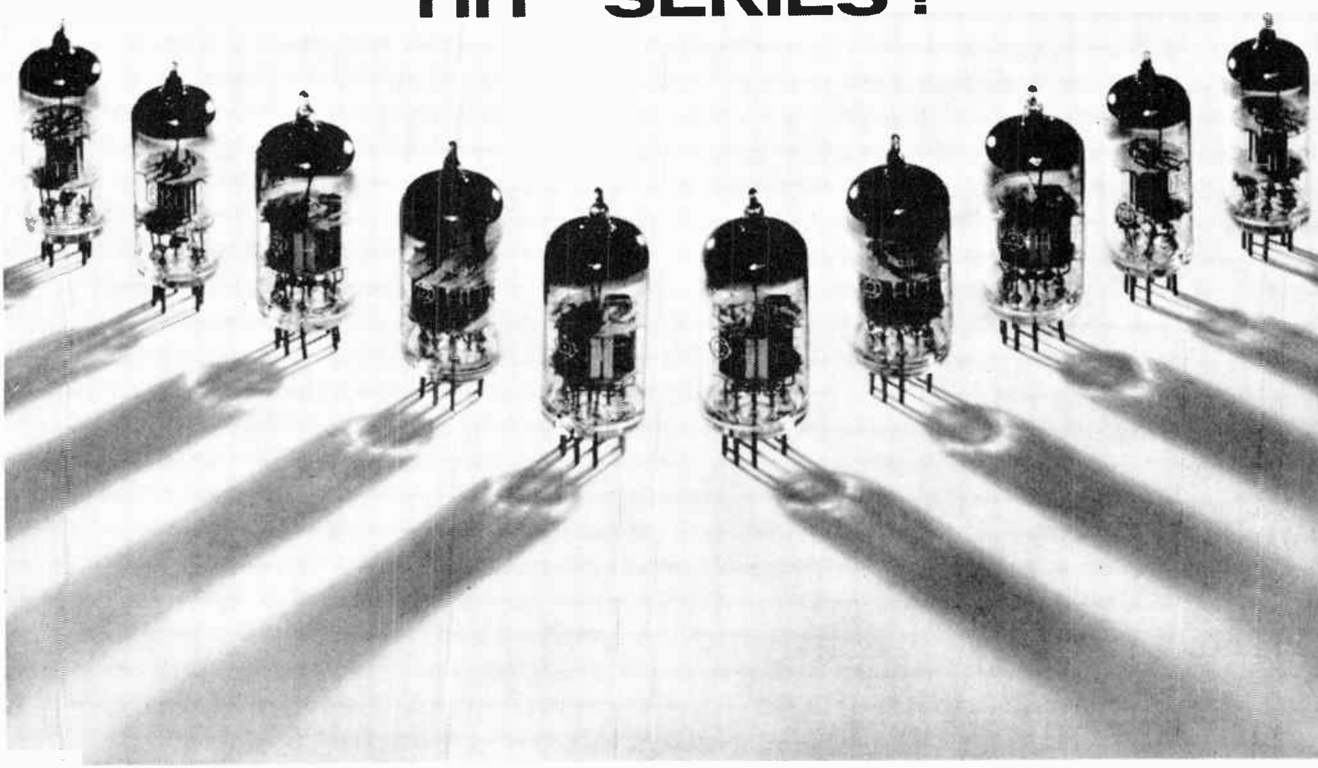
❖

Frank N. Kirby (S'49) has been appointed Riverdale Plant Manager for ACF Electronics, a division of ACF Industries, Incorporated. For the past year he had been Engineering Manager of the plant.

He joined ACF Electronics from the Missile and Space division of the Raytheon Company, Bedford, Mass., where

(Continued on page 77A)

WHAT IS THE "HH" SERIES?



The "HH" series is Hitachi's new superior line of television receiver tubes, the ultimate in far-reaching reception of television waves.

For RF amplifier of VHF television tuners, specify the 4R-HH2 and 6R-HH2 which feature very high transconductance, high sensitivity and low noise. These twin triode tubes replace the 4BQ7A and 6BQ7A without change of circuit.

For frequency convertor and local oscillator of VHF television tuners, specify 5M-HH3 and 6M-HH3 twin triodes which replace the 5J6 and 6J6 without change of circuit.

The "HH" series is another fine quality line from Hitachi, one of the most completely integrated electrical manufacturers in the world.

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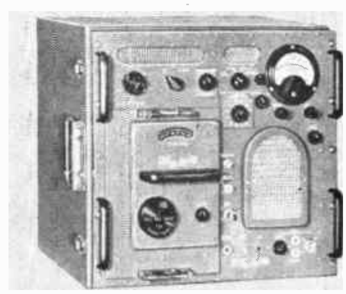
THE ONLY SIMPLE FORMULA FOR

COMPLETE

INTERFERENCE MEASUREMENTS

LOW FREQUENCY • Model NF-105

14 KC-1000 MC



HIGH FREQUENCY • Model NF-112

1000-15,000 MC

ENTIRE FREQUENCY RANGE

14 KC-15 KMC

Through years of hard field use Model NF-105 has acquired an outstanding reputation as a noise and field intensity meter for the frequency range from 150 kilocycles to 1000 megacycles. The versatility of this instrument has now been expanded through the introduction of a new tuning unit extending its coverage down to 14 kilocycles. What is more, this unique measuring equipment has been joined by Model NF-112 which covers the frequency range from 1000 to 15,000 megacycles. The same simplicity, accuracy and speed of operation and reliability of performance which made Model NF-105 so successful have been designed into Model NF-112. Each instrument uses an impulse generator as its calibrator; each combines in one basic unit the components common to all frequency ranges, including the power supply, calibrator, attenuators and metering circuits. All frequency determining components and circuits are contained in plug-in tuning units. Model NF-112, incidentally, uses one single antenna for the entire range from 1000 to 10,000 megacycles.
You save considerably in SIZE, WEIGHT and COST by letting these two instruments do your entire interference measuring job from 14 KC all the way up to 15,000 MC.

Approved for testing to 10 KMC under Specifications MIL-I-11748, MIL-I-61810, Category A and MIL-I-26600, Category A.

For complete technical information, send for Catalog 614.

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

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IRE People

(Continued from page 74A)

he was Technical Director. Prior to that he was associated with the Canadian Marconi Company, Montreal, and the Electronics division of American Machine and Foundry Company, Boston, Mass. He served four years with the RCAF during World War II on British Air Ministry experimental radar assignments and was a Reserve Technical Officer with the Royal Canadian Electrical Mechanical Engineers after World War II. He is a graduate of the University of Manitoba, Canada, and a member of the Professional Engineering Society of Quebec.

❖

Bernard G. Kuhse (M'59) has been appointed as Central Midwest Field Engineer for the Sprague Electric Company, Interference Control Field Service Laboratory, Vandalia, Ohio.

Prior to joing Sprague, he was employed by the Hallicrafters Company, Chicago, Ill. for seven years, during which he performed duties as mechanical engineer and as an electronics engineer. His mechanical engineering duties included the design and packaging of various airborne electronic systems and equipments. As an electronics engineer, he was the Radio Frequency Interference Control Engineer in Hallicrafters' Electronics Warfare Passive Division, being responsible for the design of RFI suppression techniques and RFI testing of a number of airborne systems.

Mr. Kuhse attended Iowa State College, Ames, Iowa, and the American Institute of Technology, Chicago, Ill., and holds a B.S.E.E. degree.

❖

Dr. Haldon A. Leedy (SM'46–F'61) has been elected to the Board of Directors of Stewart-Warner Corporation. He is director of the Armour Research Foundation of Illinois Institute of Technology.

He joined Armour Research Foundation as a scientist in 1938 after receiving the Ph.D. degree in physics from the University of Illinois. In 1950 he became Executive Vice President and Director of the Foundation. He also is a Director of Link-Belt Company; Signode Steel Strapping Company; Nuclear-Chicago Corporation; and Business Capital Corporation. He is a trustee of North Central College and of Chicago Theological Seminary.

Dr. Leedy is a member of the Sigma Pi Sigma, Sigma Xi and Tau Beta Pi, a fellow of the American Institute of Electrical Engineers, and a member of the American Institute of Physics and the American Physical Society.

❖

Professional Group on Audio

The IRE Professional Group on Audio is the oldest and one of the most active Groups in the IRE. Formed on June 2, 1948, the Audio Group now serves the specialized technical needs of over 4000 Group members who wish to keep abreast of progress in communication at audio frequencies, the audio portion of the radio frequency spectrum, and recording and reproducing.

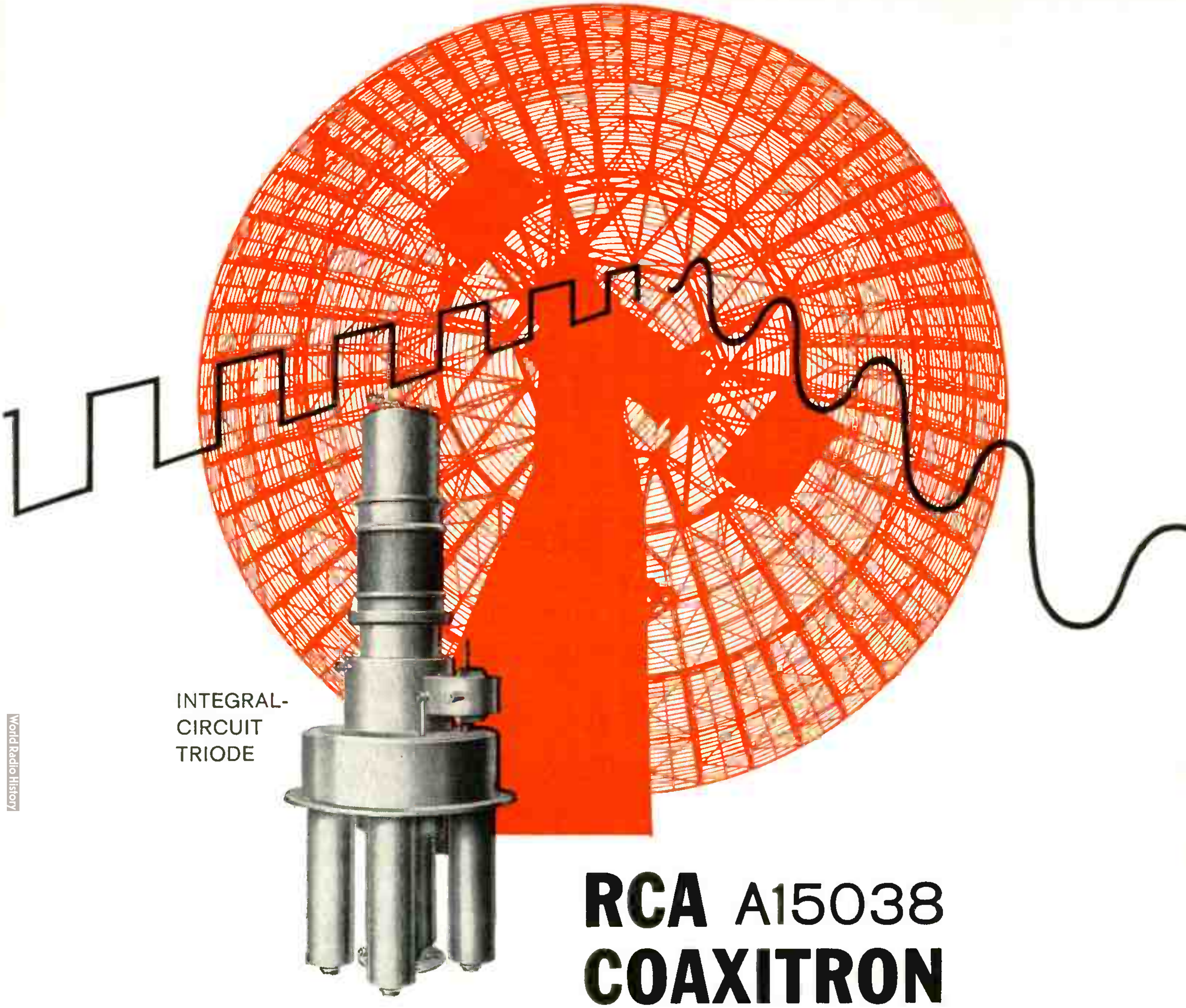
Perhaps the most important activity of the Group is its technical publication, called TRANSACTIONS, which is sent bi-monthly to all Group members who have paid the $2 assessment. The TRANSACTIONS provides an invaluable source of authoritative information concerning the latest developments in the audio field. In addition to technical papers the TRANSACTIONS contains informative technical editorials and news items pertaining to the audio field. To date, Group members have received 67 issues comprising 2250 pages of material devoted to their specific field of interest.

Supplementing its program of national meetings, the Audio Group has established 17 Chapters in cities throughout the country which, in cooperation with IRE Sections in those areas, hold frequent local meetings on subjects of timely interest to the audio engineer.

Thus the Audio Group, through its publications, national conferences, and Chapter meetings, renders a valuable service which no audio engineer can afford to be without.

Ernst Weber

Chairman, Professional Groups Committee



RCA A15038 COAXITRON

Integral Cavity Design Means Greater Reliability Over Broader Bandwidth

Broad bandwidth and outstanding power-output uniformity mark the new RCA A15038—a developmental Coaxitron—as a quality, high-mu super-power triode. A linear amplifier incorporating integral circuitry, the A15038 is designed for use in long-range search radar, broad-band multi-channel communications, and wherever electronic pulse-to-pulse frequency agility is important.

This RCA Triode Coaxitron operates with exceptional power-output uniformity in the range of 400 to 450 Mc and covers the range of 385 to 465 Mc to the −3db power-output level.

In short pulse service (pulse duration: 30 μsec), its efficiency is 43 per cent—at a gain of 13 db. Power output capability is 5 megawatts, minimum.

Operation of RCA A15038 over this bandwidth is made possible by integrating the radio-frequency input and output circuitry, high-voltage blocking circuit and the gridded tube structure within a common vacuum envelope. The tube is suited to a variety of modulation techniques—amplifying at any frequency within the pass-band at which it is driven.

Developed by RCA for the Rome Air Development Center, A15038 Coaxitron also features a low temperature matrix-oxide filamentary cathode to provide high emission, long life, and economical operation. These and other features combine to provide greater power output, broader bandwidth, higher power gain, better stability and greater reliability.

For further information contact your RCA Industrial Tube Representative or write: Marketing Manager, Industrial Tube Products, RCA Electron Tube Division, Lancaster, Pennsylvania.

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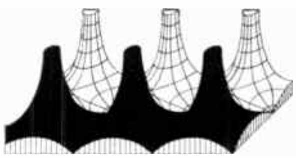
INDUSTRIAL TUBE PRODUCTS FIELD OFFICES—OEM Sales: Newark 2, N. J., 744 Broad St., HU 5-3900 • Chicago 54, Ill., Suite 1154, Merchandise Mart Plaza, WH 4-2900 • Los Angeles 22, Calif., 6801 E Washington Boulevard, RA 3-8361 • Burlingame, Calif., 1838 El Camino Real, OX 7-1620. **Government Liaison:** Harrison, N. J., 415 South Fifth St., HU 5-3900 • Dayton 2, Ohio, 224 N. Wilkinson St., BA 6-2366 • Washington 7, D.C., 1725 "K" St., N.W., FE 7-8500. **International Sales:** RCA International Division, Clark, N. J., FU 1-1000

September, 1962 *Vol. 50* *No. 9*

Proceedings of the IRE

Poles and Zeros



Seattle. In accordance with the established custom of periodically holding IRE Board of Directors' meetings near the grass roots, the Board met in Seattle, Washington, in May, in conjunction with the 7th Region Conference. Members of the Board had the opportunity to participate in the Conference and to view firsthand the strength, vigor, and problems of the region. President Haggerty gave an inspiring address, "Electronics, Machines, and Man," Director W. G. Sheppard chaired a papers session, and Director Dan Noble gave a "Crystal Ball" paper on "Electronic Nirvana." During odd moments, meeting breaks, and, for some, an extra day, the Directors managed to ride the monorail, scale the space needle, and see U. S. and International Exhibits. Most talked about were the exhibit of France in the Coliseum and "hardware art" in the Art Museum.

France. The French exhibit drove home, as only the French can do it, our heritage, the problems of today's world, and the Seven Golden Keys to the Future, which are education, research, study of modern life, modern methods of operational research and decision making, a new urbanism and, generally, an environment adapted to man, organization of leisure more in harmony with nature, and refinement of sensitiveness by the practice of art. Here's hoping the French exhibit will be reduced to film and circulated so that those who cannot visit Seattle can share in this emotional experience and the extended contemplative aftermath.

Electronics, Machine, and Man. In his Seattle address President Haggerty used vivid examples in reminding his Seattle audience that technological progress is a product of the present and very recent past and has moved at a greatly more rapid pace than any comparable achievement in history.

Among the factors enabling the United States to develop its needed effectiveness are natural resources, size of the nation and its market, governmental hands-off policy ("for most of the period") which encouraged individual effort, "frontier spirit" (exuberance, courage, willingness to risk, self-confidence, self-reliance, acceptance "of personal responsibility for one's own status in life"), the private enterprise system which rewards accomplishment, incentive and discipline resulting from profit-making ("succeed or fail"), strong emphasis on the useful resulting in demand for, obtaining of, and use of engineers in large quantity.

President Haggerty drew a word-picture of the engineer as "contributor of useful products and services to the larger society," and one who understands society's needs, is increasingly competent in theoretical and applied science, has a "solid feel" for the functioning of the economic system, is an able communicator of ideas—verbally and in writing, and has "knowledge of where our society came from and judgment as to where it is going."

Although American "organizational and managerial approaches . . . have given us our relative and enviable position in this world of ours . . . now the situation is changing." The international wage and salary differential remains "quite sizable"—in Japan, one-sixth that of the United States; in Europe, one-third—and although Western Europe's wages and salaries will increase more quickly in the next five years than those at home, in most cases they will still be at least 50 per cent below those in the United States. Wages and salaries in Japan can be expected to decline even more under pressure of a growing labor force.

Republican forms of government, the institution of private property, and private enterprise are "much more likely to succeed than communism," he said, and comprise a "much more tenable solution" of problems.

President Haggerty cited statistics reinforcing his statement that there is nevertheless "cause for concern about the scientific and technical manpower buildup in the Soviet Union, because it has become the principal source of Communist strength."

He said there are indications that the United States does not understand its own system:

—An improved living standard comes from our approach to organization, deliberately useful results of technology, and the improved productivity resulting from them.

—Attitudes seem to be developing that profits are "suspect"—at best, "a price to pay." He called profits "essential" to the American system.

—There is "failure to appreciate that a large part of the virtue of our system is its flexibility and responsiveness"; and, therefore, the system is restricted by wage laws, "undue fear of bigness, and exaggerated privilege for the mass voter, whether as represented by the farmer or labor organizations."

The engineer—particularly the electronics engineer—"has a critically important role to play in improving national effectiveness at a rate sufficient to ensure competiveness," because electronics offers "the tools . . . (which). . . augment our brain."

"It is our responsibility as engineers to improve the effectiveness of our productive machine at a rate sufficient to keep it competitive with those of our sister nations of the West and superior to the oppressive systems behind the Iron Curtain."

Standards. Copies of "Standards on Electron Tubes: Methods of Testing" are available free to members and subscribers who request them. See Page 1974 for information on this important document.

International. The establishment of Region 9 by the IRE Board, which results from a steady growth of IRE Membership in Europe and countries bordering the Mediterranean, represents a major step in development of the International character of IRE. The charter sections are Benelux, Egypt, France, Geneva, Israel, Italy, and the United Kingdom. Regional Director is H. R. Rinia of Phillips Electric, Ltd., former Chairman of the Benelux Section.—T. F. J.

G. A. Woonton

Director, 1962–1964

G. A. Woonton (A'40-SM'44-F'51) was born in London, Ontario, Canada, on July 9, 1906. He entered the University of Western Ontario in 1922 and was awarded the B.A. degree in commercial economics in 1925. After a few years with the Bell Telephone Company of Canada and a short time at the University of Chicago, he returned to the University of Western Ontario, and was granted the M.A. degree in physics in 1931. In 1955 he was given a D.Sc. degree (Hon.) by that University.

From 1931 to 1939 he held the joint appointment of Research Fellow in the Department of Physiology and Demonstrator in the Department of Physics at the University of Western Ontario. During that period he worked with a group concerned with the physiology of the cerebral cortex; in the Department of Physics he carried on research on the diffraction of electrons. With the advent of the war in 1939, he became a full-time member of the Department of Physics and a member of the team working with the National Research Council on the development of radar. In the Department of Physics, he held the appointments of Lecturer, Assistant Professor, Associate Professor, and (in 1946) Research Professor. In 1948 he was appointed Professor of Physics at McGill University, Montreal, P. Q. He became Director of the Eaton Electronics Research Laboratory in 1950, and in 1955, Macdonald Professor and Chairman of the Department of Physics.

After the war he became interested in analogies between optics and the properties of microwave radiation, in the interaction of electrons in long electron beams, and in the mechanisms responsible for the generation of noise in beams at microwave frequencies. During that same time he was a member of one of the teams engaged in the development of the prototype of the warning system now known as the Mid-Canada Line, and was also a member of the group organized at the Massachusetts Institute of Technology by the U. S. Navy to look into the problems of air defense of the North American Continent.

During the last five or six years his research has centered in the subjects which are sometimes combined under the heading of Quantum Electronics. He and his associates are interested in the interactions between the spin systems of paramagnetic ions and the vibrations of the crystal lattice, and in a group of phenomena primarily associated with the magnetic properties of solids.

Professor Woonton has served the IRE as a student representative at the University of Western Ontario, as a member of the Education Committee (1945–1946) and as Chairman of the London, Ontario section (1945). In the International Union of Scientific Radio (URSI) he was International Chairman of Commission VII (Electronics) (1952–1957). He was elected Vice President of the Union in 1957, a position which he still holds. He is a past President of the Canadian Association of Physicists (1948–1949), and a member of the American Physical Society and Sigma Xi. He was elected a Fellow of the Royal Society of Canada in 1950.

Scanning the Issue

IEEE Ballott (pp. 1892–1897).

How to Obtain the IRE Standards on Electron Tubes: Methods of Testing, 1962 (p. 1974)—This month the IRE is publishing the largest standard in its history. Because of its unusual size (160 pages), it was not feasible to include it in an issue of PROCEEDINGS. However, any IRE member or PROCEEDINGS subscriber may obtain one free copy by writing to IRE Headquarters and asking for Standard No. 62 IRE 7. S1 (IRE Standards on Electron Tubes: Methods of Testing, 1962). In addition to its large size, this standard is noteworthy for its comprehensive scope. Consisting of ten parts, it covers methods of testing receiving, cathode-ray, gas, microwave duplexer, photo, microwave, camera and cathode-ray charge storage tubes, as well as methods of measuring cathode-interface impedance and noise in linear twoports. A detailed description and contents listing continues on page 1975. The value of standards such as these to the profession is presumably well-appreciated by most engineers. But one wonders whether many people have any idea of the amount of work that goes into producing them. In the present case, 167 of the top electron tube specialists in the profession voluntarily devoted an estimated 50,000 man-hours over a seven year period in developing this Standard. This is comparable, no less, to the man-hours expended by the National Television System Committee in formulating the U. S. color television broadcasting standards in the early 1950's—probably the greatest cooperative effort our profession has witnessed. The result is a document which establishes a quantitative base for a major area of radio engineering and provides an indispensable tool for those working in the electron tube and allied fields. Interested readers are urged to write for their copy of the Standard promptly.

Quantum Effects in Communications Systems (Gordon, p. 1898)—In the radio frequency portion of the electromagnetic spectrum, the ultimate limit on the performance of a communication system is normally set by thermal noise. With communication at infrared and optical frequencies now becoming a distinct possibility, we are on the threshold of a frequency region where quantum effects come into play and, indeed, provide the ultimate limit to our ability to transmit information. This paper examines the information capacity of various communication systems, taking quantum effects into account, and establishes the relative efficiency of amplifiers, heterodyne and homodyne converters and binary counters as a function of frequency and signal or noise power. The results will be of timely value to anyone studying coherent optical communication systems, providing them a theoretical basis for evaluating such systems.

Negative Impedance Electrometer Amplifiers—Introduction (MacNichol, p. 1909)—This issue brings together three papers describing recent work done in laboratories in New York, Bethesda and Copenhagen on developing wide-band amplifiers having a very high input impedance. The papers are noteworthy in several respects. First, the laboratories in question are not electronics organizations; each is engaged in medical and biological research. Secondly, the amplifiers they have developed have a much higher input impedance than any previous wide-band vacuum-tube amplifier design. Finally, the technique by which this has been accomplished, namely, the use of negative capacitance, thus far seems to have been employed only by neurophysiologists in connection with the measurement of bioelectric potentials. It is hoped that with publication of these papers, this class of amplifier will find much broader application in fields such as photoelectric measurements, variable capacitance and piezoelectric transducers, electrostatic memory devices, and others. As a preface to the three papers which immediately follow, the Editor of the IRE TRANSACTIONS ON BIO-MEDICAL ELECTRONICS, who was instrumental in bringing these papers together, has prepared an excellent, brief introduction to the subject, to which he has added a short description of an interesting circuit which employs a field effect transistor instead of a vacuum tube.

Cathode Follower and Negative Capacitance as High Input Impedance Circuits (Guld, p. 1912)—In the first paper of the aforementioned series, the author is concerned with the problem of measuring the potential across the membrane of single cells, a problem which calls for an extremely high impedance, low grid current input circuit. He describes a cathode follower circuit and a negative capacitance circuit and analyzes and tests them with respect to neutralization of input capacitance, noise and grid current. The result is an excellent description of the theoretical and practical considerations that led to the design of an important and useful circuit.

Stabilized Wide-Band Potentiometric Preamplifiers (Moore and Gebhart, p. 1928)—The second paper of the three in this issue dealing with wide-band high impedance amplifiers, in addition to analyzing and describing several suitable circuits and providing an excellent discussion of previous work in the field, introduces the use of a chopper to stabilize this type of amplifier against drift. Also of special interest is the authors' use of a simple analog computer as a design aid in simplifying the analysis.

Bandwidth Limits for Neutralized Input Capacity Amplifier (Schoenfeld, p. 1942)—The final paper of the series uses the root-locus technique for studying the behavior of different input capacity neutralization circuits. The analysis yields design criteria for maximizing the performance of this class of amplifier and provides an excellent theoretical basis for evaluating future neutralization schemes. It also provides an excellent theoretical complement to the preceding two papers.

On the Reception of Quasi-Monochromatic, Partially Polarized Radio Waves (Ko, p. 1950)—The response of a receiving antenna to a completely (elliptically) polarized radio wave has been thoroughly discussed in the literature. However, electromagnetic waves caused by natural radiation are only partially polarized, that is, the electric field vector traces out an ellipse whose shape and orientation, instead of remaining constant, is continuously changing. Such waves arise, for example, in radio astronomy, microwave plasma diagnostics, passive radar mapping, and have long been familiar to physicists in the optics field, a field which is now rapidly becoming important to radio engineers. Thus this paper makes a noteworthy contribution both to antenna theory and practice, and, significantly, does so in terms that are compatible with modern optical theory.

Coherent FDM/FM Telephone Communication (Develet, p. 1957)—With an active communication satellite, Telstar, already in orbit, little further need be said about the timeliness of a paper which deals with the design of telephone communication systems which utilize satellite repeaters. The principal results of the paper, namely, establishing the bounds on the performance to be expected when employing coherent reception in FDM/FM satellite communication, are of broad applicability and will be widely useful, especially since the author has adhered to the internationally-approved CCIR definitions and standards in deriving his results.

A Piezoelectric-Piezomagnetic Gyrator (Onoe and Sawabe, p. 1967)—A new method has been found for realizing that somewhat elusive and intriguing circuit element, the gyrator, by means of a clever electromechanical system. This element is put to good use to produce a unilateral passive network element which, unlike previous gyrator-type isolators, maintains a high forward-to-backward ratio at all frequencies.

Scanning the Transactions appears on page 2005.

Letter Accompanying Ballot for IEEE Directors

August 14, 1962

AMERICAN INSTITUTE OF
ELECTRICAL ENGINEERS
345 East 47 Street
New York 17, N. Y.

and

THE INSTITUTE OF RADIO
ENGINEERS, INCORPORATED
1 East 79 Street
New York 21, N. Y.

To the Voting Members of the AIEE and IRE:

We are happy to report that the Voting Members of our two Institutes have ratified the Agreement of Merger (as published in Part II of the April, 1962, issue of the PROCEEDINGS OF THE IRE, and in Section II of the April, 1962, issue of ELECTRICAL ENGINEERING, by a vote of 29,464 *aye*, 4383 *no* in AIEE, and 36,221 *aye*, 5489 *no* in IRE. This agreement provides, in paragraph 3, that the IRE and AIEE Boards of Directors shall nominate a slate of 25 Directors for the continuing organization (to be named IEEE), one of whom is to be designated President and one designated a Vice President, and that the Voting Members of the AIEE and IRE shall vote to accept or reject the slate as a whole. The enclosed ballot contains this slate. If you are a Voting Member of both the IRE and AIEE, you will receive two ballots and are entitled to vote both. Please mark the ballot for or against the slate as a whole, insert it in the plain envelope, seal the plain envelope and insert it in the mailing envelope enclosed, and write your signature in the place indicated on the mailing envelope.

In order to be counted, your ballot must be received at the Headquarters designated on the mailing envelope before 12:00 noon on October 1, 1962.

For your information, we enclose biographies of each of the 25 nominees. If elected, they will serve, as provided in the Agreement of Merger, until the next (1964) Annual Assembly, that is, for a term of approximately one year.

Also provided for your information is a statement on the proposed future organization of the Board of Directors of the continuing organization.

Please mark your ballot "for" or "against" the slate as a whole and return it promptly.

Very sincerely,

B. Richard Teare, Jr.
President, AIEE

Patrick E. Haggerty
President, IRE

Ballot

I hereby vote

FOR ☐ AGAINST ☐

the following persons to be Directors of the continuing Corporation after the merger of the American Institute of Electrical Engineers and The Institute of Radio Engineers, Incorporated:

President—ERNST WEBER (AIEE Fellow '34; IRE Fellow '51)

Vice President—B. RICHARD TEARE, Jr. (AIEE Fellow '42; IRE Fellow '51)

Director—LLOYD V. BERKNER (IRE Fellow '47; AIEE Fellow '47)

Director—HENDLEY BLACKMON (AIEE Fellow '49; IRE Senior Member '57)

Director—WARREN H. CHASE (AIEE Fellow '51; IRE Senior Member '51)

Director—W. RUSSELL CLARK (AIEE Fellow '61; IRE Senior Member '47)

Director—JOHN W. DAVIS (AIEE Member '55)

Director—JOSEPH H. ENENBACH (AIEE Member '61)

Director—PATRICK E. HAGGERTY (IRE Fellow '58; AIEE Member '62)

Director—FERDINAND HAMBURGER, Jr. (AIEE Fellow '48; IRE Fellow '53)

Director—JOHN T. HENDERSON (IRE Fellow '51; AIEE Member '57)

Director—HERBERT O. HODSON (AIEE Fellow '62)

Director—LYNN C. HOLMES (IRE Fellow '49; AIEE Fellow '51)

Director—TITUS G. LeCLAIR (AIEE Fellow '40)

Director—CLARENCE H. LINDER (AIEE Fellow '57; IRE Senior Member '62)

Director—J. ELIOT McCORMACK (AIEE Fellow '44)

Director—RONALD L. McFARLAN (IRE Fellow '61; AIEE Member '62)

Director—DANIEL E. NOBLE (IRE Fellow '47)

Director—BERNARD M. OLIVER (IRE Fellow '54)

Director—WALTER E. PETERSON (IRE Senior Member '50; AIEE Associate Member '46)

Director—HARADEN PRATT (IRE Fellow '29; AIEE Fellow '37)

Director—JOHN D. RYDER (AIEE Fellow '51; IRE Fellow '52)

Director—WILLIAM G. SHEPHERD (IRE Fellow '52; AIEE Member '51)

Director—EUGENE C. STARR (AIEE Fellow '49)

Director—F. KARL WILLENBROCK (IRE Senior Member '61)

Note: Nominees for Directors are listed alphabetically, with date of highest grade of membership.

Future Organization of the Board of Directors of the Continuing Corporation

THE AGREEMENT OF MERGER and its Exhibits provide for the appointment of Officers and the designation of Directors to particular duties. The Resolution of Merger provides for a 14-Man Merger Committee to "prepare for and take all necessary steps to implement the merger." Pursuant to this Resolution and in the event that the slate of Directors presented herewith is approved by the memberships of the two Institutes, it is the intention of the Merger Committee to suggest to the elected Board, the following designations:

1) Vice President Teare as representing the Sections Committee
2) Director Oliver as a Vice President representing the Professional Technical Groups Committee
3) Director Pratt as Secretary
4) Director Clark as Treasurer
5) Director Ryder as Editor
6) Director Blackmon as representing the Technical Operating Committee
7) Director Berkner (Senior Past President IRE)
8) Director Chase (Junior Past President AIEE)
9) Director Haggerty (Junior Past President IRE)
10) Director Linder (Senior Past President AIEE).

It is further the intention of the Merger Committee to suggest to the Board the establishment of nine geographical Regions. The preliminary boundaries, which are subject to adjustment, of seven Regions in North America, including Hawaii and Alaska, are indicated in the accompanying map. The Committee, using the best information available to it, established these tentative boundaries for the purpose of indicating the general areas of the Regions so as to have a guide for selecting candidates for Regional Directors. The final boundaries will not be established until after recommendations have been submitted after consideration by all member groups directly interested. The Committee will suggest the establishment of a Region 8, to cover the membership generally in Europe, the Near East, and North Africa, including existing Sections in Benelux, Egypt, France, Israel, Italy, Geneva, and the United Kingdom, and the establishment of a Region 9 to comprise those areas of the World not covered in Regions 1 thru 8. The Committee will suggest the designation of the following elected Directors as Regional Directors:

Region 1—Director Holmes
Region 2—Director Hamburger
Region 3—Director Davis
Region 4—Director Enenbach
Region 5—Director Hodson
Region 6—Director Noble
Region 7—Director Henderson
Region 9—Director McFarlan

Note: The Director for the proposed Region 8 will be selected in accordance with the Consitution and By-laws of the continuing Corporation.

The Committee will further suggest the creation of the Honorary title of Director Emeritus and will suggest that this title be conferred on Dr. Alfred N. Goldsmith and Mr. Elgin B. Robertson. Their biographies are provided for the information of the Voting Members.

The members of the 14-Man Merger Committee are:

Warren H. Chase, Co-chairman for AIEE (AIEE F'51; IRE SM'51)
Junior Past President AIEE; Vice President, Ohio Bell Telephone Co., Cleveland Ohio.

Patrick E. Haggerty, Co-chairman for IRE (IRE F'58; AIEE M'62)
President IRE; President, Texas Instruments Incorporated, Dallas, Tex.

Lloyd V. Berkner (IRE F'47; AIEE F'47)
Junior Past President IRE; President, Graduate Research Center of the Southwest, Dallas, Tex.

Hendley Blackmon (AIEE F'49; IRE SM'57)
Director AIEE; Engineering Manager, Association Activities, Centra Laboratories, Westinghouse Electric Corp., Pittsburgh, Pa.

W. Russell Clark (AIEE F'61; IRE SM'47)
Treasurer AIEE; Assistant to the Vice President, Technical Affairs, Leeds and Northrup Company, Philadelphia, Pa.

John T. Henderson (IRE F'51; AIEE M'57)
Past President IRE; Principal Research Officer, National Research Council, Ottawa, Canada.

Clarence H. Linder (AIEE F'57; IRE SM'62)
Senior Past President AIEE; Vice President, General Electric Company, New York, N. Y.

Ronald L. McFarlan (IRE F'61; AIEE M'62)
Senior Past President IRE; Consultant Chestnut Hill, Mass.

Walter E. Peterson (IRE SM'50; AIEE AM'46)
Past Chairman IRE Los Angeles Section; President, Automation Development Corporation, Culver City, Calif.

Haradan Pratt (IRE F'29; AIEE F'37)
Secretary IRE; Consultant, Pompano Beach, Fla.

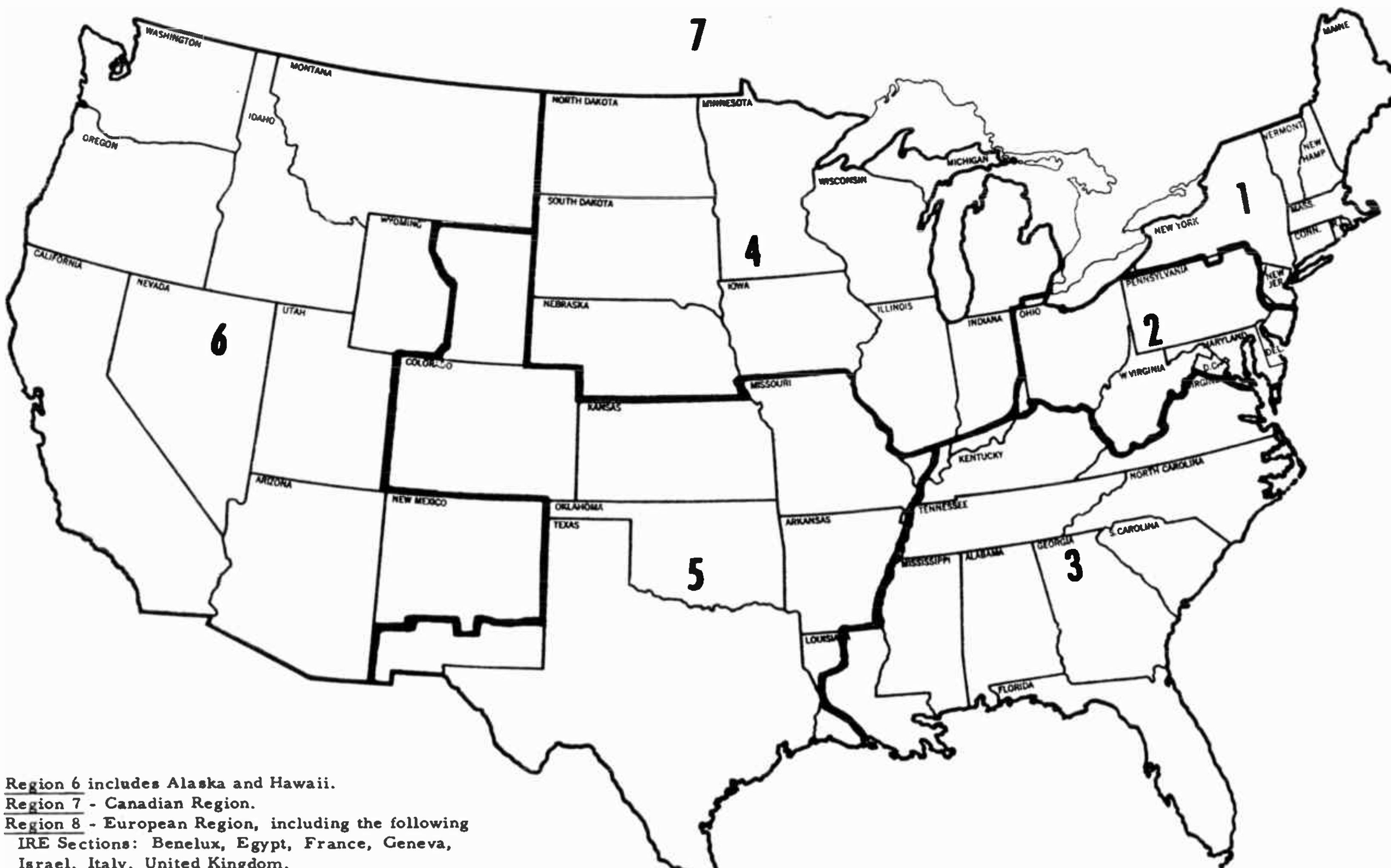
Elgin B. Robertson (AIEE F'45)
Past President AIEE; President, Elgin B. Robertson, Inc., Dallas, Tex.

Lawrence M. Robertson (AIEE F'45)
Past Vice President AIEE; Engineering Manager, Public Service Co. of Colorado, Denver, Colo.

John D. Ryder (IRE F'52, AIEE F'51)
Past President IRE; Dean, College of Engineering, Michigan State University, East Lansing, Mich.

B. Richard Teare, Jr. (AIEE F'42; IRE F'51)
President AIEE; Dean, College of Engineering and Science, Carnegie Institute of Technology, Pittsburgh, Pa.

TENTATIVE REGIONAL BOUNDARIES OF IEEE (Subject to Adjustment)



Region 6 includes Alaska and Hawaii.
Region 7 - Canadian Region.
Region 8 - European Region, including the following IRE Sections: Benelux, Egypt, France, Geneva, Israel, Italy, United Kingdom.
Region 9 - The rest of the world, including the following IRE Sections: Buenos Aires, Chile, Colombia, India, Rio de Janeiro, Tokyo.

Biographies of Nominees

(In order as listed on the ballot.)

ERNST WEBER—President, Polytechnic Institute of Brooklyn, Brooklyn, N. Y. Born 1901. Ph.D. (1926) University of Vienna; D.Sc. (1927) Technical University, Vienna. President IRE, 1959; Vice President IRE, 1962; IRE Director, 1952–62. AIEE Committees, 1937 to date; IRE Committees, 1941 to date. AIEE Fellow (1934) citation: "For investigations and publications on electron flow, field transients, nonlinear circuits, and electromagnetic units." IRE Fellow (1951) citation: "For his outstanding service and achievement in the field of engineering education and for his many contributions to electromagnetic theory." Recipient, Electrical Engineering Education Medal, 1960

B. RICHARD TEARE, JR.—Dean, College of Engineering and Science, Carnegie Institute of Technology, Pittsburgh, Pa. Born 1907. D. Eng. (1937) Yale University. President AIEE, 1962; Vice President AIEE, 1957–59; AIEE Director, 1957–59, 1961 to date; AIEE Committees, 1934 to date; IRE Committee, 1944. AIEE Fellow (1942) citation: "For research and development of control and conversion devices and contributions to electrical engineering education." IRE Fellow (1951) citation: "For contributions in teaching and research in the fields of engineering methods, network theory, and servomechanisms." Recipient, George Westinghouse Award, 1947.

LLOYD V. BERKNER—President, Graduate Research Center of the Southwest, Dallas, Tex. Born 1905. B.S.E.E. (1927) University of Minnesota. D.Sc. (hon.) (1955) Polytechnic Institute of Brooklyn and honorary doctorate from Uppsala, Sweden; Calcutta, India; U. of Notre Dame; Columbia University; University of Edinburgh, Scotland, IRE President, 1961; IRE Director, 1959–62; IRE Committees, 1937 to date; AIEE Committees, 1947–51, 1958–59. IRE Fellow (1947) citation: "For his investigation of ionospheric phenomena and his contribution to airborne radar development." AIEE Fellow (1947) citation: "For pioneering exploration of the properties of the ionosphere." Recipient, Gold Medal of Congress.

HENDLEY BLACKMON—Engineering Manager, Association Activities, Central Laboratories, Westinghouse Electric Corp., Pittsburgh, Pa. Born 1903. B.Sc. (1925) Georgia Institute of Technology. AIEE Director, 1959–62. AIEE Committees, 1949 to date; IRE Committees, 1943. AIEE Fellow (1949) citation: "For special proficiency in contributions to electrical literature." Recipient, Westinghouse Order of Merit.

WARREN H. CHASE—Vice President, Ohio Bell Telephone Co., Cleveland, Ohio. Born 1898. S.B. (1924) Harvard University School of Engineering. President AIEE, 1961–62; Vice President AIEE, 1958–61; AIEE Director, 1958–62; AIEE Committees, 1954 to date. AIEE Fellow (1951) citation: "For contributions to the engineering of a large telephone system." Recipient, Outstanding Engineer of the Year Award, Cleveland Society of Professional Engineers, 1960.

W. RUSSELL CLARK—Assistant to the Vice President, Technical Affairs, Leeds and Northrup Co., Philadelphia, Pa. Born 1907. D.Sc. (1938) University of Pennsylvania. Treasurer AIEE, 1960–62; AIEE Director, 1956–62; AIEE Committees, 1944 to date. AIEE Fellow (1961) citation: "For contributions by invention, development and design to recording, measuring and control instruments and systems."

JOHN W. DAVIS—Assistant Vice President, Southern Bell Telephone and Telegraph Co., Atlanta, Ga. Born 1898. B.S. in physics (cum laude) (1921), Georgetown College. D.Eng. (hon.) (1960) Clemson College. AIEE Vice President, 1960–61; AIEE Director-at-large, 1961–62; AIEE Committees, 1960–62.

JOSEPH H. ENENBACH—District Marketing Manager (Data), Illinois Bell Telephone Co., Chicago, Ill. Born 1923. B.S.E.E. Illinois Institute of Technology. Chairman AIEE Chicago Section. Past President National Electronics Conference. AIEE Committees, 1958–62.

PATRICK E. HAGGERTY—President, Texas Instruments Incorporated, Dallas, Tex. Born 1914. B.S.E.E. (1936) Marquette University; LL.D. (hon.) (1959) St. Mary's University; LL.D. (hon.) (1960) Marquette University. President IRE, 1962. Director IRE, 1960–62; IRE Committees, 1954 to date. IRE Fellow (1958) citation: "For leadership in the advancement of semiconductor devices."

FERDINAND HAMBURGER, JR.—Professor and Chairman, Electrical Engineering Department and Director, Radiation Laboratory, The Johns Hopkins University, Baltimore, Md. Born 1904. D. Eng. (1931) The Johns Hopkins University. IRE Editor, 1960–61; IRE Director, 1950–51, 1959–61; AIEE Committees, 1933–39, 1951–62; IRE Committees, 1951–61. AIEE Fellow (1948) citation: "For contributions in the fields of electrical insulation and radio engineering; for consulting and outstanding leadership in engineering education." IRE Fellow (1953) citation: "For his leadership as a teacher and author in the electronics and electrical engineering fields."

JOHN T. HENDERSON—Principal Research Officer, National Research Council, Ottawa, Canada. Born 1905. Ph.D. (1932) University of London. President IRE, 1957; IRE Director, 1953–59. IRE Fellow (1951) citation: "For his contributions in the field of radio direction-finding systems and, in particular, for his work in the development of Canadian radar during the war." Member of the Order of the British Empire, 1943.

HERBERT O. HODSON—Vice President in charge of engineering and power, Southwestern Public Service Co., Amarillo, Tex. Born 1909. B.S.E.E. (1930) South Dakota School of Mines and Technology. AIEE Committees, 1951–59. AIEE Fellow (1962) citation: "For contributions to the engineering, construction and operations of a rapidly growing electric utility system." Recipient, Outstanding Engineer of the Year Award, Texas Society of Professional Engineers, 1953.

LYNN C. HOLMES—Executive Assistant, Research and Engineering, General Dynamics/Electronics, Rochester, N. Y. Born 1904. M.E.E. (1932) Rensselaer Polytechnic Institute. Vice President AIEE, 1958–60; AIEE Director, 1958–62; AIEE Committees, 1947 to date. IRE Fellow (1949) citation: "For his contributions to theory and practice in the field of magnetic recording." AIEE Fellow (1951) citation: "For important contributions to the development of sound recording."

TITUS G. LECLAIR—Reactor Applications Manager, General Atomic Division, General Dynamics Corp., San Diego, Calif. Born 1899. B.S.E.E. (1921) D.Sc. (hon.) (1951) University of Idaho. President AIEE, 1950–51; Vice President AIEE, 1946–48; AIEE Director, 1941–51. AIEE Fellow (1940) citation: "For original contributions to relaying and protective devices and the design of high voltage transmission systems."

CLARENCE H. LINDER—Vice President, General Electric Co., New York, N. Y. Born 1903. M.S.E.E. (1927) University of Texas; D. Eng. (hon.) (1955) Worcester Polytechnic Institute; D. Eng. (hon.) (1956) Clarkson College of Technology. President AIEE, 1960–61; Treasurer AIEE, 1958–60; AIEE Director, 1956–61; AIEE Committees, 1956 to date. AIEE Fellow (1957) citation: "For his contributions in coordinating electrical development and cultivating engineering talent."

J. ELIOT MCCORMACK—Vice President, Consolidated Edison Co. of New York, New York, N. Y. Born 1904. E.E. (1926) Polytechnic Institute of Brooklyn. AIEE Committees, 1934–38, 1944–48, 1959–62. AIEE Fellow (1944) citation: "For engineering designs and innovations in connection with large power supplies."

RONALD L. MCFARLAN—Consultant, Chestnut Hill, Mass. Born 1905. Ph.D. (1930) University of Chicago. President IRE, 1960;

IRE Director, 1957–62; IRE Committees, 1958–62. IRE Fellow (1961) citation: "For contributions to systems applications of electronic computers and for effective administrative activities during formative periods of technical advancement."

DANIEL E. NOBLE—Director and Executive Vice President, Communications, Semiconductor, Solid State Systems and Military Electronics Divisions of Motorola, Inc., Phoenix, Ariz. Born 1901. B.S. (1929) University of Connecticut; D.Sc. (hon.) (1957) Arizona State College. IRE Director, 1957–62; IRE Committees, 1940 to date. IRE Fellow (1947) citation: "In recognition of his contributions to the design and application of very-high-frequency voice communication systems for police and other emergency services."

BERNARD M. OLIVER—Vice President, Research and Development, Hewlett-Packard Co., Palo Alto, Calif. Born 1916. Ph.D. (1940) California Institute of Technology. IRE Director, 1959–61; IRE Committee, 1954–61. IRE Fellow (1954) citation: "For his contributions to communications, particularly in the field of information theory and coding systems."

WALTER E. PETERSON—President, Automation Development Corp., Culver City, Calif. Born 1921. B.S.E.E. (1943) University of California. Chairman, IRE Los Angeles Section, 1955–56. Chairman WESCON Board of Directors, 1961. AIEE Committee, 1955.

HARADEN PRATT—Consultant, Pompano Beach, Fla. Born 1891. B.S. (1914) University of California. IRE President, 1938; IRE Treasurer, 1941–42; IRE Secretary, 1943–62; IRE Director, 1935–62. IRE Fellow (1929) citation: "In recognition of his engineering contributions to the development of radio, of his work in the extension of communications facilities to distant lands, and of his constructive leadership in Institute affairs." AIEE Fellow (1937) citation: "For contributions to the extension of international radio communication and to the standardization of radio engineering systems and devices." Recipient, IRE Medal of Honor (1944); IRE Founders Award (1960).

JOHN D. RYDER—Dean, College of Engineering, Michigan State University, East Lansing, Mich. Born 1907. Ph.D. (1944) Iowa State University. IRE President, 1955. IRE Editor, 1958–59; IRE Director, 1952–59; IRE Committees, 1945–46, 1948 to date; AIEE Committees, 1945–60. AIEE Fellow (1951) citation: "For inventions in the field of instrumentation and control, and contributions to electrical engineering literature and education." IRE Fellow (1952) citation: "For his contributions in industrial applications of electronics circuits and to education in radio and allied fields."

WILLIAM G. SHEPHERD—Professor and Head of Electrical Engineering, University of Minnesota, Minneapolis, Minn. Born 1911. Ph.D. (1937) University of Minnesota. IRE Director, 1960–62; IRE Committees, 1951–60. IRE Fellow (1925) citation: "For his contributions to the development and design of electron tubes, particularly the reflex klystron."

EUGENE C. STARR—Consultant, Bonneville Power Administration, Portland, Ore. Born 1901. E.E. (1938) Oregon State College. AIEE Director, 1958–62; AIEE Committees, 1941–62. AIEE Fellow (1949) citation: "For research in high voltage phenomena and contributions to electrical engineering education."

F. KARL WILLENBROCK—Associate Dean and Director of Laboratories, Division of Engineering and Applied Science, Harvard University, Cambridge, Mass. Born 1920. Ph.D. (1950) (Applied Physics) Harvard University. IRE Director, 1962; Chairman, IRE Boston Section, 1959–60.

Note. The following are biographies of two members not listed on the enclosed ballot. The Merger Committee will suggest to the elected Board that they each be designated "Director Emeritus."

ALFRED N. GOLDSMITH—Consulting Engineer, New York, N. Y. (IRE Founder and Charter Member). Born 1889. Ph.D. (1911) Columbia University. Sc.D. (hon.) (1935) Lawrence College. IRE President, 1928; IRE Secretary, 1918–27 ;IRE Editor, 1912–28, and 1930–53; IRE Editor Emeritus, 1954 to date; IRE Director, 1912–62; IRE Committees, 1913–53. IRE Fellow (1915) citation: "For his contributions to radio research, engineering and commercial development, his leadership in standardization, and his unceasing devotion to the establishment and up-building of the Institute and its PROCEEDINGS." AIEE Fellow (1920) citation: "For pioneering in the development of radio communications." Recipient IRE Medal of Honor, 1941; IRE Founders Award, 1954.

ELGIN B. ROBERTSON—President, Elgin B. Robertson, Inc., Dallas, Tex. Born 1893. E.E. (1915) University of Texas. D. Eng. (hon.) (1954) Southern Methodist University. AIEE President, 1953–54; AIEE Director, 1947–56; AIEE Committees, 1944–61. AIEE Fellow (1945) citation: "For original designs of outdoor substation structures, important developments in transformers and switchgear, and outstanding service to the War Production Board." Honorary Member, AIEE, 1959.

Quantum Effects in Communications Systems*

J. P. GORDON†

Summary—**The information capacity of various communications systems is considered. Quantum effects are taken fully into account. The entropy of an electromagnetic wave having the quantum statistical properties of white noise in a single transmission mode is found, and from it the information efficiency of various possible systems may be derived. The receiving systems considered include amplifiers, heterodyne and homodyne converters and quantum counters. In the limit of high signal or noise power (compared to $h\nu B$, where h is Planck's constant and ν and B are, respectively, the center frequency and bandwidth of the channel) the information efficiency of an amplifier can approach unity. In the limit of low powers the amplifier becomes inefficient, while the efficiency of the quantum counter can approach unity. The amount of information that can be incorporated in a wave drops off rather rapidly when the power drops below $h\nu B$.**

I. Introduction

With the advent of the possibility of broad-band communications at frequencies in the infrared and optical range, it has become important to investigate the effects of the quantization of radiation on the capacity of electromagnetic waves to transmit information. Unlike the situation prevailing in the microwave range, where thermal noise generally provides an ultimate limit to our ability to transmit information, in the infrared and optical range this limit is provided by what may be called quantum noise.

Our work stems principally from the classic work of Shannon[1] on discrete and continuous information channels. Gabor[2,3] introduced the concept of quantization into electromagnetic communication channels and coined the term "quantum noise." In consideration of the problem of field measurements by a receiver, he used an electron beam probe. The shot noise in the beam influenced his results in an important and, in the light of present knowledge, unnecessary way. Stern[4,5] has considered information rates in "photon channels." His conclusion[5] that the information efficiency of a linear amplifier can be no greater than 50 per cent conflicts with the results presented here. The major difference may be traced to the fact that he takes no account of the information that may be stored in the signal phase; and phase information approaches 50 per cent of the total possible information in the large signal-to-noise case where the quantum theory and the classical theory approach one another. Lasher[6,7] has also obtained expressions for information capacity based on quantum mechanical principles. His results agree qualitatively with ours; the quantitative differences presumably arise from the approximate methods which he used. We[8] have previously discussed some of the ideas which are utilized in this paper. In other recent work the important question of the statistical properties of quantum noise in linear amplifiers has been studied.[9,10,11]

Our ruminations will be limited to waves existing in a transmission system for which only a single transmission mode of the field is utilized. That is, the polarization and distribution of the field over any plane perpendicular to the direction of propagation are considered invariant. This situation is typical of transmisssion in a coaxial line or in a waveguide. It will also very likely be true for long-distance broad-band optical communication systems. A possible departure from such a single-mode system would involve the use of the two orthogonal field polarizations to provide two independent channels.

During the course of passage from transmitter to receiver, the signal is presumed to suffer a large attenuation and, in general, to be supplemented by some amount of additive white[12] noise power. At the receiver

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† Bell Telephone Laboratories, Murray Hill, N. J.

[1] C. E. Shannon and W. Weaver, "The Mathematical Theory of Communication," University of Illinois Press, Urbana, Ill.; 1949.

[2] D. Gabor, "Communication theory and physics," *Phil. Mag.*, vol. 41, pp. 1161–1187; 1950.

[3] D. Gabor, "Lectures on Communication Theory," Res. Lab. of Electronics, M.I.T., Cambridge, Mass., Tech. Rept. No. 238; April 3, 1952.

[4] T. E. Stern, "Some quantum effects in information channels," IRE Trans. on Information Theory, vol. IT-6, pp. 435–440; September, 1960.

[5] T. E. Stern, "Information rates in photon channels and photon amplifiers," 1960 IRE International Convention Record, pt. 4, pp. 182–188.

[6] G. J. Lasher, "A quantum statistical treatment of the channel capacity problem of information theory," in "Advances in Quantum Electronics," J. R. Singer, Ed., Columbia University Press, New York, N. Y., pp. 520–536; 1961.

[7] G. J. Lasher, "Channel capacity of optical frequencies," presented at the NATO-SADTC Symp. on Technical and Military Applications of Laser Techniques, The Hague, Netherlands; April, 1962.

[8] J. P. Gordon, "Information capacity of a communications channel in the presence of quantum effects," in "Advances in Quantum Electronics," J. R. Singer, Ed., Columbia University Press, New York, N. Y., pp. 509–519; 1961.

[9] W. H. Wells, "Quantum formalism adapted to radiation in a coherent field," *Ann. Phys.* (*N. Y.*), vol. 12, pp. 1–40; January, 1961.

[10] J. P. Gordon, W. H. Louisell, and L. R. Walier, "Quantum fluctuations and noise in parametric processes. II," to be published.

[11] J. P. Gordon, W. H. Louisell, and L. R. Walker, "Quantum statistics of maser amplifiers and attenuators," to be published.

[12] Since we are concerned with a very broad range of frequencies, neither thermal noise nor quantum noise is truly "white," as this would imply a uniform spectral density. Rather, the noise is "colored"; its spectral density is generally a function of frequency. Since, however, the systems we consider are all narrow band, in the sense that the bandwidth is always much smaller than the carrier frequency, the noise may be considered to be white within the bandwidth. Cases in which the spectral density of the noise varies appreciably across the band may be treated by dividing the band up into smaller segments, and treating each such segment as an independent channel.

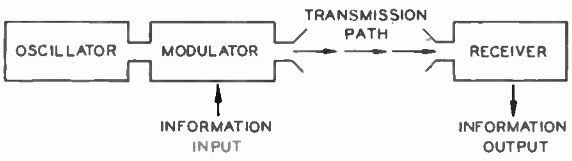

Fig. 1—Typical communication system.

as much as possible of the information remaining in the received wave is extracted. The receiver may incorporate an amplifier at the carrier frequency or it may not. We will investigate both of these cases. Fig. 1 shows a typical communications channel such as we have described.

So long as the electromagnetic waves may be described classically, *i.e.*, without quantization, Shannon[1] has shown that the information capacity C of a signal of average power S in the presence of additive white noise power N in a channel of bandwidth B is given by

$$C = B \log\left(1 + \frac{S}{N}\right) \tag{1}$$

If the logarithm is taken to the base 2, C is in units of bits per second. To realize this capacity the signal must be modulated in such a way as to have also the statistical randomness of white noise.

In deriving (1) Shannon noted that information and entropy were closely allied quantities. In fact he identified information as *prescribed* entropy. He was able to show that the entropy rate R of a continuous wave, having the statistical properties of narrow-band white noise and average power P, could be expressed as

$$R = B \log\left(\frac{P}{P_0}\right) \tag{2}$$

where the constant P_0 is arbitrary. To obtain (1) he subtracted the entropy rate for the noise alone from the entropy rate for the combined signal and noise. The latter also has the statistical properties of white noise when both signal and noise have these properties independently. Thus the constant is cancelled out and

$$C = B \log\left(\frac{S+N}{P_0}\right) - B \log\left(\frac{N}{P_0}\right) = B \log\left(1 + \frac{S}{N}\right).$$

C is the additional entropy occasioned by the presence of the signal. Since the signal is completely prescribed, the added entropy is prescribed entropy, or information.

Eq. (1) says that the information capacity approaches infinity as the signal-to-noise ratio approaches infinity. This is because as the noise decreases we can make more and more accurate measurements of the state of the signal field. However, the uncertainty principle of quantum mechanics tells us that in fact we cannot measure a field to arbitrary accuracy, and so as $N \to 0$, fundamental quantum limitations on information capacity make their appearance.

II. Entropy of White Noise

The fact that an electromagnetic wave is quantized allows us to obtain an absolute value for its entropy without the arbitrary constant of (2). Consider the wave in a transmission line traveling toward the receiver. Assume that the wave velocity is v and that there is no dispersion. Then, in time t, the receiver measures the field which had previously occupied a length $L = vt$ of the line. To describe this field we can expand it into a series of orthogonal modes, and then measure the state of excitation of each mode as well as possible. A commonly used expansion is a spatial Fourier series. For this expansion the qth mode varies with distance and time according to the exponential factor

$$\exp\left[jq\frac{2\pi}{L}(z - vt)\right].$$

The condition for orthogonality of the modes is that the different values of q differ by integers. It is also clear from the above expression that the mode q has frequency qv/L. Thus the frequency separation between adjacent modes is $\Delta\nu = v/L$. In a bandwidth B there are $B/\Delta\nu = BL/v$ orthogonal modes. Since $L = vt$ we see that in time t the receiver measures the state of excitation of Bt such modes. The rate of arrival of independent field modes at the receiver is therefore B.

The complete description of the field requires measurement of the state of excitation of each mode. Classically this would involve independent simultaneous measurements of the amplitude and phase of each mode, or equivalently simultaneous measurement of the electric and magnetic fields associated with each mode. Thus, classically, we make $2B$ independent measurements per second to identify the wave. In quantum mechanics the measurements of electric and magnetic fields are not independent, so we must consider that we make only B independent measurements per second, each measurement specifying the state of one particular field mode.

Now we know that a white noise wave must have the most random possible excitation of the various modes consistent with the average power in the wave. This allows us to calculate the entropy of such a wave. Let us specify the state of each mode by assigning to it exactly m photons, *i.e.*, an excitation energy $mh\nu$. From statistical mechanics[13] we know that the entropy per mode for a large number of modes is given by the expression

$$H = -\sum_m p(m) \log p(m)$$

where $p(m)$ is the probability that a mode will contain just m photons. The average energy per mode is given by

$$\overline{E} = h\nu\overline{m} = h\nu \sum_m m p(m)$$

and of course since $p(m)$ is a probability, the $p(m)$'s must fulfill the requirement that

$$\sum_m p(m) = 1.$$

To find the most random possible excitation consistent with a given average power, we must maximize H by varying the probabilities $p(m)$ while keeping $\Sigma p(m)$ and $\Sigma m p(m)$ constant. This is a simple problem in the calculus of variations. The set of $p(m)$ which maximize H are

$$p(m) = \frac{1}{1+\overline{m}}\left(\frac{\overline{m}}{1+\overline{m}}\right)^m.$$

The average power P in this wave is

$$P = EB = \overline{m}h\nu B$$

since $E=\overline{m}h\nu$ is the average energy per mode and B modes per second are incident on the receiver. This exponential probability distribution for the excitation of the modes is consistent with the exponential power distribution which we know is characteristic of white noise. The entropy per mode for white noise is thus

$$\begin{aligned} H &= -\sum p(m) \log p(m) \\ &= \sum p(m)\left[\log(1+\overline{m}) + m \log\left(\frac{1+\overline{m}}{\overline{m}}\right)\right] \\ &= \log(1+\overline{m}) + \overline{m}\log\left(1+\frac{1}{\overline{m}}\right). \end{aligned} \qquad (3)$$

Since $\overline{m} = P/h\nu B$ where P is the average power in the wave, we may express the entropy per mode as

$$H = \log\left(1+\frac{P}{h\nu B}\right) + \frac{P}{h\nu B}\log\left(1+\frac{h\nu B}{P}\right).$$

One may object that the specification of the excitation of each mode in terms of exact numbers of photons is not the only possible way. However, the number of distinguishable excitations within an energy range from E to $E+\Delta E$ should be independent of the quantities used for the field specification, and so we are free to choose the most convenient specification, as we have done. Finally we note that the rate of arrival of entropy at the receiver for a white noise wave is

$$R = HB = B\log\left(1+\frac{P}{h\nu B}\right) + \frac{P}{h\nu}\log\left(1+\frac{h\nu B}{P}\right). \qquad (4)$$

Eq. (4) is the quantum equivalent of (2).

Of the terms in (4) the first has a form quite similar to the classical expression and predominates when the average number of photons per mode is large compared to unity. We can call it the mode entropy. It is equal to the rate of arrival of modes, B, times the logarithm of $\overline{m}+1$, which may be thought of rather loosely as the number of frequently occurring mode occupation numbers in a typical noise wave. By mode occupation number we mean the number of photons in the mode.

The second term of (4) is of fundamental quantum origin. It is the predominant term at power levels less than $h\nu B$ where the mean occupation number $\overline{m}$ becomes less than unity. We can call it the photon entropy. It is equal to the rate of arrival of photons, $P/h\nu$, times the logarithm of the number of frequently occurring intervals (*i.e.*, modes) for each photon. We shall see that at least part of this entropy can take the form of information which is recoverable if we use a photocell or some other energy-sensitive device as a receiver.

If we approach classical theory by the frequently used artifice of supposing that h becomes very small, it may be seen that (4) approaches (2) with the arbitrary constant evaluated as

$$P_0 = h\nu B/e$$

where e is the Naperian base for natural logarithms. Since the arbitrary constant contains h, it is clear that it could not be determined from a classical description.

III. Entropy and Information

In Section II we found an absolute expression for the entropy of white noise, utilizing a particular quantum mechanical description of the possible excitations of the field modes. It is not obvious, however, that all of this entropy can be prescribed as a signal, and so constitute information. This is not to say that we cannot modulate a CW carrier wave in such a way as to give the resulting wave the statistical properties of white noise in the prescribed bandwidth B, but rather that there is very likely some part of the resulting entropy which is essentially irretrievable as information. We must confess that we do not know at present the answer to this problem. In any event the entropy of the wave is certainly an upper limit to the amount of information it may contain, and as such it is a useful quantity.

[13] R. C. Tolman, "The principles of statistical mechanics," Oxford University Press, Oxford, England, 1938. See also Shannon and Weaver.[1]

IV. Information Capacity in the Presence of Additive Noise

Suppose we have a signal with average power S accompanied by additive white noise with average power N. Following the ideas of Shannon we note that the information in the wave can be no greater than the entropy of the combination of signal plus noise less the uninformative entropy of the noise alone. The entropy of the combined signal and noise is maximized when the total wave has the statistics of white noise. Quantum mechanically as well as classically, this implies that the signal alone should also have the characteristics of white noise. The entropy rate for the combined wave is then given by (4) with $P=S+N$, while the entropy rate for noise alone has $P=N$. The upper limit to the information in the wave, which we will label C_{wave}, for a signal of average power S in the presence of white noise of average power N is thus given by

$$C_{\text{wave}} = R_{(P=S+N)} - R_{(P=N)}$$

or

$$C_{\text{wave}} = B \log\left(1 + \frac{S}{N + h\nu B}\right)$$

$$+ \frac{S+N}{h\nu} \log\left(1 + \frac{h\nu B}{S+N}\right) - \frac{N}{h\nu} \log\left(1 + \frac{h\nu B}{N}\right). \quad (5)$$

For a bandwidth of 10^9 cps and an additive noise power N taken as arising from a black body at 290°K, *i.e.*,

$$N = h\nu B\left[\exp\left(\frac{h\nu}{290k}\right) - 1\right]^{-1},$$

the information limit, C_{wave}, is plotted in Fig. 2 as a function of frequency for power levels ranging from 10^{-7} to 10^{-13} watt.

A. Classical Limit

If the noise power N is considerably greater than $h\nu B$, we have a situation where a classical description of the wave should be adequate. Expansion of (5) to first order in the small quantities $h\nu B/N$ and $(h\nu B)/(S+N)$ yields

$$C_{\text{wave}} = B\left[\log\left(1 + \frac{S}{N}\right) - \frac{h\nu BS}{2N(S+N)} \log e \cdots\right].$$

Under the assumed condition $N \gg h\nu B$, the second term is always much smaller than the first, independent of the value of S/N, so the classical description which results in (2) is quite good.

If there is no additive noise, but the signal is much larger than $h\nu B$, we find

$$C_{\text{wave}} = B\left[\log\left(1 + \frac{S}{h\nu B}\right) + \log e + \cdots\right].$$

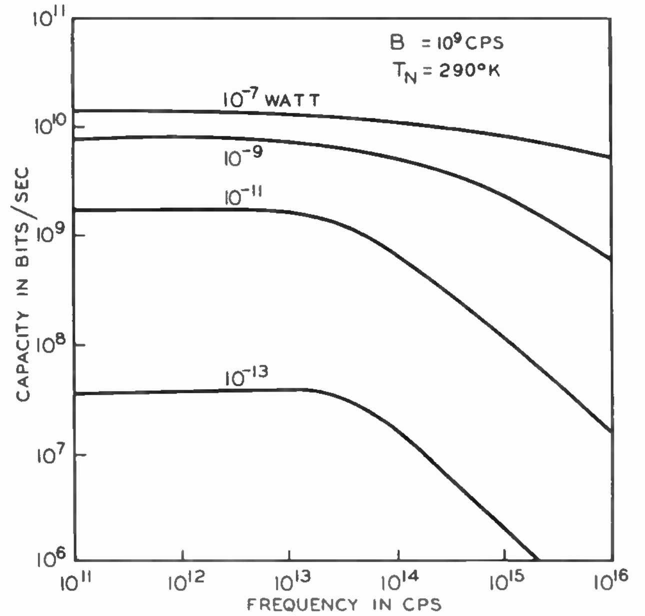

Fig. 2—Upper limit to the information that may be incorporated into an electromagnetic wave in a single transmission mode. Thermal noise, as originating from a black body at 290°K, is assumed to accompany the wave.

In the limit of very high signal power this expression is nearly the same as one would obtain from the classical expression (1), by assuming the presence of an equivalent "zero-point" noise power $h\nu B/e$. Note, however, that this equivalence is not exact.

V. Information Capacity After Transmission

As our transmitted signal travels toward the receiver, it is attenuated and usually some noise power is added to it. If we assume that the added noise is white then the information capacity of the received wave is limited by (5) where S is the received signal power and N the added noise power.

VI. Information Capacity After Coherent Amplification

Suppose now that the first element of the receiver is an amplifier at the carrier frequency. This could be a maser, a nondegenerate parametric amplifier or any other type of linear amplifier. Assume that the amplifier has high gain. There is always internal white noise generated in such an amplifier which, referred to the input, may be described by an effective input noise, N_{eff}. In the case of the maser this noise is known to be

$$N_{\text{eff}} = Kh\nu B,$$

where $K = n_2/(n_2 - n_1)$ and n_2 and n_1 are, respectively, the densities of upper-state and lower-state atoms in the active medium. In terms of a negative temperature of the active medium T_m, we have

$$K = \left[1 - \exp\left(-\frac{h\nu}{k|T_m|}\right)\right]^{-1}.$$

For the parametric amplifier N_{eff} may be written in a similar way, with K also greater than or equal to unity.[14] After much amplification the additive noise, given by the gain times the sum of the incident noise plus the effective input noise,[15] is *always* much greater than $h\nu B$ and so the classical formula applies for the information capacity. We find, therefore, that after much amplification the information capacity of the wave is reduced to

$$C_{\text{amplifier}} = B \log\left(1 + \frac{S}{N + Kh\nu B}\right) \tag{6}$$

where S is the incident signal, N is the incident noise and $K \geq 1$. Thus the best possible amplifier, for which $K = 1$, retains only the first term in the incident wave information limit, (5). We now can define the information efficiency of an amplifier as $C_{\text{amplifier}}/C_{\text{wave}}$. For the interesting case of a perfect amplifier this is plotted for various values of signal strength in Fig. 3. The incident noise is assumed the same as for Fig. 2.

After much amplification we may assume that all of the information remaining in the wave can be extracted, so (6) also gives the information capacity of a system using a high gain coherent amplifier at the carrier frequency as the first element of the receiver.[16] For small N the efficiency drops off for signal levels less than about $h\nu B$, indicating a substantial loss of information in this region for such a system.

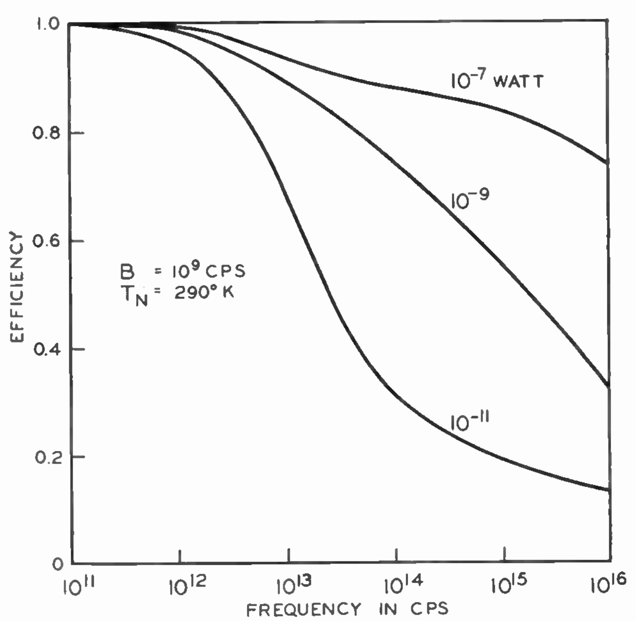

Fig. 3—Information efficiency for an ideal amplifier of high gain. Because of spontaneous emission, the ideal amplifier has an effective input noise power of $h\nu B$, which is responsible for the lowering of its efficiency at high frequencies.

VII. The Heterodyne Receiver

Instead of amplifying the wave we might immediately make use of a photoelectric device in a heterodyne receiver.[17,18] To do this we might let the signal and power from a CW local oscillator fall simultaneously on a photosensitive element.

Then the photocurrent is proportional to the instantaneous power P_{inst} incident on the element. If the quantum efficiency of the photosensitive device is ϵ, the current is given by

$$I = \frac{\epsilon P_{\text{inst}}}{h\nu} q$$

where q is the electronic charge. Let the signal frequency be ω_{sig} and the local oscillator frequency be ω_{local}. If the local oscillator power is much greater than the signal power, the instantaneous power will have the form

$$P_{\text{inst}} \approx P_{\text{local}} + 2\sqrt{P_{\text{sig}} P_{\text{local}}} \cos(\omega_{\text{sig}} - \omega_{\text{local}})t + \cdots$$

where P_{sig} is the instantaneous input signal power and P_{local} is the local oscillator power. The photocurrent thus consists of a dc component

$$I_0 = \frac{\epsilon q}{h\nu} P_{\text{local}}$$

and a signal current at the intermediate frequency whose mean square is

$$\overline{I_{\text{sig}}^2} = 2\left(\frac{\epsilon q}{h\nu}\right)^2 S P_{\text{local}}$$

where S is the average input signal power. Because of the dc current there will be shot noise, whose mean square is

$$\overline{I_N^2} = 2qI_0B = 2\left(\frac{\epsilon}{h\nu}\right) q^2 P_{\text{local}} B.$$

[14] W. H. Louisell, A. Yariv, and A. E. Siegman, "Quantum fluctuations and noise in parametric processes. I," *Phys. Rev.*, vol. 124, pp. 1646–1654; December, 1961.

[15] It has now been established[9,10,11] that the simple addition (voltagewise) of the amplified effective input noise to the classically amplified signal and real input noise accounts for all fluctuations in the output wave. That is, if the signal wave leaving the transmitter has the form $v_s \cos(\omega_s t + \phi_s)$, then the amplified wave has the form $(G/L)^{1/2} v_s \cos(\omega t + \phi_s) + G^{1/2} v_n \cos(\omega t + \phi_n)$, where G and L are the gain and loss of the amplifier and attenuator, respectively, and where the added term in the amplified wave is the fluctuating white noise voltage. This is rigorously true no matter how small, in terms of quanta per mode, the signal may be at the amplifier input. We are of course assuming that the gain and loss are not subject to fluctuations caused by such things as variations in the density of attenuating or amplifying particles, variations in pumping of a parametric amplifier, etc.

[16] It might appear that we are departing somewhat from common usage here by speaking of the information capacity of a system using a specific receiver. The reason for it is that in quantum mechanics the properties of the measuring apparatus (*i.e.*, the receiver) inevitably influence to some extent the quantities to be measured. Thus, while we can obtain from entropy considerations an upper limit to the capacity of *any* system, from which we may derive "efficiencies" for particular systems, this upper limit cannot be termed a capacity. It would seem that we cannot obtain any expression which might properly be called a channel capacity unless we include as an essential part of the channel such elements of the receiver as are necessary to insure that subsequent measurement can be performed with no further appreciable reaction back on the channel itself.

[17] A. Javan and R. Kompfner, private communication.

[18] B. M. Oliver, "Signal-to-noise ratios in photoelectric mixing," Proc. IRE, vol. 49, pp. 1960–1961; December, 1961.

The ratio $\overline{I_{sig}^2}/\overline{I_N^2}$ is the signal-to-noise ratio at the IF, which comes out to be simply $\epsilon S/h\nu B$. This implies an information capacity for the IF signal of

$$C_{\text{heterodyne}} = B \log \left(1 + \epsilon \frac{S}{h\nu B}\right).$$

It is not difficult to include the effect of incident additive noise coming in with the signal. This simply reduces the signal-to-noise ratio at the IF to

$$\frac{S}{N + \frac{1}{\epsilon} h\nu B}$$

and the information capacity to

$$C_{\text{heterodyne}} = B \log \left[1 + \frac{S}{N + \frac{1}{\epsilon} h\nu B}\right].$$

The information capacity of a system using a heterodyne receiver thus has the same form as that of a system using a coherent amplifier, with K replaced by ϵ^{-1}.

VIII. The Homodyne Receiver

It was pointed out by B. M. Oliver[18,19] that the homodyne receiver has quite interesting properties. In this case we confine the modulation to *amplitude* modulation, along with an allowed phase shift of π, and then use a local oscillator in the receiver which has exactly the same frequency and phase as the signal carrier. Since $\cos(\omega_{sig} - \omega_{local})t$ is then always equal to ± 1, the instantaneous power incident on the photocell is

$$P = P_{\text{local}} + 2\sqrt{P_{\text{sig}}}\sqrt{P_{\text{local}}} + \cdots$$

where the quantity $\sqrt{P_{\text{sig}}}$ may range through positive and negative values according to the modulation amplitude and phase. The dc component of the photocurrent is again

$$I_0 = \frac{\epsilon q}{h\nu} P_{\text{local}}.$$

For this case, however, the signal current is at baseband and has bandwidth $B/2$, where B is the high-frequency band used for transmission. The mean-square shot current at baseband is therefore

$$\overline{I_N^2} = 2qI_0(B/2) = \left(\frac{\epsilon}{h\nu}\right) q^2 P_{\text{local}} B,$$

while the mean-square signal current is now

$$\overline{I_{\text{sig}}^2} = 4\left(\frac{\epsilon q}{h\nu}\right)^2 S P_{\text{local}}$$

where again S is the average signal power, *i.e.*, the average of P_{sig}. The signal-to-noise ratio is therefore

$$\overline{I_{\text{sig}}^2}/\overline{I_N^2} = 4 \frac{\epsilon S}{h\nu B},$$

and so the information capacity of the baseband signal is

$$C_{\text{homodyne}} = \frac{B}{2} \log \left(1 + 4 \frac{\epsilon S}{h\nu B}\right).$$

As in the heterodyne case we may include incident noise without too much difficulty. The result is

$$C_{\text{homodyne}} = \frac{B}{2} \log \left[1 + \frac{2S}{N + \frac{1}{2\epsilon} h\nu B}\right]$$

where N is the average received noise in the high-frequency band B.

Oliver pointed out that in this case the equivalent input quantum noise is only half as large as that occurring in the heterodyne receiver or in the equivalent maser. At first sight this is somewhat curious. In fact it simply indicates that perhaps one cannot always deduce the effects of quantum noise simply on the assumption of some fixed equivalent input noise which is the same in all situations. In no case is the capacity of a system using a homodyne receiver greater than the capacity limit, (5), of a wave of average power S in the presence of the average incident noise N. Such a result would be truly surprising. In Fig. 4 the information efficiency for

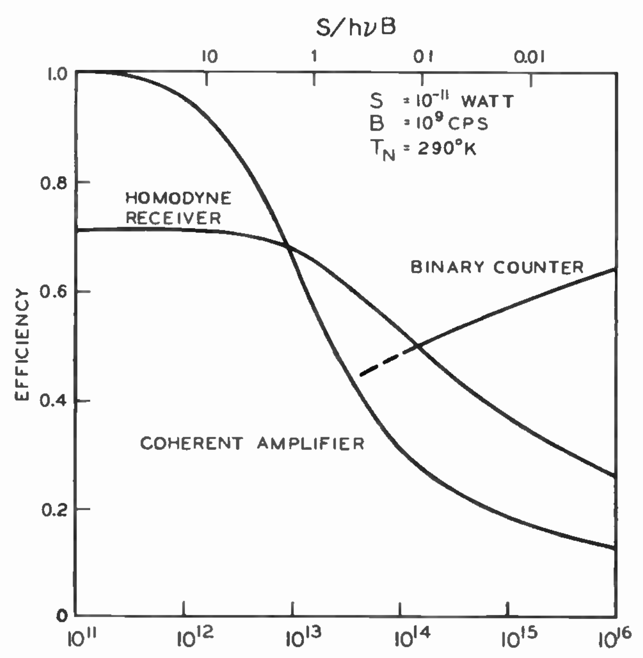

Fig. 4—Information efficiency for various receivers for an average received signal power of 10^{-11} w, a bandwidth of 10^9 cps, and an external noise temperature of 290°K. Note that at the higher frequencies the coherent amplifier is not as good as the other types of receivers.

[19] B. M. Oliver, "Comments on 'Noise in photoelectric mixing,'" Proc. IRE (*Correspondence*), vol. 50, pp. 1545–1546; June, 1962.

an ideal ($\epsilon = 1$) homodyne system is also plotted against frequency, for a signal power of 10^{-11} w and an external noise temperature of 290°K. For comparison the information efficiency of an ideal amplifier is plotted also, as well as that of an ideal detector using a binary quantum counter (see Section IX-A, and note that for frequencies of 10^{14} cps or greater the external noise may be completely neglected).

IX. The Quantum Counter

Instead of using any of the aforementioned receivers, we might simply allow the signal to fall on some photoelectric device and count the photoelectrons as they are produced. If we could do this with unity quantum efficiency and with perfect discrimination between different numbers of photoelectrons, we would surely have an ideal power-sensitive receiver. The information capacity for this general case can in principle be found since the probability distribution for the various numbers of received photons resulting from the transmission of some known number of photons has been computed.[20] Unfortunately, attempts to calculate the information capacity of a communication system using such a receiver encounter rather great computational difficulties. Nevertheless in some simple cases the problem can be solved approximately. When $S/h\nu B$ is either much larger or much smaller than unity, we may obtain approximately correct values for the capacity.

A. *The Binary Counter*

For the case $S \ll h\nu B$, the average number of photons per independent field mode is much smaller than unity, so that only the two events, no photon received or one photon received, have appreciable probabilities. Consider, then, the following communication system. The transmitted signal consists of a series of pulses, each of duration $1/B$ and of constant amplitude. The pulses occur in a statistically random sequence with the probability Q of sending a pulse in any particular time interval. A typical transmitted message would then appear as in Fig. 5. The average power in the signal is Q times the pulse power, or if the energy in each pulse is E the average power is QEB. The receiver measures the number of received photons in each time interval $1/B$; thus it makes B measurements per second, which is consistent with the notion that there are B independent field modes received per second. If the receiver simply distinguishes between no photons received or some photons received, we will have a system which should do nearly as well as possible when the average number of photons received per interval is much smaller than unity but of course is rather inefficient for larger average numbers of photons. This system has the advantage that one can compute its information capacity exactly, and we shall now proceed to do this.

Fig. 6 shows the communications channel under consideration. In each time interval $1/B$ the transmitter either emits a pulse or it does not. The probability of occurrence of a pulse in any particular time interval is Q. If the receiver detects at least one photon in any time interval, it records a 1; if not, it records a 0. To simplify matters, let us assume that the quantum efficiency of the receiver is unity, and at first let us assume that there is no noise in the channel. In this case if the transmitter does not send a pulse, the receiver definitely records a 0. This is indicated in Fig. 6. On the other hand if the transmitter sends a pulse, the receiver does not definitely record a 1. There is a finite probability that no photons reach the receiver even when the pulse is sent. This probability is known, however. So long as the number of photons in the transmitted pulse is reasonably well known, the probability distribution $q(m)$ for the various numbers m of photons received after large transmission loss is a Poisson distribution, from which

$$q(m) = \frac{s^m}{m!} e^{-s}.$$

Here the average or expected number of received photons in the pulse is labeled s. Thus the probability of receiving no photons is e^{-s}, and the probability of receiving at least one is of course $1 - e^{-s}$. These probabilities are also indicated on Fig. 6.

Now to compute information capacity we must use some further results of Shannon's work.[1] He showed that the information I per symbol (*i.e.*, time interval)

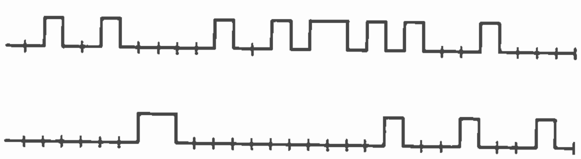

Fig. 5—Typical sequence of pulses in a message suitable for a binary communication system. The statistical probability for the occurrence of a pulse is 0.25 in this message.

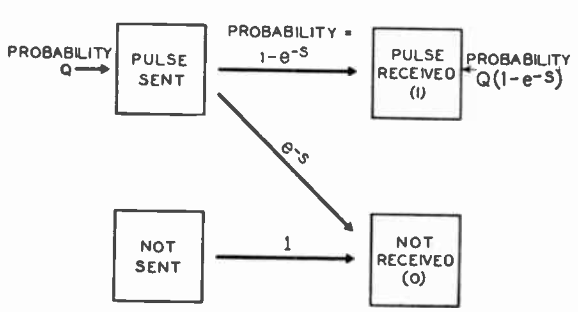

Fig. 6—Schematic diagram for a noiseless binary channel. The various probabilities necessary for the solution of the information problem are indicated on the diagram.

[20] K. Shimoda, H. Takahasi, and C. H. Townes, "Fluctuations in the amplification of quanta with applications to maser amplifiers," *J. Phys. Soc. Japan*, vol. 12, pp. 686–700; June, 1957.

for such a discrete communication channel is given by

$$I = H(y) - H_x(y)$$

where $H(y)$ is the entropy per symbol of the received message, given by

$$H(y) = -\sum p(y) \log p(y)$$

summed over the probabilities $p(y)$ for the possible received symbols y, while $H_x(y)$ is the conditional entropy of the received message, given by

$$H_x(y) = -\sum_x p(x) \sum_y p_x(y) \log p_x(y).$$

Here the quantities

$$\left[-\sum_y p_x(y) \log p_x(y)\right]$$

are the entropy per symbol of the received message when the transmitted symbol (x) is known, and $H_x(y)$ is this entropy averaged over the probability distribution $p(x)$ for transmitted symbols. Thus $H(y)$ is the total received entropy, while $H_x(y)$ is that part of the received entropy which does not contain information.

With the help of these formulas we are able to compute the information capacity of the channel. For a probability Q of sending a pulse, the total probabilities for receiving a 1 or a 0 are

$$p(1) = Q(1 - e^{-s}); \qquad p(0) = 1 - Q(1 - e^{-s})$$

while the conditional probabilities are

$$P_{\text{pulse}}(0) = e^{-s};\ P_{\text{pulse}}(1) = 1 - e^{-s},\ P_{\text{no pulse}}(0) = 1,$$
$$P_{\text{no pulse}}(1) = 0.$$

The received entropy is then

$$H(y) = -Q(1 - e^{-s}) \log [Q(1 - e^{-s})] - [1 - Q(1 - e^{-s})] \log [1 - Q(1 - e^{-s})]$$

and the conditional entropy is

$$H_x(y) = -Q[e^{-s} \log e^{-s} + (1 - e^{-s}) \log (1 - e^{-s})] - (1 - Q)[0].$$

Subtracting the two, we find

$$I = -Q(1 - e^{-s}) \log Q - [1 - Q(1 - e^{-s})] \cdot \log [1 - Q(1 - e^{-s})] + Qe^{-s} \log e^{-s}.$$

To find the maximum information per symbol we must maximize I with respect to Q, under the constraint that the average power remain constant. Now since s is the average number of received photons per pulse, and Q the probability of sending a pulse, the average number of photons per time interval is Qs. This is the quantity which must remain constant and was called $\bar{m}$ in Section I. If we therefore substitute $Q = \bar{m}/s$, where $\bar{m}$ is a constant, into I, differentiate with respect to s and set the result equal to 0, we obtain the condition for maximum I. This is

$$\log_e \left[\frac{s}{\bar{m}} + e^{-s} - 1\right] = \frac{s}{\left(\dfrac{e^s}{s+1} - 1\right)}.$$

To find $I_{\max}$ this transcendental equation must be solved for s, assuming some value of $\bar{m}$, and then the result used to evaluate I. In Fig. 7, s is plotted against $\bar{m}$. It may be seen that s does not drop off very rapidly for small $\bar{m}$. Finally $I_{\max}$ can then be calculated, and the information capacity of this system

$$C_{\text{binary}} = I_{\max} B$$

may be compared to the information limit for a noiseless wave of the same average power (*i.e.*, $\bar{m}h\nu B$) at the receiver input. The efficiency of the system

$$C_{\text{binary}}/C_{\text{wave}}$$

is plotted in Fig. 4. It may be seen to approach unity slowly at small signal levels. One can in fact show that at very very small signal levels, *i.e.*, for $\log_e(1/\bar{m}) \gg 1$, the information per symbol approaches

$$I_{\max} \to \bar{m} \log \frac{1}{\bar{m}}.$$

This may be compared to H of (3).

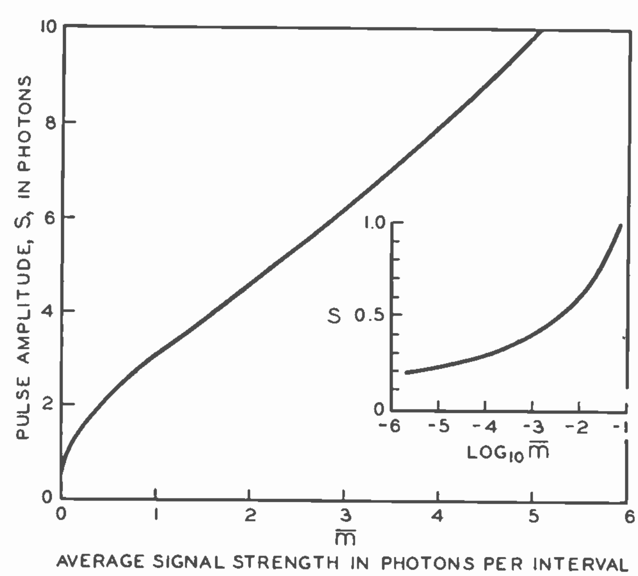

Fig. 7—Optimized average received pulse amplitude for the noiseless binary channel as a function of the average number of received photons per available time interval. The probability of sending a pulse is given by $Q = \bar{m}/s$.

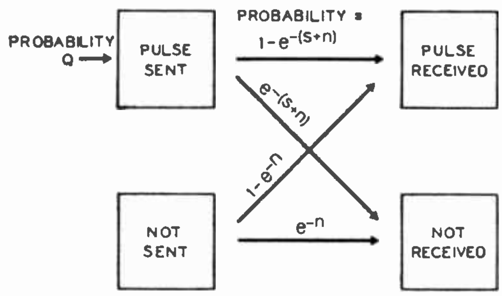

Fig. 8—Schematic diagram for a noisy binary channel.

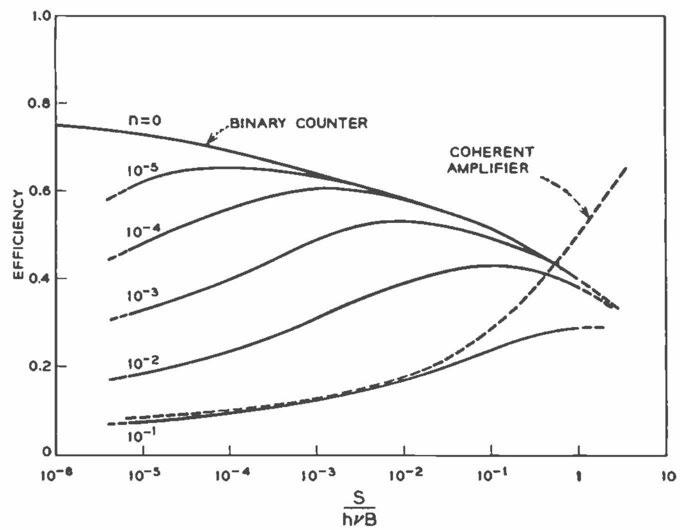

Fig. 9—Information efficiency of a binary counter perturbed by various amounts of incoherent noise. The numbers n represent the average number of noise photoelectrons per pulse interval B^{-1}. Note that the efficiency drops off when $nh\nu B > S$, and that for n greater than 0.1, the efficiency of the binary counter is always less than that of the coherent amplifier.

From this fairly simple example the mathematical complexity of the quantum counter system is reasonably evident. However, the case of the binary counter with noise is also simple enough to be calculated. There are two cases of possible interest involving noise. The first is when the noise in the counter results from noise power in the transmission mode accompanying the signal. It is then important to consider the effects of interference between noise and signal. However, there seems little use in calculating information capacities for this case since, for a given noise temperature, the number of noise photons per interval $1/B$, as a function of frequency, is for the greater part of the frequency range either greater than one—when $h\nu < kT$, or much less than one—when $h\nu > kT$. In the former case the binary counter is clearly not the most efficient receiver, in the latter the noise may be ignored so long as it is much less than the signal.

The second case involving noise in the counter is when the noise and signal are statistically independent. This would occur if the noise results from dark current in the photodetector, or from the effects of stray light incident in the receiver from modes other than the transmission mode or at frequencies outside of the useful band. If we assume that the noise photoelectrons arise from a large number of statistically independent causes, then the probability distribution of noise photoelectrons is also a Poisson. From this the conditional probabilities given in Fig. 8 follow. The results of a calculation of I_{max} for these probabilities, based on the equations of the previous section, are given in Fig. 9. For comparison, I_{max} for the noiseless case, is plotted there also.

B. The Quantum Counter When $S \gg h\nu B$

In the previous subsection we considered a particular communication system using a quantum counter for which the capacity could be calculated exactly, but which approximates an ideal receiver only when $S+N \ll h\nu B$. We can also obtain an approximate result for an ideal quantum counter system which is valid at high power levels. We assume again that the transmitter sends out a sequence of pulses, each of duration $1/B$, but with varying amplitudes. We suppose that the receiver tells the exact number of photons it receives in each such time interval, and we may assume that in the great majority of intervals it receives a reasonably large number of photons.

The calculation is carried out in the Appendix and gives the results that an exponential probability distribution for the energy of the transmitted pulses is approximately optimum, and that for this distribution and no noise the information capacity of the counter system is [see (13)]

$$C_{\text{counter}} = \tfrac{1}{2}C_{\text{wave}} - B[\log\sqrt{2\pi} + 0.289 \log e]. \qquad (7)$$

In the limit of very high power the constant term can be neglected, and the quantum counter then extracts half of the information in the wave. It is likely that when the wave capacity is small enough so that the second term begins to be significant, the exponential distribution is no longer optimum.

Having gone this far we perhaps should go on to add noise to the wave and again calculate the information capacity. In fact one can do so using similar approximate methods. Again one finds that in the limit of high power the counter system achieves half the capacity of the wave. The calculation is much like that in the Appendix, and in order not to bore the reader excessively, we shall omit it here.

The fact that a system using an energy-sensitive receiver has a capacity no greater than half of the wave capacity in the limit of high power (*i.e.*, high signal-to-noise ratio) is just what we might expect. In this limit the classical theory should give an adequate description of physical phenomena. Classically, when the signal-to-noise ratio is high, then equal amounts of information may be obtained from measurements of amplitude and measurements of phase; and the energy sensitive receiver automatically rejects all phase information.

X. Summary

We have found an expression for the absolute rate at which entropy is carried by an electromagnetic wave having the statistical properties of white noise, in a transmission medium which supports a single transmission mode. Further, we have found an upper limit for the information capacity of a wave consisting of signal with average power S in the presence of white noise with average power N. We have investigated the information capacity of a number of communication systems. The results of this investigation may be summarized as follows:

1) When the received signal or noise power is much larger than $h\nu B$, where ν is the center frequency of the wave and B its bandwidth, a receiver using an ideal coherent amplifier or an ideal heterodyne converter can extract essentially all the information that can be incorporated in the wave, while an ideal energy-sensitive receiver is limited to about half the capacity of the wave. The ideal homodyne converter is intermediate between these two.
2) When the total received power is much less than $h\nu B$, a binary quantum counter can extract essentially all the information that can be incorporated in the wave, while the other types of receivers become increasingly less efficient.
3) For a given power and bandwidth, the upper limit to the information which can be incorporated in an electromagnetic wave begins to drop off fairly rapidly when ν increases beyond P/hB. Viewed from another angle, for a given frequency and bandwidth there is a kind of threshold for received power below which the information capacity of a communications channel drops off rapidly. When external noise is absent, this power level is about $h\nu B$.

Appendix

The Quantum Counter When $S \gg h\nu B$

Our first step will be to calculate the conditional entropy $H_x(y)$. If the transmitter sends a pulse of M photons in a particular interval with any reasonably small uncertainty and there is no additive noise, the probability distribution for received photons is, as before, known to be a Poisson; that is, the probability of reception of m photons when M were sent is

$$P_M(m) = \frac{(\bar{m})^m e^{-\bar{m}}}{m!}.$$

Here $\bar{m}$ is the expected number of received photons. $\bar{m}$ is of course MT, where T is the transmission coefficient of the transmission line. By supposition, T is much less than unity. The conditional entropy of the received signal, $H_M(m)$, may then be written as

$$H_M(m) = -\sum_M p(M) \sum_m P_M(m) \log P_M(m).$$

Now

$$\log P_M(m) = m \log \bar{m} - \bar{m} \log e - \log (m!)$$

and since we assume m is large in the great majority of instances we may use Stirling's approximation for $\bar{m}!$, which is

$$\log m! \approx (m + \tfrac{1}{2}) \log m - m \log e + \log \sqrt{2\pi}.$$

Using this relation we find

$$\log P_M(m) \approx \log \bar{m} - (m + \tfrac{1}{2}) \log m - (m - \bar{m}) \log e - \log \sqrt{2\pi},$$

whence

$$\sum_m P_M(m) \log P_M(m) = \bar{m} \log \bar{m} - \log \sqrt{2\pi} - \sum_m P_M(m)(m + \tfrac{1}{2}) \log m.$$

It remains to evaluate the last term. To do this we expand $\log m$ in a power series in $(m - \bar{m})/\bar{m}$, according to the prescription

$$\log m = \log \bar{m} + \log \left[1 + \frac{m - \bar{m}}{\bar{m}}\right] = \log \bar{m} + \left[\frac{m - \bar{m}}{\bar{m}} - \frac{1}{2} \frac{(m - \bar{m})^2}{\bar{m}^2} + \cdots \right] \log e.$$

We can then make the necessary summation in terms of the moments of the Poisson distribution $P_M(m)$, for which we know that

$$\sum_m P_M(m)(m - \bar{m}) = 0, \qquad \sum_m P_M(m)(m - \bar{m})^2 = \bar{m}$$

$$\sum_m P_M(m)(m - \bar{m})^3 = \bar{m}, \qquad \text{etc.}$$

Doing this we find that

$$\sum_m P_M(m)(m + \tfrac{1}{2}) \log m = (\bar{m} + \tfrac{1}{2}) \log \bar{m} + \tfrac{1}{2} \log e + O(1/\bar{m}).$$

Substituting this in (8) we find the relation

$$-\sum_m P_M(m) \log P_M(m) = \tfrac{1}{2} \log (2\pi e \bar{m}) + O\left(\frac{1}{\bar{m}}\right),$$

and so the conditional entropy is given very nearly by

$$H_M(m) = \frac{1}{2} \sum_M p(M) \log (2\pi e \bar{m}).$$

Since M is exceedingly large over most of its significant range, we can replace this summation by an integral, thus

$$H_M(m) \approx \frac{1}{2} \int_0^\infty dM\, p(M) \log (2\pi e \bar{m})$$

and finally since $\bar{m}$ is a known function of M, we have

$$H_M(m) \approx \frac{1}{2} \int_0^\infty d\bar{m}\, p(\bar{m}) \log (2\pi e \bar{m}) \qquad (9)$$

where $p(\bar{m})$ is the probability distribution of the expected value of the number of photons incident on the receiver and $p(\bar{m})d\bar{m} = p(M)dM$.

Now let us ask how the conditional entropy varies with the choice of the expected signal distribution $p(\bar{m})$. We must know this in order to maximize the information content of the signal. As before we can expand $\log(2\pi e\bar{m})$ in a power series in $\bar{m} - \bar{\bar{m}}$, where $\bar{\bar{m}}$ is the average value of $\bar{m}$ over the distribution $p(\bar{m})$. Thus

$$\begin{aligned}
&\log 2\pi e\bar{m} \\
&= \log 2\pi e m + \log\left(1 + \frac{\bar{m} - m}{m}\right) \\
&= \log 2\pi e m + \log e \\
&\cdot\left[\left(\frac{\bar{m} - m}{m}\right) - \frac{1}{2}\left(\frac{\bar{m} - m}{m}\right)^2 + \frac{1}{3}\left(\frac{\bar{m} - m}{m}\right)^3 - \cdots\right]
\end{aligned}$$

where

$$m = \int_0^\infty \bar{m}\, p(\bar{m}) d\bar{m}.$$

Substituting this in (9), we find

$$\begin{aligned}
&H_M(m) \cong \tfrac{1}{2}\log 2\pi e m + \tfrac{1}{2}(\log e) \\
&\cdot\int_0^\infty\left[-\frac{(\bar{m} - m)^2}{2m^2} + \frac{1}{3}\,\frac{(\bar{m} - m)^3}{m^3} - \cdots\right] p(\bar{m})d\bar{m}. \quad (10)
\end{aligned}$$

It is clear from (10) that if m is large, and if the distribution $p(\bar{m})$ is any reasonably sharp distribution, the conditional entropy is very nearly given by the first term alone; *i.e.*,

$$H_M(m) \approx \tfrac{1}{2}\log(2\pi e m)$$

and thus is dependent only on the average signal power. Thus the problem of maximizing the information in the signal subject to a given average power (therefore, a given value of m) reduces simply to the problem of maximizing the received entropy. This we already know how to do. It requires an exponential probability distribution for received photons. For this distribution the received entropy is given by (3) with m replacing $\bar{m}$, and for large m may be approximated by

$$H(m) \cong \log m + \log e = \log(em). \qquad (11)$$

The most straightforward way to obtain an exponential probability distribution at the receiver is to generate an exponential probability distribution at the transmitter. Our final task is then to check whether our result for the conditional entropy at the receiver is valid for this very broad distribution as well as for narrow ones. For the exponential distribution with a reasonably large average, the probability distribution for the expected number of received photons may be assumed to be very nearly continuous and given by

$$p(\bar{m}) = \frac{1}{\bar{\bar{m}}}\exp\left(-\frac{\bar{m}}{\bar{\bar{m}}}\right).$$

For this distribution the series expansion (10), for the conditional probability converges embarrassingly slowly, so we must go back to the integral form (9). Doing this we obtain for the conditional entropy

$$H_M(m) \approx \frac{1}{2}\int_0^\infty d\bar{m}\left(\frac{1}{\bar{\bar{m}}}\right)\exp\left(-\frac{\bar{m}}{\bar{\bar{m}}}\right)\log(2\pi e\bar{m}).$$

Substituting $x = \bar{m}/\bar{\bar{m}}$ we find

$$H_M(m) \approx \frac{1}{2}\log(2\pi e\bar{\bar{m}}) + \frac{1}{2}\int_0^\infty dx \exp(-x)\log x.$$

The integral evaluates to $0.577 \log e$, so that

$$H_M(m) = \tfrac{1}{2}\log(2\pi e\bar{\bar{m}}) + 0.289 \log e. \qquad (12)$$

Comparison of this result with (10) shows that by going to the broad exponential distribution we have slightly increased the conditional entropy, but probably not enough to invalidate the conclusion that for large $\bar{\bar{m}}$ the exponential distribution is the optimum one.

Finally, we obtain for the information per symbol obtained by the ideal quantum counter,

$$I = H(m) - H_M(m)$$

where $H_M(m)$ is given by (12) and $H(m)$ by (11). We can express this as

$$I = \tfrac{1}{2}H(m) - (\tfrac{1}{2}\log 2\pi + 0.289 \log e).$$

Since $BH(m)$ for thise case is just the information capacity of the wave, C_{wave}, we find for the information capacity of the ideal quantum counter, at high power levels, the expression

$$C_{\text{counter}} = BI \approx \tfrac{1}{2}C_{\text{wave}} - B(\tfrac{1}{2}\log 2\pi + 0.289 \log e). \quad (13)$$

ACKNOWLEDGMENT

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Negative Impedance Electrometer Amplifiers—Introduction*

EDWARD F. MACNICHOL, JR., MEMBER, IRE
Editor, IRE Transactions on Biomedical Electronics

The following brief article, prepared at the invitation of the Editor, serves as an excellent introduction to the three papers which follow it.—*The Editor*

IT SOMETIMES happens that a group of scientists will develop a useful device for its own needs which could have application in a much wider field, but which remains practically unknown to the engineering profession as a whole. A case in point was the development of stable low-level dc amplifiers in physiological laboratories under pressure of the need for measuring small bioelectric potentials. Before World War II it was not uncommon for a physicist or an engineer faced with the problem of measuring a few microvolts from a source having an impedance of the order of a megohm to borrow a suitable instrument from the physiology laboratory of the local medical school, since one was not obtainable elsewhere. During World War II the technique of dc amplification was greatly advanced to fulfill the needs of such equipment as automatic tracking radar and analog computers used in gun pointing and precision bombing.

The steady drift inherent in vacuum-tube dc amplifiers was avoided by the use of contact modulators (choppers) that convert the dc input signal to a low-frequency carrier that can be amplified and demodulated by conventional means. The bandwidth limitation of the chopper amplifier was circumvented by the invention of the chopper stabilized amplifier which uses a dc amplifier in parallel with a chopper amplifier that compares its output, suitably attenuated, with the input and applies a drift correction. The chopper stabilized amplifier appears to be the ultimate in drift-free amplification of signals in the range of zero frequency to at least several megacycles. Unfortunately, however, its input impedance, as ordinarily used, is limited to a few megohms at most.

The class of amplifiers discussed in the next three papers has a much higher input impedance than any previous wide-band vacuum-tube amplifier design. This high impedance is achieved both by the use of negative feedback, which also stabilizes the gain, and by the use of positive feedback through a capacitance that has the effect of cancelling out not only the input capacitance of the amplifier, but that of the source as well.

The use of capacitance neutralization of narrow-band amplifiers goes back to the "neutrodyne" circuit of the 1920's which made possible stable triode RF and IF amplifiers and therefore radio receivers having sufficient power output to drive a loudspeaker. Even though the pentode has supplanted the triode in receiving equipment, neutralized amplifiers are still used in most transmitting equipment.

Capacitance-neutralized wide-band amplifiers, on the other hand, are a comparatively recent development. As in the case with low-level dc amplifiers they have been brought to their present stage of development through necessity in physiology laboratories. All the authors of the three succeeding papers are intimately associated with electrophysiological research. However, it is to be hoped that when better known the circuits discussed will have much broader applications in fields such as photoelectric measurements, variable capacitance and piezoelectric transducers, electrostatic memory devices, etc. It is for this reason that these three papers are presented together in the PROCEEDINGS.

It is of interest that all three authors discuss basically the same circuit which is the most straightforward of the many configurations that have been employed to accomplish the same results. In this writer's opinion, none of the others, while admittedly workable, offer any real advantages and only make mathematical circuit analysis more complicated. The original circuit was developed by P. R. Bell[1] during World War II, as a pickup device for an electrostatic storage tube used in a radar moving target indicator. This writer was privileged to be associated with Dr. Bell at the time and subsequently he and others have used the circuit in electrophysiological amplifiers.[2] However, Solms, Nastuk, and Alexander[3] should be given priority for first describing its use for this purpose.

* Received July 23, 1962.

† Thomas C. Jenkins Department of Biophysics, The Johns Hopkins University, Baltimore, Md.

[1] P. R. Bell, "Negative-capacity amplifiers," in "Waveforms," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 19, especially Appendix A; 1949.

[2] See bibliographies in next three papers.

[3] S. J. Solms, W. L. Nastuk, and J. T. Alexander, "Development of a high fidelity preamplifier for use in recording bioelectric potentials with intracellular electrodes," *Rev. Sci. Instr.*, vol. 24, pp. 960–967; October, 1953.

The three papers presented here represent the latest developments of the same basic principles in three different laboratories. The first paper, by Guld, describes an elegant circuit in use in his laboratory in Denmark and gives in detail the theoretical and practical considerations that led to its design.

The second paper, by Moore and Gebhart, also describes a circuit that is in use in the laboratory and the design considerations involved. Two items are worthy of special attention. The first is the use of an analog computer to simulate the circuit on a much longer time scale, thus avoiding the effect of stray capacitances, parasitic inductances, etc. This permits determining which circuit parameters are really important and tests the closeness of fit between the real circuit and the mathematical model represented by the analog. The second item is the application of chopper stabilization to reduce drift. To this writer's knowledge, this is the first time chopper stabilization has been applied to very high impedance amplifiers.

The third paper, by Schoenfeld, provides a theoretical basis for optimizing gain, bandwidth, and noise figure within the limitations of existing circuit elements. Although it does not arrive at any practical circuit design for a particular application, it provides the designer with sufficient insight into the problem for doing so. It should form a sound basis for future developments.

Also of potential future importance is the application of neutralization of negative impedance to solid-state amplifiers so that they may compete with vacuum-tube electrometer circuits. Although positive and negative feedback can be applied to junction transistors to raise the input impedance to several megohms the noise increases rapidly, particularly at low frequencies, as the source impedance is raised. The advent of commercially available field effect transistors with guaranteed noise figures as low as 0.5 db when used with a 1 megohm source impedance[4] has opened up new possibilities.

A circuit[5] using a field effect transistor is under development by this writer. It is shown in its present, admittedly tentative form in Fig. 1. It comprises an amplifier package that uses a silicon field effect transistor Q_1 as an input device followed by junction transistors Q_2

[4] Crystalonics Type C624, Fairchild Type FSP-400.
[5] Supported by National Science Foundation Grant G-18886.

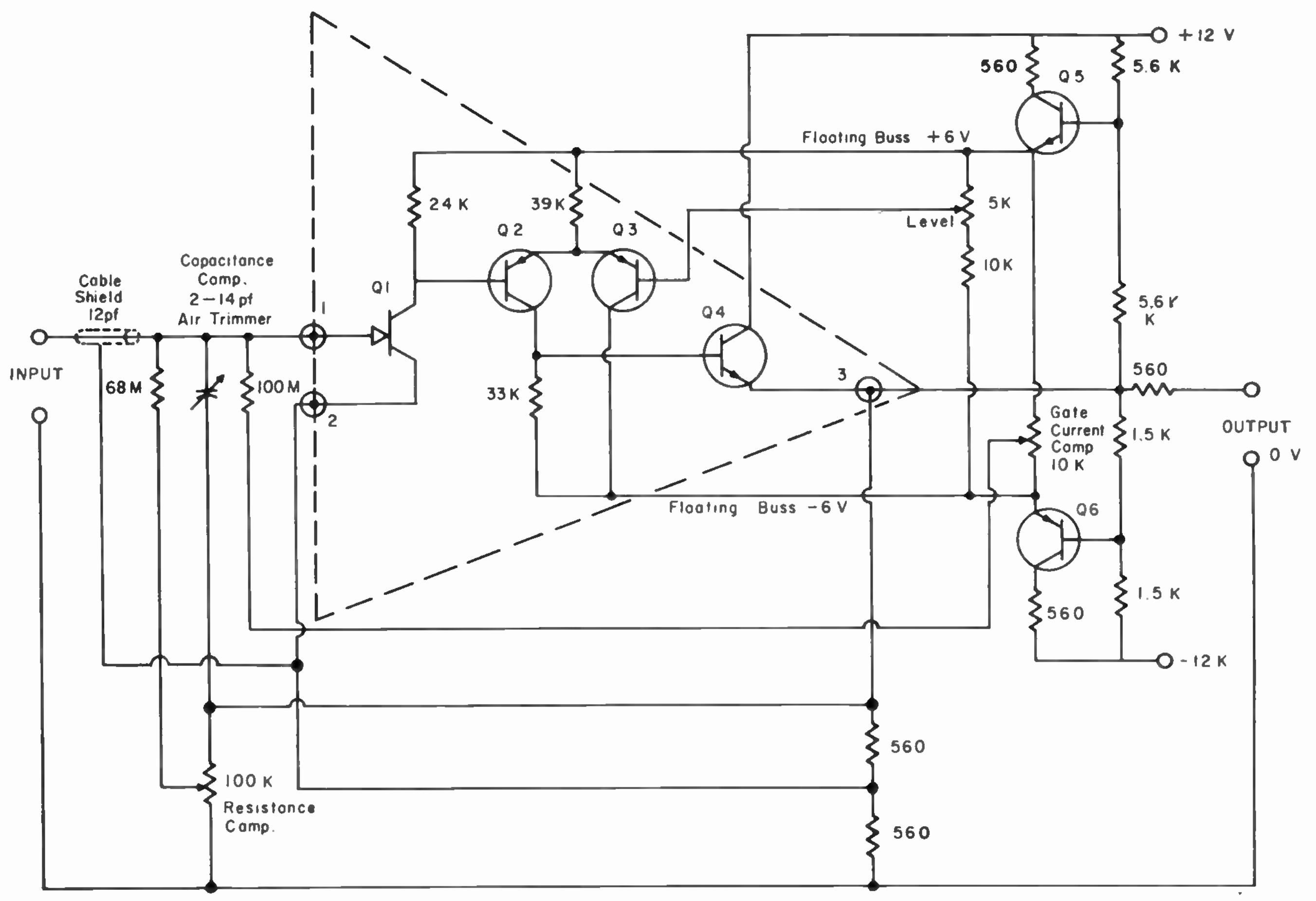

Fig. 1—Circuit diagram of field effect transistor electrometer. All resistors under 1 megohm are 1/2 w, metal film 1 per cent. All high resistors are Victoreen Hi Meg. Q_1 Crystalonics C622 silicon field effect transistor. Q_2, Q_3, Q_6 Raytheon 2N1037 silicon *PNP* transistors. Q_4, Q_5 Raytheon 2N336 silicon *NPN* transistors.

and Q_4 as an amplifier and emitter follower. The junction transistor Q_5 is used as a level setting device and provides a low impedance for the emitter of Q_2. The amplifier package has an open-loop voltage gain of approxiately 50.

Negative feedback ($\beta = 0.5$) is applied to the source of Q_1 in analogy with the vacuum-tube circuit to give the amplifier a stable voltage gain of very nearly 2. In addition, "floating" 6-v positive and negative supplies are provided by voltage divider networks and the transistors Q_5 and Q_6. These have the same variational voltage as the source of Q_1 but are displaced from it by ± 6v. These supplies have two functions: 1) they increase the dynamic range of the amplifier by supplying electrode potentials that are independent of the magnitude of the input signal and 2) they permit compensation by supplying current through a 100-megohm resistor connected to a source of adjustable voltage that has the same variational voltage as the source of Q_1 which makes the compensating current very nearly independent of the size of the input signal.[6] Compensation for input capacitance is obtained by connecting a variable capacitor between the output and the input as in the equivalent vacuum-tube circuit. However, vacuum tubes have an input resistance at low frequencies of thousands of megohms, so that compensation for input resistance is unnecessary. This is not so in the case of the field effect transistor and the circuit was found to have an input impedance of slightly over 100 megohms. This can be increased greatly by connecting an effectively negative resistance of the same magnitude to the amplifier input. This negative resistance is obtained by connecting a 68-megohm resistor to an adjustable fraction of the amplifier output.

Fig. 2 shows the response of the amplifier to a 1000-cps square wave. The top line shows the response from a low impedance source and indicates a voltage gain of 2. The second line shows the response to the same signal applied through 100 megohms with 10 $\mu\mu$f shunting the amplifier input. Capacitance compensation was adjusted for optimum response but resistance compensation was disconnected. The signal voltage attenuation of nearly 2 indicates an effective input resistance of slightly over 100 megohms. The bottom line shows the effect of adding resistance compensation to restore initial gain of 2.

The noise figure, although not yet measured directly, appears to compare favorably with most of the equivalent vacuum-tube circuits, as shown in Fig. 3. The bottom trace shows the output with a 10-μv PP square wave from a low impedance source. The top line shows the output when a 1-mv PP square wave is applied through 100 megohms with the amplifier adjusted for best compensation. The noise voltage is certainly less than an order of magnitude greater than the 125-μv rms Johnson noise that would have been generated in a 10-kc bandwidth in a 100-megohm resistor at room temperature. In addition, the troublesome microphonics generated in vacuum-tube amplifiers were conspicuously absent.

It must be emphasized that the above circuit is only a first attempt at applying the principle of impedance neutralization to wide-band solid-state amplifiers. The power consumption is only 60 mw and could probably be made much less. Microminiaturization is certainly possible so that it should be feasible to locate an array of many electrodes and preamplifiers in a small space and thus record from a large number of nerve cells simultaneously so that their interactions can be studied.

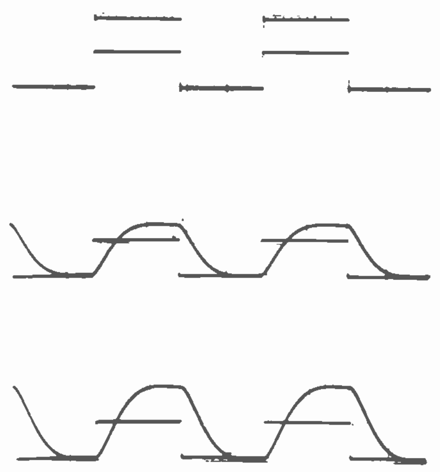

Fig. 2—Response to 1-kc square wave. Small rectangle on each trace is 100-mv *PP* 1-kc square-wave input at same oscilloscope gain as amplifier output. *Top trace*—response to square wave from low impedance source. *Middle and bottom traces*—amplifier response to 100-mv square wave through 100 megohms with 10-pf shunting amplifier input. Middle trace with capacitance but without resistance compensation. *Bottom trace*—resistance compensation adjusted to give gain of 2. Capacitance compensation optimal.

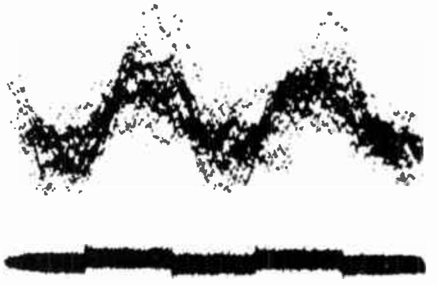

Fig. 3—Amplifier noise. *Top trace*—response of amplifier with 1-mv *PP* 1-kc square wave applied to amplifier input shunted by 10 pf through 100 megohm resistor. *R* and *C* compensation optimal. *Bottom trace*—10-v *PP* square wave applied to amplifier from 50-ohm source.

[6] J. Y. Lettvin, B. Howland, and R. C. Gesteland, "Footnotes on a headstage," IRE TRANS. ON MEDICAL ELECTRONICS, vol. PGME-10, pp. 26–28; March, 1958.

Cathode Follower and Negative Capacitance as High Input Impedance Circuits*

CHRISTIAN GULD†

Summary—Action potentials recorded across membranes may have a maximum rate of rise of up to 10^3 v/sec and the impedance of the electrode plus the generator is of the order of 10 MΩ. Therefore, to reduce distortion of the recorded signal, it is necessary to diminish the effective capacitance of the input circuit to about 1 $\mu\mu$f. It is also requisite to reduce the currents which pass through the biological specimen, both the input grid current ($<10^{-13}$ A) and the current charging the input capacitance ($<10^{-9}$ A). The performance of a cathode follower and a negative capacitance as to reduction of input capacitance was measured by the damping factor a and the time constant T' of the second-order transfer function. An equivalent input time constant $T_{eq}=aT'$ of 10 μsec ensures small distortion and negligible current through the cell. Whether a cathode follower or a negative capacitance is the more suitable depends on the value of that part of the input-ground capacitance C_g, which cannot be removed by screening. When C_g is large (microelectrode deeply immersed in the specimen) a negative capacitance is advantageous; with a small value of C_g (electrode immersed <1 mm) the cathode follower may neutralize to a $T_{eq}=30$ μsec as does a negative capacitance with a cutoff frequency of 200 kc. A $T_{eq}<10$ μsec was obtained with a negative capacitance by extending its frequency range to a cutoff frequency of several megacycles or by the use of an oscillatory response ($a<1$). The last mentioned procedure ascertains low noise. Compensating the grid current through a large resistor (10^{12} Ω) a small value of T_{eq} can be combined with a low noise factor. T_{eq} and the resistance of the electrode can be tested accurately by applying the test signal to the input via a small condenser.

I. List of Symbols

Symbol		Definition
A'	Gain of the feedback path	Appendix I (31)
A	Amplitude of step voltage	Appendix I
A_0	Low-frequency gain of A'	Fig. 1
a	Damping factor on the second-order transfer function	(3)
C_{eq}	Equivalent input capacitance after neutralization	(8)
C_f	Feedback capacitance	Fig. 1
C_g	Total input-ground capacitance	(25)
C_g'	Input-ground capacitance without signals	Fig. 7
C_k'	Cathode-ground capacitance in cathode followers	(10)
C_k	Capacitor for application of test signal	Fig. 7(b)
C_p	Stray parallel capacitance of model resistor	Fig. 7(a)
$C_i=C_f+C_g$	Total input capacitance	Section III-D
$\overline{e_{nt}^2}$	Total mean-square noise from amplifier input and electrode	(17)
F	Noise factor	(18)
f_i	Cutoff frequency of the feedback path (*cf.*, T_i)	Figs. 10 and 12
I	Strength of impulse	(24)
i_c	Current charging the equivalent input capacitance	(8a)
R_e	Resistance of the electrode	Fig. 1
R_{eq}	Equivalent noise resistance at short-circuited input	Appendix II (52)
T'	Normalizing time constant	(2)
T_{eq}	Equivalent input time constant after neutralization	(7)
T_i	Time constant of the feedback path (*cf.*, f_i)	Fig. 1
V_e	Laplace transform of the voltage v_e applied to the electrode tip	Fig. 1
V_i	Laplace transform of the voltage v_i at the impulse grid	Fig. 1
V_k	Laplace transform of the test voltage v_k applied to C_k	Fig. 7(b)
V_0	Laplace transform of the output voltage v_0, referred to unity gain	Fig. 1
V_0'	Laplace transform of the output voltage v_0' at gain A_0	Fig. 1
Δv_{01}	Error on the action potential proportional to its first derivative	(4b)
Δv_{02}	Error on the action potential proportional to its second derivative	(4c)

II. Introduction

The POTENTIAL across the membrane of single cells is usually measured with an electrolyte-filled glass pipette microelectrode placed inside the cell. Perforation of the cell with a microelectrode with a diameter of for example 2 μ injures the membrane and

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† Institute of Neurophysiology, University of Copenhagen, Denmark.

inevitably changes its state. The errors in measurement caused by the injury are of minor importance when the diameter of the tapering tip of the electrode is less than 0.5 μ [1]. These electrodes, filled with 3 m KCl, have impedances in the range of 5 to 50 MΩ.

The potential measured across the membrane is also influenced by the circuit composed of amplifier input impedance, electrode impedance and membrane impedance. When the membrane potential changes a current flows in the circuit and charges the input to ground capacitance; in addition there is a steady flow of grid current through the electrode and the cell membrane. Two types of error arise from these currents:

1) The currents may induce a change in the state of the cell membrane or may even damage it.

2) The currents cause a voltage drop over the electrode resistance+membrane impedance and the voltage recorded differs from the correct transmembrane potential by this voltage drop.

To eliminate these errors input circuits are required which have a high input impedance and a low grid current. The input impedance can be increased by reducing the capacitance of the amplifier input and microelectrode, and the types of input circuit used in biology are:

1) The input cathode follower introduced as a triode cathode follower [2] and used later as a pentode cathode follower [3]. In these circuits the input capacitance is reduced by connecting the screen around the input lead to the cathode.

2) The so called "negative capacitance" [3]–[9]; the input capacitance is reduced by introducing positive feedback to the amplifier input via a capacitance.

Each of the two circuits has its own merits and the choice depends on the demands of the experiments in question. These requirements are in general

1) the frequency spectrum to be transmitted through the recording system (microelectrode+input stage) without distortion,
2) the amount which can be tolerated of the current charging the input capacitance,
3) the amount of grid current which can be tolerated, and
4) the maximum acceptable noise.

The question is then which intrinsic properties of the input stage satisfy a certain set of requirements, and which are feasible? A rigorous analysis of the performance of the cathode follower and the negative capacitance would provide an answer to these questions and in particular put the choice between the two circuits on a more rational basis.

Previously the frequency spectrum transmitted by the negative capacitance has been calculated solely for the critically damped response [10], [11]. In the calculation of the frequency spectrum of the cathode follower [3], [4], [12] the input-ground capacitance was disregarded. It will be shown below that this capacitance determines whether the response of a cathode follower is oscillatory or not. The analysis of the noise problem in these circuits has been qualitative [8], [9] or approximate [13] and the dependence of noise upon the degree of reduction of input capacitance is not known.

The purpose of the present study is to fill these gaps by an analysis of the performance both of the cathode follower and of the negative capacitance not restricted to the critically damped response. It will be shown that other types of response may be advantageous; and that frequency response, noise and grid current are determined somewhat by the same circuit parameters. This interdependence is unfavorable in that improvement of one characteristic often involves deterioration of others. A compromise between the different requirements is sometimes inevitable and I will discuss below how the best compromise can be obtained. The applicability of the theory is examined on a cathode follower and a negative capacitance of improved performance, whereby it is shown that the performance of the given circuits can be predicted from theory. A prerequisite for this comparison is a correct measurement of the input impedance and electrode impedance and the methods to achieve this are analyzed and discussed.

III. Neutralization of Input Capacitance[1]

A. General Considerations

The theory of capacitance neutralization is similar for the cathode follower and for the negative capacitance. The general circuit is shown in Fig. 1. C_f represents the sum of capacitances exposed to feedback from the amplifier output to the amplifier input. C_g represents the sum of input-ground capacitances from the electrode and from the amplifier input. The feedback amplifier A' is characterized by the low-frequency amplification A_0 and by the time constant T_i. The analysis is idealized by the use of a pure resistance R_e as model of membrane and electrode, and by assuming that the output impedance of the amplifier is zero.

The system composed of amplifier and electrode (Fig. 1) has the transfer function [Appendix I (35), (36) and (38)]

$$\frac{V_0}{V_e} = \frac{1}{1 + saT' + s^2(\frac{1}{2}T')^2}. \tag{1}$$

This is a general transfer function of second order characterized by the normalizing time constant T' and the damping factor a. The constants of amplifier and electrode determine a and T' by the equations

$$T' = 2[R_e(C_f + C_g)T_i]^{1/2} \tag{2}$$

[1] The term "Neutralization of input capacitance" indicates the reduction of input capacitance in the cathode follower and in the negative capacitance. The term "negative capacitance amplifier" is somewhat misleading and "negative capacitance" or "negative input capacitance" are used instead.

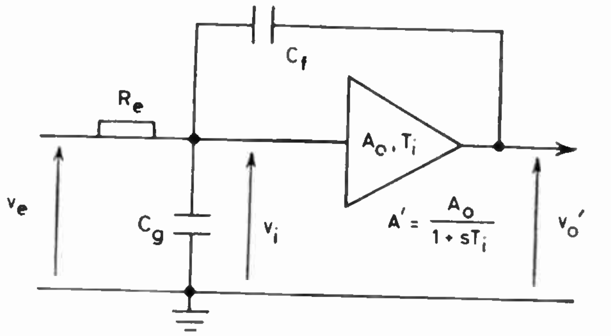

Fig. 1—Input capacitance neutralization in the cathode follower and in the negative input capacitance. R_e—electrode resistance, C_g—input-ground capacitance, C_f—feedback capacitance, A_0—low-frequency gain, T_i—time constant of the feedback path, v_e—voltage at the electrode tip, v_i—voltage at the input grid, v_0'—voltage at the amplifier output. $v_0 = v_0'/A_0$.

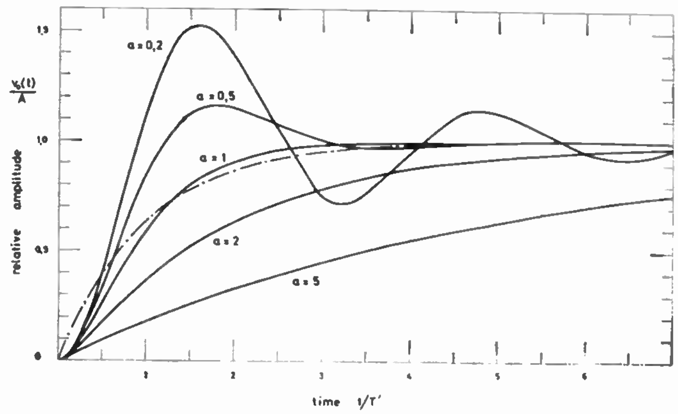

Fig. 3—Solid curves—Step responses corresponding to Fig. 2 [calculated from (40)–(42)]. The step applied to the electrode tip has amplitude A volt. Stippled curve—step response of a first-order transfer function with time constant T'[10].

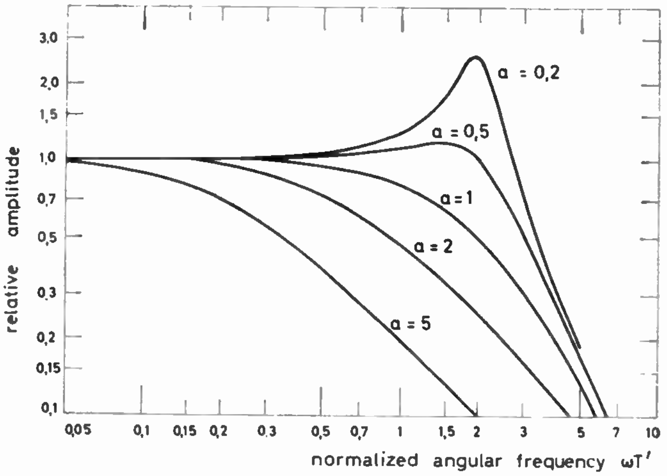

Fig. 2—Amplitude frequency response from the second-order transfer function (1) for the neutralizing input capacitance of Fig. 1. The response is calculated from (39). ω—angular frequency, T' normalizing time constant (2) and a damping factor (3).

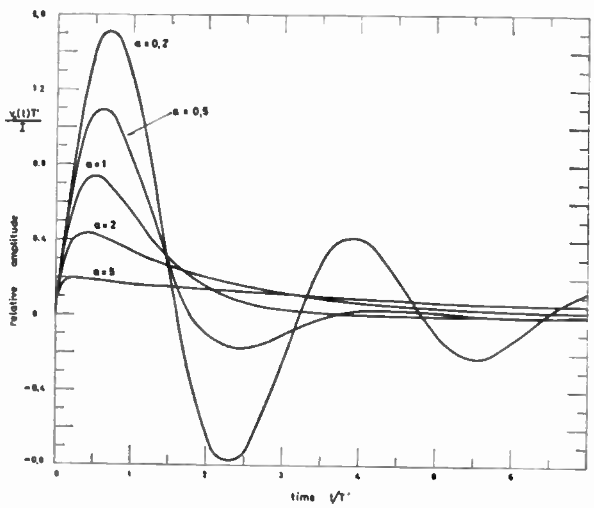

Fig. 4—Impulse response corresponding to Fig. 2 [calculated from (43)–(45)]. The impulse applied to the electrode tip is I volt × sec.

and

$$aT' = R_e(C_g + C_f - A_0C_f) + T_i. \tag{3}$$

The significance of a and T' is seen from the amplitude-frequency responses (Fig. 2), the step responses (Fig. 3) and the impulse responses (Fig. 4) associated with the transfer function (1) (see Appendix I). T' determines the range of frequencies transmitted or the time to a given response and a the amplitudes of the responses at a normalized frequency or time. The response is critically damped when $a = 1$, damped oscillatory when $0 < a < 1$ and overdamped when $a > 1$.

The analysis of the transfer function is subdivided in

1) a determination of the values of a and T' required to obtain the desired accuracy of recording (Section III-B), and
2) a determination of the extent to which these values of a and T' can be obtained with a cathode follower (Section III-C) and with a negative capacitance (Section III-D).

B. Required Values of a and T'

The necessary values of a and T' are determined from the requirements of 1) a transmission of the recorded membrane potential without distortion and 2) a charging current to the input capacitance which does not change the properties of the membrane.

When neutralizing the input capacitance, the input circuit is described by the transfer function of second order (1). Its critically damped response is comparable with that of a transfer function of the first order with the time constant T' (see Fig. 3). Therefore, at $a = 1$, it was suggested to determine distortion [10] and input charging current [5] by the methods available for the first-order transfer function [4], [12b]. The following analysis of the second-order transfer function concerns also the oscillatory and overdamped cases ($a \gtrless 1$).

By rearranging (1) and translating to the time domain the following equations are obtained:

$$v_e(t) = v_0(t) + \Delta v_{01} + \Delta v_{02} \tag{4a}$$

with

$$\Delta v_{01} = aT' \frac{d(v_0(t))}{dt} \tag{4b}$$

$$\Delta v_{02} = \frac{1}{4} T'^2 \frac{d^2(v_0(t))}{dt^2} \cdot \tag{4c}$$

Thus the membrane potential $v_e(t)$ can be calculated from the recorded potential and its first and second derivatives when a and T' are known. Instead of correcting for the distortion an attempt is made below to determine the values of a and T' at which the distortion can be disregarded when recording nerve and muscle potentials with a fast initial phase. In this evaluation the maximum values are considered. The fast transmembrane action potential has an amplitude of about 100 mv; at half amplitude of the rising phase the first derivative reaches its maximum value of about

$$(dv_0/dt)_m = 10^3 \text{ v/sec} \tag{5}$$

and the second derivative reaches maximum at the foot and near the top of the rising phase with values of about

$$(d^2v_0/dt^2)_m = \pm 10^7 \text{ v/sec}^2. \tag{6}$$

These values are considered independent of a and T', since the main effect of small distortions is a delay of the action potential [4].

The errors Δv_{01} and Δv_{02} as functions of a and T' are seen in Fig. 5. When T' is less than 100 μsec and a larger than 0.2, Δv_{02} is small relative to Δv_{01}, and the error on the action potential is proportional to the time constant (4b)

$$T_{eq} = aT' \tag{7}$$

denoted as the equivalent input time constant. This error can be reduced either by using a small value of T' or by using an oscillatory response ($a<1$). With an error of 10 mv at half amplitude of the rising phase the equivalent input time constant $T_{eq}=aT'$ required is 10 μsec. This time constant is obtained for example with $T'=30$ μsec and $a=0.3$; the error Δv_{02} being still less than 3 mv.

The final question is whether the current, charging the input capacitance is sufficiently low with an equivalent input time constant of 10 μsec. With an electrode resistance R_e and from (7) the equivalent input capacitance is

$$C_{eq} = \frac{T_{eq}}{R_e} = \frac{aT'}{R_e} \cdot \tag{8}$$

The charging current is then

$$i_c = C_{eq} \cdot (dv_0/dt). \tag{8a}$$

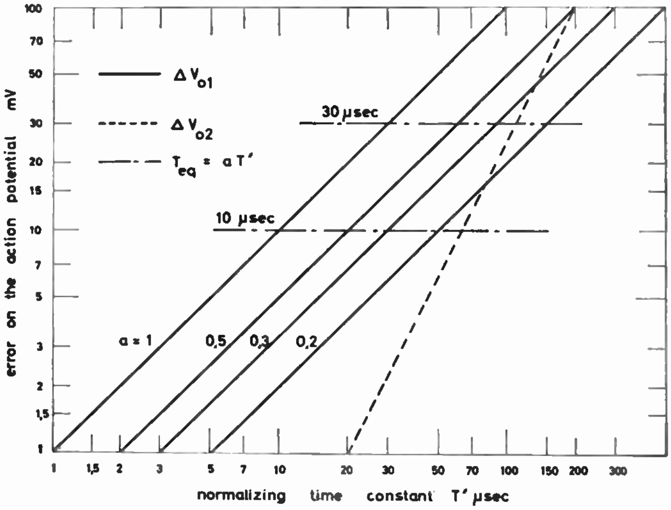

Fig. 5—Maximum errors on the rising phase of an action potential as function of the normalizing time constant T' and the damping factor a. Δv_{01} (4b), is given at $dv_0/dt=10^3$ v/sec and Δv_{02} (4c), at $d^2v_0/dt^2=\pm 10^7$ v/sec^2.

With $R_e=10$ MΩ, the equivalent input capacitance, corresponding to $T_{eq}=10$ μsec, is $C_{eq}=1$ $\mu\mu f$, giving a charging current (8a), (5) of 10^{-9} A. Let us assume that a charging current which causes a voltage drop of less than 1 mv does not change the properties of the membrane. With a charging current of 10^{-9} A, the membrane impedance seen from the electrode tip can be up to 1 MΩ. The smallest cell whose transmembrane properties have been measured is the toad motor neuron; the total dc resistance of the membrane is 5 MΩ [14]. The membrane impedance falls during the rising phase of the action potential, and it is reasonable to assume that it is less than 1 MΩ at half amplitude, *i.e.*, at maximum charging current. In most instances, therefore, the charging current of 10^{-9} A is negligible, and a T_{eq} of 10 μsec ensures distortion free recording of the action potential and negligible current through the cell. Amatniek [7] and Bak [8] suggest that time constants of the order of 2–3 μsec are necessary; in fact this is so only, as at nodes of Ranvier, when the membrane resistance approximates 50 MΩ.

C. The Cathode Followers

The principal circuit for neutralization (Fig. 1) applies to the cathode follower when:

1) the feedback capacitance C_f is equal to the capacitance between the input lead and its cathode-connected screen plus the capacitance between grid and cathode of the input tube.
2) the input-ground capacitance C_g is equal to the sum of the capacitances from the electrode, the stray input capacitance and the grid anode capacitance of the input tube.
3) A_0 is the amplification from the input to the cathode of the cathode follower [15], thus

$$A_0 = [1 + \mu^{-1} + (g_m R_k)^{-1}]^{-1} \tag{9}$$

where R_k is the cathode resistor, μ the amplification factor and g_m the transconductance of the tube.

4) the time constant of the feedback path is that of the cathode output

$$T_i = \frac{1}{g_m} C_k' \tag{10}$$

where C_k' is the capacitance between cathode and ground, including the capacitance of the output cable.

To analyze the neutralization in the cathode follower, the conditions (9) and (10) are inserted in (2) and (3). After rearranging and applying (8) we obtain the equivalent input capacitance C_{eq} and the normalized time constant T' as

$$C_{eq} = \frac{aT'}{R_e} = C_g + C_f\left(\frac{1}{\mu} + \frac{1}{g_m R_k}\right) + C_k' \frac{1}{g_m R_e} \tag{11}$$

provided μ and $g_m R_k \gg 1$

$$T' = 2[R_e(C_f + C_g)C_k'/g_m]^{1/2}. \tag{12}$$

Eq. (11) gives the contributions to the equivalent input capacitance of the total input-ground capacitance C_g, the total input cathode capacitance C_f and the cathode-ground capacitance C_k'. The proper choice of the constants of the cathode follower makes the contributions of C_f and C_k' to C_{eq} negligible (see below). Then (11) is reduced to

$$aT' = R_e C_g \tag{13}$$

which shows that the equivalent input time constant $T_{eq} = aT'$ (7) in a cathode follower can be reduced to the time constant of the electrode resistance and the stray input-ground capacitance; the damping factor a is given by the relation between T_{eq} and the time constant T'(12). Thus a critically damped response is obtained when $T' = R_e C_g$. When $T' < R_e C_g$ the response is oscillatory and when $T' > R_e C_g$ the response is overdamped. To obtain a certain value of T' mainly depends upon the selection of a proper value of the cathode-ground capacitance C_k'.

To obtain a small contribution from C_f and C_k' to C_{eq} (11), it is essential to use a pentode cathode follower, *i.e.*, a cathode follower in which the screen grid follows the cathode potential. Due to its high μ the reduction of C_f (11) is then obtained by a large value of $g_m R_k$. Furthermore there is no contribution to C_g from the grid-anode capacitance, since the capacitance to the screen grid is included in C_f. When the input (including an input switch, the tube and the electrode holder), is surrounded by a screen connected to the cathode, the stray input capacitance is reduced to a degree that the capacitance of the electrode alone determines C_g. The minimum value of the electrode capacitance is 2–3 $\mu\mu$f and is obtained when only the tip of the microelectrode is immersed. The equivalent input time constant T_{eq}, which corresponds to an electrode with a resistance of 10 MΩ, is 30 μsec. This is close to the value of 10 μsec required to ensure that the recorded action potential is but little distorted and that the input current is so small as not to affect the membrane (see Section III-B). When T' is 30 μsec as well [$a=1$ in (13)] the error from the second derivative is negligible (Fig. 5). The conditions necessary to obtain the desired value of T' are described in Section VII-A.

D. The Negative Capacitance

Fig. 1 represents a negative capacitance with a single feedback amplifier, which has at least two stages of amplification to obtain the positive feedback of a gain A_0 larger than unity. The amplifier is assumed to have a maximally flat amplitude frequency characteristic whose upper limiting frequency f_i determines the internal amplifier time constant $T_i = 1/(2\pi f_i)$.[2] The feedback capacitance C_f may be separate or include the capacitance between the inner core and the screen of an input cable.

To analyze the negative capacitance (2) and (3) are rearranged, using (8)

$$C_{eq} = \frac{aT'}{R_e} = C_g + C_f(1 - A_0) + \frac{T_i}{R_e}. \tag{14}$$

$$T' = 2[R_e(C_f + C_g)T_i]^{1/2}. \tag{15}$$

With the amplification A_0 exceeding unity, the contribution of C_f in (14) to the equivalent input capacitance C_{eq} is negative and compensates that from the input-ground capacitance C_g. By adjusting the three parameters A_0, C_f and C_g either singly or in combination, C_{eq} and T_{eq} can be reduced to values near zero. Eq. (15) shows that the normalizing time constant T' at a certain electrode resistance R_e is determined by the product of the total input capacitance $C_f + C_g$ and the time constant T_i of the feedback amplifier. This product can be reduced to a degree that T' becomes less than 10 μsec (see below). Thus values of T_{eq}, T' and of the damping factor $a = T_{eq}/T'$ can be obtained which are requisite for negligible distortion and input current when transmembranal action potentials are recorded (see Section III-B). C_{eq} and a are determined by a small difference between the two first terms on the right-hand side of (14). A small variation in C_g, A_0 or C_f causes a large variation in C_{eq} and a, hence these three parameters must be stable. On the other hand, the damping factor a varies only with the half power of the electrode resistance R_e [(14) after insertion of T'(15)]. This is an advantage because R_e is rather unstable.

[2] When the characteristic falls off with 20 db/decade, f_i is the frequency at amplitude 3 db down; with a fall of 40 db/decade, f_i is half the frequency at amplitude 6 db down.

T' is brought to a minimum when the product of total input capacitance $C_t = C_f + C_g$ and the internal time constant T_i in (15) is minimum. It is necessary to find the minimum of the product C_tT_i as a reduction of one of the factors may result in an increase of the other factor, the product remaining constant. The minimum of the product C_tT_i occurs at a certain low gain A_0. Eq. (14) shows that an increase in A_0 reduces C_f at a given value of C_g. On the other hand, since the "gain-bandwidth product" is constant (*cf.*, the theory of wide-band amplifiers) an increase in A_0 is also associated with an increase in T_i. A minimum in T_i is obtained by cascading a certain number of amplifying stages; at low amplification this number is small. With gains of 2 to 4 times and with four stages of amplification the minimum of C_tT_i is rather flat. A placement of the input tube close to the biological specimen to reduce C_t is of no advantage since T_i is increased by the capacitance of the connection between the input tube and the main amplifier. The lowest value of C_tT_i is obtained with a short input cable, using the capacitance between its inner core and its screen as the feedback capacitance C_f. The only efficient way of reducing C_tT_i is by the use of tubes with high transconductance. However, since low input grid current is required, these tubes can only be used in the stages following the input tube.

IV. The Noise

In most instances the noise from high impedance electrodes exceeds the noise from the amplifier itself. The neutralization of capacitance extends the frequency range of the input circuit and thus increases the noise from the high impedance electrode. However, when a high degree of neutralization is obtained, there is an additional increase in noise, the total noise exceeding the electrode noise. The additional noise originates from the amplifier-input noise, which is increased by neutralization. The increase is explained by the transfer function from a voltage E_i in the amplifier input (Fig. 6) to the output voltage V_0 given by

$$\frac{V_0}{E_i} = \frac{1 + sR_e(C_f + C_g)}{1 + saT' + s^2(\frac{1}{2}T')^2} \tag{16}$$

[Appendix II, (50)].

This transfer function differs by the term $sR_e(C_f + C_g)$ from the transfer function from a voltage at the electrode tip (1). Thus, the higher frequencies are transmitted from the amplifier input with an amplitude response which increases by 20 db/decade above the response from the electrode tip [8]. The noise from the amplifier input is thereby increased relative to the noise from the electrode and at a high degree of neutralization the amplifier noise may be the larger. Therefore, in the input stage of a neutralizing input capacitance the same precautions are required to reduce noise as in an amplifier for low-level signal recording. The excess noise is present independent of whether the neutralization

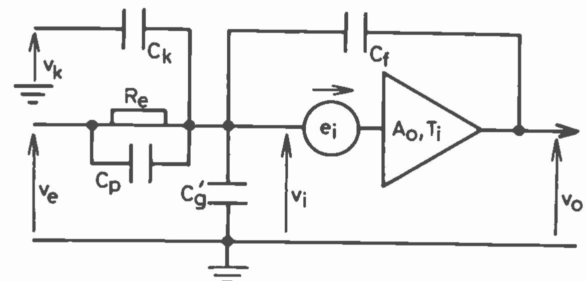

Fig. 6—General circuit for calculation of transfer functions and time responses of the neutralizing input capacitance. e_i: equivalent input noise with the input grid connected to ground. Other symbols as in Figs. 1 and 7.

is obtained in a cathode follower or in a negative capacitance, since the transfer function for noise (16) is valid for both amplifiers. The mean-square noise of a neutralizing input capacitance is calculated in Appendix II. The total mean-square noise from electrode and amplifier in volt² at room temperature is

$$\overline{e_{nt}^2} = \frac{4 \times 10^{-21}}{aT'}\left[1 + \frac{R_{eq}}{R_e} + \frac{R_{eq}(C_f + C_g)}{T_i}\right]. \tag{17}$$

R_{eq} is the equivalent noise resistance of the amplifier with short-circuited input.

The factor in front of the parenthesis is the noise from the resistance of the electrode in the frequency range obtained by the neutralization. Therefore the terms in parenthesis represent a noise factor F indicating how many times the total noise exceeds the electrode noise. As the frequency range is determined by the equivalent input time constant

$$T_{eq} = aT' = a2\sqrt{R_e(C_f + C_g)T_i}$$

the noise from the electrode is independent of whether T_{eq} is reduced by a small damping factor a obtained by an oscillatory response or reduced by a small internal time constant T_i to reduce T'. The noise factor F is independent of the damping factor a, but increases when T_i is diminished. Hence the lowest noise is obtained when the equivalent input time constant T_{eq} is reduced by the use of an oscillatory response ($a < 1$). On the other hand, a certain reduction of T' is necessary to diminish the error originating from the second derivative of the action potential (4c). The question then arises how to compromise between error and noise. With electrodes of high impedance ($R_e \gg R_{eq}$) the noise factor is independent of the electrode resistance R_e. The noise factor is given by

$$F = 1 + R_{eq}(C_f + C_g)/T_i. \tag{18}$$

When the noise from the amplifier is equal to the noise from the electrode ($F = 2$) the time constant T_i of the feedback path is [from (18)]

$$T_{im} = R_{eq}(C_f + C_g) \tag{19a}$$

and the corresponding normalizing time constant T' is [from (2) and (19a)]

$$T'_m = 2\sqrt{R_e R_{eq}}\,(C_f + C_g). \qquad (19b)$$

T_{im} and T'_m are figures of merit of a neutralizing input capacitance. A good performance as to both noise and error is obtained by minimizing T_{im} and T'_m. This demands a minimum of the total input capacitance C_f+C_g composed of all capacitances connected to the input whether grounded or exposed to feedback and input signals [Appendix I, (37)]. Including the capacitance of a deeply immersed electrode (10 $\mu\mu$f) the total input capacitance is hardly less than 20 $\mu\mu$f. Minimizing T_{im} and T'_m demands the same precautions to reduce noise in the input stage as in a low-level input amplifier. The minimum equivalent noise resistance R_{eq} is about 10 kΩ. With a 10 MΩ electrode resistance the T'_m corresponding to $R_{eq}=10$ kΩ and $C_f+C_g=20$ $\mu\mu$f is 13 μsec. Since low distortion could be obtained with T' less than 30 μsec in both the cathode follower and the negative capacitance there is little additional noise when underdamping is used to reduce distortion of the first derivative type. When utilizing the step response directly as in voltage clamp experiments (for references see [19]) underdamping may be a disadvantage and a fast response with critical damping may be preferred. In this case, in spite of the increase in noise, T' is reduced to its minimum by reducing T_i.

The increase in input noise was due to the rise in amplitude of 20 db/decade of higher frequencies transmitted from the input grid of the amplifier (16). Therefore it has been suggested that the noise be reduced by a 40 db/decade low-pass filter following the neutralizing input capacitance [8]. However the denominator in the transfer function (19) provides a 40 db/decade cutoff at high frequencies which is identical with a low-pass filter in the output.

The time constant T'_m (19b) which gives a noise factor 2 was also calculated by Weimann [13], assuming a flat frequency response from the electrode tip from dc to a frequency f_1 and zero response of frequencies above f_1. A comparison between Weimann's results and the author's show that the time constant T'_m found by Weimann's less accurate calculation is a factor $2\sqrt{3}$ too small.

V. The Input Grid Current

A low input grid current is usually obtained by running the input tube at a small anode current [16], for example 10 μa [5]. Thereby the transconductance of the input tube is reduced [17a] and the equivalent noise resistance increased [17b]. The low transconductance furthermore reduces the gain-bandwidth product of the feedback amplifier and thereby limits its frequency range, which determines the normalizing time constant T'. Thus, minimum grid current does not comply with minimum input rise time and minimum noise, and it is of interest to reduce the grid current in other ways than by lowering the anode current.

The grid current can be reduced by compensation, and this procedure may allow an anode current of the order of 0.1 ma, since the corresponding grid current of 10^{-11} A (average on 100 EF804) can be compensated such that the current through the biological specimen is less than 10^{-13} A. This input current is sufficiently small for most purposes and is of the same order as that obtained with an input electrometer tube but the procedure has the advantage of lower microphony and noise and a wider amplifier frequency range.

The compensation is provided through a large resistor connected to the input grid. Unless this resistor is sufficiently large it loads the cell with a current larger than the resulting current from the grid, because the cell generally introduces dc voltages in the input circuit of the order of 100 mv. A 10^{10} Ω resistor causes an input current of 10^{-11} and compensation to 10^{-14} A offers no advantage. If a 10^{12} Ω resistor (the largest available) is not sufficient for this purpose, the resistance can be increased by positive feedback [9].

To obtain this high input impedance Teflon and ceramics are used as insulating material in the input circuit. To ascertain that the insulation is maintained during experiments a long test pulse is applied to the input through another large resistor.

VI. Test Procedures

The resistance of the microelectrode is mainly situated in the tip and determined by its diameter, whereas the capacitance is distributed along the electrode with about 1 $\mu\mu$f/mm. Since the transmembrane potential acts on a few μ of the tip, the electrode can be represented to a first approximation by a pure series resistor and a capacitance to ground in parallel with the input (Figs. 1 and 7). It is therefore customary to test the response of a neutralized input capacitance by applying a rectangular pulse to a resistor in series with the input [3], [5], [7]. This test procedure is not suitable when neutralizing to input capacitances of the order of 1 $\mu\mu$f, because resistors have a parallel capacitance of the same order [Fig. 7(a)]. This parallel capacitance diminishes the rise time of the response to a rectangular pulse [18]. For this reason several investigators [4], [5], [7] found responses with much shorter time constants than should be expected, calculating from (2) with the parameters of their apparatus. To avoid this error it has been suggested [9] that the test resistor R_e be connected to ground and that the test voltage v_k be fed to the input via a small condenser C_k [Fig. 7(b)].

A. Measurement of a and T'

When a test voltage is applied to

1) a resistor with parallel stray capacitance C_p [Fig. 7(a)],

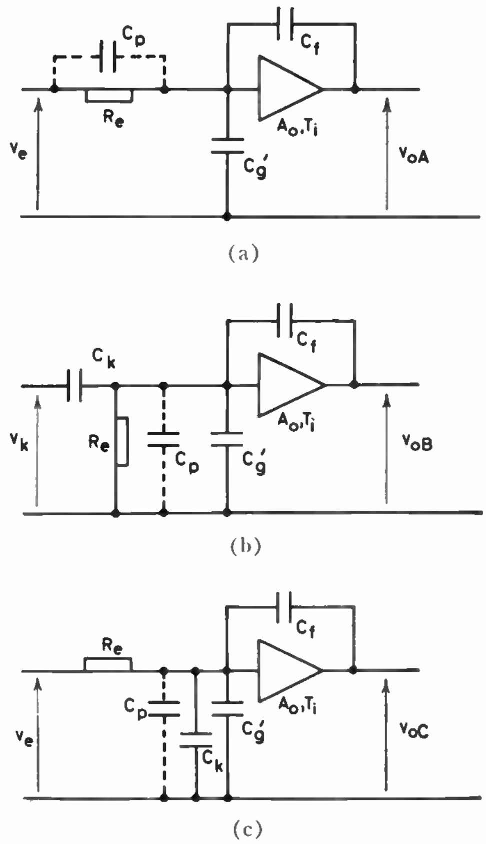

Fig. 7—Measurement of the performance of capacitance neutralization.

(a) The test signal V_e was applied to a resistor R_e with stray capacitance C_p.
(b) The test signal V_k was applied to a small condenser C_k (C_p was grounded).
(c) The test signal V_e was applied to a pure resistor R_e (*e.g.*, the tip of a microelectrode) (C_k and C_p were grounded).

C_g': Input-ground capacitance without signals. For other symbols, see Fig. 1.

2) a small condenser C_k connected to the input [Fig. 7(b)],

3) a pure resistor R_e [Fig. 7(c)],

the transformed responses are, respectively [from Appendix I (34)]

$$V_{0A} = \frac{V_e}{N} + R_e C_p \frac{sV_e}{N}, \tag{20}$$

$$V_{0B} = R_e C_k \frac{sV_k}{N}, \tag{21}$$

$$V_{0C} = \frac{V_e}{N} \tag{22}$$

where the denominator is

$$N = 1 + saT' + s^2(\tfrac{1}{2}T')^2 \tag{23}$$

with a and T' determined from (2) and (3) by the constants of the circuit A_0, T_i, C_f and C_g.

By applying a step voltage V_e to a pure resistor [(22), Fig. 7(c)] the step responses of the second-order transfer function are recorded (Fig. 3). A step voltage V_k of amplitude A_k applied via the small condenser [(21), Fig. 7(b)] gives the impulse responses of the second-order transfer function (Fig. 4), the strength I of the impulse being

$$I = R_e C_k A_k \text{ volt} \times \text{sec.} \tag{24}$$

When a step voltage V_e is applied to a resistor with a parallel capacitance [(20), Fig. 7(a)] a sum of step and impulse responses is recorded and the resulting rise time is shorter than T' due to the fast rising phase of the impulse response. The rise time is reduced about 4 times by neutralizing to an equivalent input capacitance of the same size as the stray parallel capacitance C_p [18]. On the other hand, a reliable measure of the response constants a and T' is obtained when the test impulse is applied through the small condenser: with a *step voltage* V_k applied to C_k, the peak of the impulse response (*cf.*, Fig. 4) measures the normalizing time constant T' and the undershoot gives a measure of the damping factor a. With V_k as an *integrated rectangular* pulse (*i.e.*, a triangular pulse) the step-type response (Fig. 3) is proportional to the response to a rectangular pulse applied to the electrode tip [(21), Fig. 7(c)] and the rising phase and the overshoot measure a and T'.

The total input-ground capacitance

$$C_g = C_g' + C_p + C_k \tag{25}$$

is identical in Fig. 7(b) and (c) in spite of the different connection of C_k, *i.e.*, independent of whether the test voltage is applied to C_k or to the pure resistor R_e; therefore the same values of T' and a are measured by the two responses [see Appendix I (37)].

Thus, the performance of a neutralizing input capacitance with a model resistor of known value is measured by comparing its frequency or pulse responses with those of Figs. 2–4. This comparison gives the values of a and T'. Conversely, the frequency or pulse response of a neutralized input capacitance can be fitted to a response of desired values of a and T' by adjustment of the appropriate constants of the cathode follower or of the negative capacitance (Figs. 11 and 13).

B. Testing of Electrode Impedance

The resistance of an electrode has previously been determined before recording of an action potential by measuring the voltage division between the electrode and a resistor of known value [3], [4]. To use the same procedure as control of the neutralization [4] is inaccurate, since the parallel stray capacitance of the test resistor introduces an error of the same type as testing performance by a series input resistor [Fig. 7(a)]. Testing of performance through the small condenser [Fig. 7(b)] gives in addition to a measure of a and T' a correct value of the electrode resistance and the test can be performed during and after recording of an action potential. Testing with a *triangular voltage* gives a step response

(Fig. 3) whose final amplitude is proportional to the electrode resistance R_e (21). An approximate triangular pulse can be obtained by a large rectangular voltage of amplitude V_r applied to an integrating RC filter in front of C_k [9]. When $C \gg C_k$ the load on the RC filter from the amplifier input is negligible, and $V_k = V_r/(1+sRC)$. Inserting this value of V_k in (21) and assuming $T' \ll RC$, the transfer function is

$$V_{0B} = \frac{R_e C_k}{RC} \times \frac{V_r}{N}. \tag{26}$$

With $N=1$, (23), V_{0B} is the final amplitude of the response, indicating that convenient scales for R_e as a function of V_{0B} can be obtained by adjustment of C_k, V_r or of the time constant RC.

When a *rectangular pulse* was applied to C_k the response consisted of two short spikes of opposite sign each being of the impulse type (Fig. 4). The currents, which flow through the specimen during the two spikes, compensate each other and the resulting current is small relative to that from testing with a ramp applied to C_k. From (24) and Fig. 4 the amplitude of the spike is given as

$$A_s = A_{rs}(a) \frac{R_e C_k}{T'} A_k \tag{27}$$

where $A_{rs}(a)$ is the maximum of the relative amplitude of the impulse response as function of the damping factor a. Inserting (2) in (27) gives

$$A_s = A_{rs}(a) C_k A_k \left[\frac{R_e}{(C_f + C_g) T_i}\right]^{1/2}. \tag{28}$$

The amplitude A_s is proportional to $R_e^{1/2}$ and represents a measure of R_e. Since this amplitude also depends on the damping factor a, A_s is a less accurate measure of the electrode resistance than is the amplitude (26) of the response to a triangular test pulse. Furthermore A_s varies with the total input capacitance $C_f + C_g$. Therefore, to maintain a constant value of $C_f + C_g$ during immersion of an electrode, the neutralization in the negative capacitance (Fig. 8) is adjusted by variation of a condenser in parallel with the input when the amplification A_0 of the feedback amplifier and the feedback capacitance C_f are constant. $C_f + C_g$ remain constant because the increase in capacitance due to immersion of the microelectrode corresponds to a decrease in capacitance of the variable input capacitor. A calibrated variable capacitor can be used to measure the electrode capacitance as a function of depth of immersion.

To obtain a correct measurement of the rise time of a neutralizing input capacitance, Haapanen [6] suggested that the test pulse be delivered via a transiently conducting diode. The input capacitance was charged during the transient and the passive decay after cutoff represented the measurement of the equivalent input time constant. However, after cutoff, the diode operates as a small condenser introducing a test signal as described in Fig. 7(b) and (21). Hence, when the output is thought to be underdamped, it is still overdamped, as indicated by the absence of oscillations associated with underdamping (Fig. 7 D in [6]).

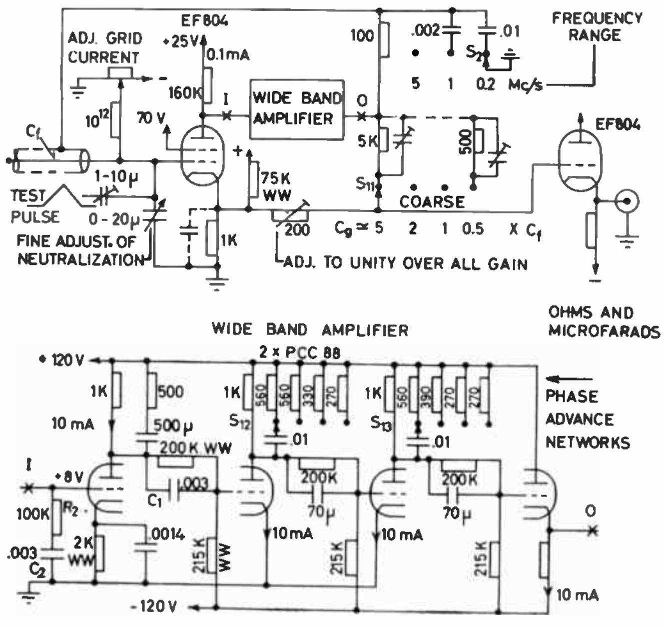

Fig. 8—Negative capacitance with extended frequency range (modified from [5]). The input coaxial cable operates as the feedback capacitance C_f. The phase advance networks in the wide-band amplifier and the capacitance compensation of the feedback voltage divider is adjusted to flat frequency response (Fig. 12). C_1 reduces the noise of the amplifier; the influence of C_1 on loop gain is compensated by R_2C_2. Designed for low drift; ¼-watt carbon film resistors unless otherwise specified.

VII. Measurements

A. The Cathode Follower

In a single cathode follower the frequency range f_i of the feedback is limited by the cathode-ground capacitance C_k' (10). This consists mainly of the capacitance of the output cable and the stray capacitance to ground of the screen of the input cable. To obtain a wide bandwidth, C_k' should be as small as possible. For this purpose a three-stage cathode follower (Fig. 9) is used in which the cathode of the input tube is loaded only by the capacitance of the grids of two additional cathode followers, one of which drives the screen of the input cable, and the other supplies the output. The former has a large transconductance and a frequency range of more than 5 Mc. The over-all frequency range of the feedback is therefore determined by the input tube and is about 1 Mc (Fig. 10). By additional capacitances C_{k1} this frequency range can be reduced to 200 kc and to 80 kc.

The performance of the cathode follower was ascertained by measuring the frequency response and the noise with a 10 MΩ model electrode (inset in Fig. 11). To measure the frequency response a sine wave was applied to the input via a RC-C_k filter. The experiments 1 through 3 (Table I, Fig. 11) demonstrate how the response changes with the frequency range f_i of the feedback path. When f_i is reduced from 1 Mc to 80 kc, the frequency response is transformed from an over-damped ($a \sim 2$) via a critically damped response ($a \sim 1$) to an underdamped response ($a \sim 0.6$). The degree of reduction of input "cathode" capacitance C_f is found by increasing it from 30 $\mu\mu$f in experiment 1 to 130 $\mu\mu$f in experiment 4. The approximate estimation of C_{eq} shows a corresponding increase of 0.9 $\mu\mu$f or 110-fold reduction of input cable capacitance. This is consistent with the amplification of the two CF stages (9).

$$A_0 = A_{01} \times A_{02} \simeq (1 - 3 \times 10^{-3})(1 - 6 \times 10^{-3})$$

$$\simeq 1 - 9 \times 10^{-3}.$$

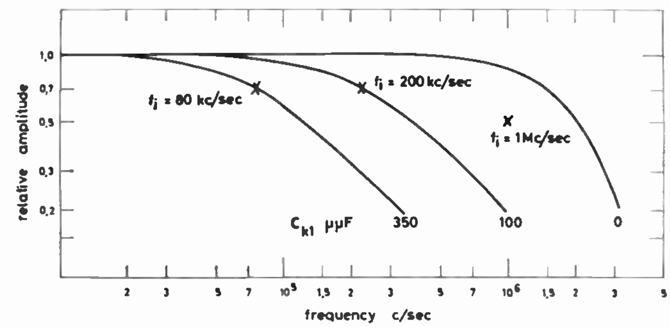

Fig. 10—Frequency responses of the feedback path of the two first stages of the cathode follower in Fig. 9. The cutoff frequency f_i is varied by means of the cathode-ground capacitance C_{k1} of the input cathode follower.

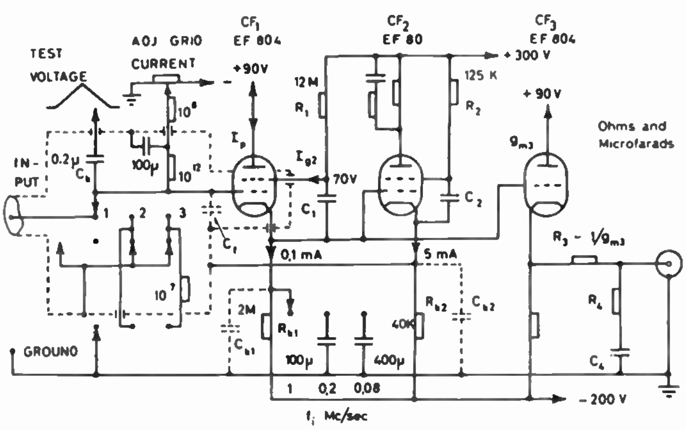

Fig. 9—Three-stage cathode follower with wide frequency range in the feedback path.

CF_1—with low input grid current and small cathode-ground capacitance C_{k1}.

CF_2—with wide-frequency range ($i_a = 5$ ma) to drive the input screen.

CF_3—output cathode follower.

CF_1 and CF_2 operate as pentode-CF at ac and as triode-CF at dc. The gain variation of CF_1 is compensated by the circuit in the output [see (28) and (29)]. The input switch connects the input grid to 1) the electrode, 2) ground, and 3) the 10-MΩ model electrode. Grid current compensation is adjusted with open input.

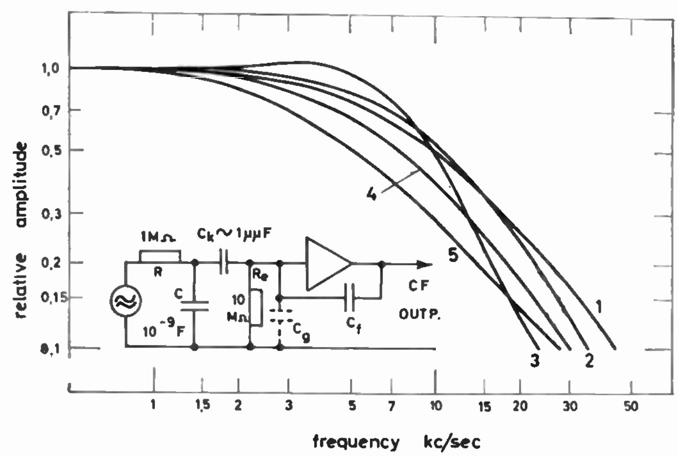

Fig. 11—Frequency responses of the three-stage cathode follower (Fig. 9) with a 10 MΩ resistor as model electrode, measured by way of the small condenser method [Fig. 7(b)]. V_k is a sine wave integrated through the filter RC. The figures on the curves denote the experiment number and refer to the data of Table I. Curve 1 also represents the response through a 10 MΩ micropipette electrode placed in a screened electrode holder and with the tip of the electrode immersed 1 mm.

TABLE I

NEUTRALIZATION OF INPUT CAPACITANCE IN THE 3-STAGE CF IN FIG. 9

Exp. No. (cf., Fig. 11)	Upper Cutoff Freq. of the Feedback Path (Fig. 10) f_i c/sec	Total Input Capacitance $C_f + C_g$ $\mu\mu$f	Normalizing Time Const. (2)* T' μsec	Equiv. Input Time Const. T_{eq}† μsec	Equiv. Input Capacitance (8)* C_{eq} $\mu\mu$f	Damping Factor (7) $a = T_{eq}/T'$	Total Noise e_n, μV rms: Measured at the Response in Fig. 11	Calculated from (17)*	Noise Factor F from (18)*
1	10^6	30	15	(26)	(2.6)	2	78	90	6
2	2.10^5	30	29	30	3.0	1	55	55	2.5
3	8.10^4	30	50	(30)	(3.0)	(0.6)	47	47	1.5
4	10^6	30+100‡	31	(35)	(3.5)	1	135	160	25
5	10^6	30+3‖	15	52	5.2	3.5	54	60	6

* $R_e = 10$ MΩ, $T_i = 1/(2\pi f_i)$, $R_{eq} = 20$ kΩ.

† $T_{eq} = 1/(2\pi f_{eq})$ where f_{eq} is the frequency in Fig. 11 with amplitude 0.7 in exp. 1, 4 and 5 and half the frequency with amplitude 0.5 in exp. 2 and 3.

‡ Added to C_f between input and "cathode."

‖ Added to C_g between input and ground.

Thus, the unavoidable input-cathode capacitance of 30 $\mu\mu$f is reduced to 0.3 $\mu\mu$f and represents only 10 per cent of the equivalent input capacitance C_{eq} of 3 $\mu\mu$f (measured in experiment 2). Therefore the main contribution to C_{eq} is from stray capacitance between input and ground, including the parallel capacitance of the model electrode. A 10 MΩ microelectrode in a screened electrodeholder with only its tip immersed in solution gave a $C_{eq}\sim 3$ $\mu\mu$f, *i.e.*, the same value as found with the 10 MΩ model electrode.

The measured and calculated rms noise of the *CF* were nearly identical (Table II). This confirms the existence of noise factors F larger than 1 indicating an excess noise in cathode followers due to neutralization of the input capacitance. Experiment 1 shows the disadvantage of a too-wide frequency range of the feedback path since the noise factor is larger than 2 [*cf.*, (19a)]; experiment 4 shows the importance of a small total input capacitance [*cf.*, (19b)]. The conditions in experiment 2 represent a good compromise between low noise and a frequency response such as to distort the recorded action potential but little (*cf.*, Fig. 5).

TABLE II

MEASURED TOTAL NOISE e_{nt} OF THE NEGATIVE CAPACITANCE SHOWN IN FIG. 8 WITH THE FREQUENCY RESPONSES OF FIG. 13

Upper Cutoff Frequency (Fig. 12) f_i Mc	e_{nt} μV rms		
	$a=1$	$a=0.5$	$a=2$
4.0	190	280	125
0.8	86	120	60
0.16	44	62	32

The use of *pentode coupling* in the input cathode follower and in the cathode follower which drives the input screen is essential to obtain the low equivalent input capacitance of 3 $\mu\mu$f. In the cathode followers in Fig. 9 pentode coupling is obtained by supplying the screen grid via a resistor and by connecting the screen grid to the cathode by a capacitor. This coupling replaces the battery previously used for this purpose [3], but introduces two minor disadvantages:

1) The resistor in the screen grid is in parallel to R_k and reduces the gain (9).
2) With a dc input the input tube operates as a triode cathode follower, the gain being lower at dc than at higher frequencies.

This difference in gain is compensated by the equalizing circuit in the output. The ratio between the resistors R_3 and R_4 in the equalizing circuit is given by the ratio between the gains of the triode and the pentode couplings. With a large cathode resistor R_{k1} and a large screen grid resistor R_1 a first approximation is

$$\frac{R_4}{R_3} = \mu_{g2g1} \tag{29}$$

where μ_{g2g1} is the amplification factor of the tube as a triode. Assuming the current to the screen grid I_{g2} to be a constant fraction n of the anode current I_p, the equalizer has the time constant

$$(R_3 + R_4)C_4 = \frac{R_1C_1}{1+n} \tag{30}$$

where R_1C_1 is the time constant in the screen grid of the input tube. This equalizing circuit can be adjusted to compensate the difference in gain rather exactly. With a symmetrical input of two pentode cathode followers to a difference amplifier a common mode rejection ratio of about $1:10^5$ has been obtained at high amplification, including the frequency range of transition from triode to pentode gain.

B. The Negative Capacitance

The measurements were performed on the negative capacitance shown in Fig. 8. The stages following the input stage compensate for its poor high-frequency response and provide sufficient gain for the positive feedback. This was obtained by a negative feedback to the cathode of the input tube [5]. A wide frequency range was obtained by a three-stage wide-band amplifier (Fig. 8). The networks for phase and frequency correction of the negative feedback loop could be adjusted stepwise simultaneously with the gain to ensure a flat frequency response at high frequencies (S_1 in Fig. 8). The frequency range of 5 Mc decreased slightly with increasing amplification (Fig. 12). To obtain a small noise factor and to enable neutralization of the capacitance of deeply immersed microelectrodes [18], the frequency range could be reduced to 1 Mc or to 200 kc (S_2 in Fig. 8).

The frequency response for a signal through a 10 MΩ resistor as model electrode was measured by a sine wave applied to the input through a RC-C_k filter (Fig. 13). By adjustment of the neutralization the shape of the frequency response experimentally found was fitted to the theoretical frequency responses (Fig. 2), in the critically damped ($a=1$) the underdamped ($a=0.5$) and the overdamped ($a=2$) condition, at the three frequency ranges of the feedback amplifier (Fig. 12). The rms noise determined at the different frequency responses in Fig. 13 are given in Table II.

At critical damping ($a=1$) the calculated and the measured value of the normalizing time constant T' and of the total rms noise e_{nt} are compared in Table III. Experimental and calculated values agreed at $f_i=0.8$ Mc and 0.16 Mc. At $f_i=4$ Mc the experimentally determined value of T' was larger and the noise was less than the calculated: the values of T' and noise measured at $f_i=4$ Mc were those to be expected at $f_i=2.6$ Mc. The author has no explanation to offer for the discrepancy.

The noise measured when $a=0.5$ and $a=2$ (Table II) differed by a factor 2, and was related to the noise at $a=1$ by $\sqrt{2}$, corresponding to the variation of the rms noise e_{nt} with $a^{-1/2}$ (17).

The increase of the amplifier noise above that from the electrode [given by F (18)] is shown in Table IV. To test the calculations C_f+C_g was increased and T_i was decreased five times. The measured and the calculated noise agree in experiment 1.A and 1.B. In experiment 2.B in which T_i was extremely small the deviation was the same as in Table III when $f_i=4$ Mc. Since T' is not changed from A to B (Table IV) the noise from the electrode resistance is identical (17) and the increase in noise in B originates solely from an increase in the noise of the amplifier input, as indicated by the marked increase in the noise factor.

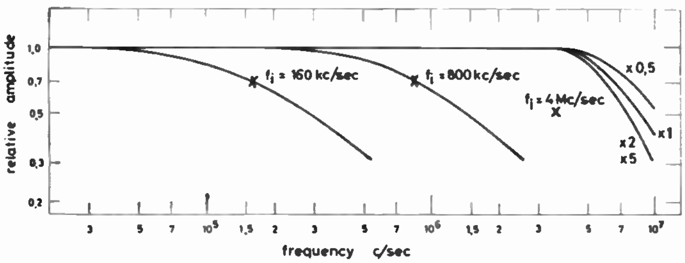

Fig. 12—Frequency responses of the positive feedback path of the negative capacitance in Fig. 8 measured at the "feedback capacitance gains" 0.5, 1, 2 and 5 [A_0-1 in (14)] at the widest frequency range, and at "gain" 2 at the two lower frequency ranges. f_i—the measured upper frequency limits giving the rise times $T_i=1/(2\pi f_i)$ used in Tables III and IV.

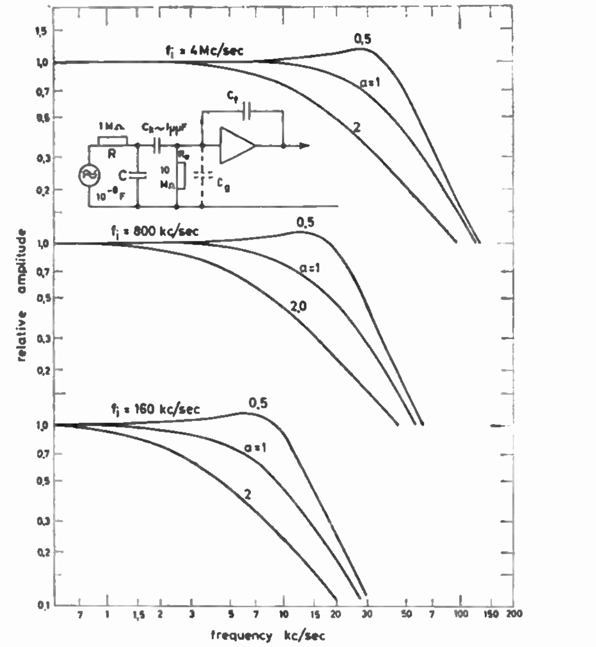

Fig. 13—Frequency responses of the negative capacitance in Fig. 8 with a 10 MΩ resistor as model electrode measured by way of the small condenser [Fig. 7(b)]; V_k is a sine wave integrated through the filter RC. f_i—the frequency ranges of the positive feedback amplifier (from Fig. 12). Neutralization is adjusted to responses which fit those in Fig. 2 at corresponding values of the damping factor a. Measurements of noise at these responses are given in Table II, and a comparison between measured and calculated noise e_{nt} and time constant T' is given in Table III.

TABLE III

MEASURED AND CALCULATED NORMALIZING TIME CONSTANT T' AND TOTAL NOISE e_{nt} OF THE NEGATIVE CAPACITANCE IN FIG. 8 WHEN ADJUSTED TO A CRITICALLY DAMPED RESPONSE ($a=1$)

Upper Cutoff Freq. (Fig. 12) f_i Mc	T' μsec		e_{nt} μV rms	
	Measured on Fig. 13* $a=1$	Calculated from (2)†	Measured (Table II, $a=1$)	Calculated from (17)†
4.0	6	7.5	190	270
0.8	14	15	86	90
0.16	30	30	44	41

* $T'=1/\pi f'$, where f' is the frequency with amplitude 0.5 in Fig. 13, $a=1$.
† $R_e=10$ MΩ, $C_f+C_g=20$ μμf, $T_i=1/(2\pi f_i)$, $R_{eq}=20$ kΩ.

TABLE IV

TOTAL NOISE e_{nt} OF THE NEG. CAPACITANCE IN FIG. 8 AT DIFFERENT VALUES OF C_f+C_g AND T_i WHEN $T'=2\sqrt{R_e(C_f+C_g)T_i}$ IS CONSTANT

Exp. No.	Total Input Capacitance C_f+C_g μμf	Time Constant of the Feedback Amplifier T_i μsec*	Equiv. Input Time Constant T' μsec†	e_{nt} μV rms		Noise Factor F (18)†
				Measured	Calculated (17)†	
1. A	20	1.0	30	44	41	1.4
1. B	110	0.2	30	100	110	11
2. A	20	0.2	14 (15)	86	90	3
2. B	110	0.04	14 (15)	240	380	51

* $T_i=1/(2\pi f_i)$, where f_i is from Fig. 12.
† $R_e=10$ MΩ, $R_{eq}=20$ kΩ, $a=1$.
Frequency response: in experiment 1 is as Fig. 13 lower curve $a=1$, in experiment 2 as Fig. 13 middle curve $a=1$.

VIII. Discussion and Conclusions

This report describes an analysis of the cathode follower (*CF*) and a similar analysis of the negative capacitance (*NC*) with respect to neutralization of input capacitance, noise and grid current.

As for neutralization of input capacitance, the *NC* is generally considered superior to the *CF* since the *NC* can neutralize a capacitance C_g between input and ground whereas the *CF* only reduces the capacitances C_f connected to its cathode. Whether a *NC* or *CF* is more suitable depends on the design of the circuits and on the experimental situation. When only the tip of the microelectrode is immersed the equivalent input capacitance C_{eq} is about 3 $\mu\mu$f both with one of the commonly used *NC* circuits [5], [7]–[9] and with the *CF* circuit described in this report. The *CF* has the advantage that adjustment of neutralization is unnecessary and that there are no self-oscillations, which occur in a *NC* when the neutralization is adjusted to compensate too large a capacitance.

With the *NC* a C_{eq} of less than 1 $\mu\mu$f can be obtained when the frequency range of the positive feedback is extended to several megacycles. As long as only the very tip of the microelectrode is immersed in the specimen it is only the high-frequency *NC* which is advantageous as compared with the *CF*. On the other hand when the microelectrode is deeply immersed in the specimen it contributes about 10 $\mu\mu$f to the input-ground capacitance and only a negative capacitance can reduce this amount of input-ground capacitance. A negative input capacitance whose positive feedback amplifier has a bandwidth of a few hundred kilocycles is sufficient for this purpose. A higher degree of neutralization is not feasible because the capacitance of the deeply immersed electrode is mixed with a small portion of the electrode resistance in a distributed manner [18].

A value of 0.2–0.3 $\mu\mu$f has been suggested as the proper equivalent capacitance of an input circuit [7], [8]. At present it is hardly possible to obtain such values. It is also questionable if the electrode properties allow the use of such small values of C_{eq} [18]. An estimation of the relation between the error on the action potential and the value of the equivalent input capacitance has shown that a C_{eq} of about 1 $\mu\mu$f is sufficient for nearly all types of intracellular measurement of action potentials. With this C_{eq} the error at the maximum rate of rise of an action potential is less than 10 mV. The current which passes through the cell under these conditions does not change the state of the cell membrane even when recording from small cells (toad motor neuron) with a "total membrane resistance" of 5 MΩ [14]. This assumption seems justified because the current causes a voltage drop over the membrane of less than 2 mv.

In a negative input capacitance it is customary to adjust neutralization to the so-called critically damped response. The critically damped response is of advantage in voltage clamp experiments [19] in which the potential level of the membrane is changed in a controlled way. When action potentials are recorded across the membrane it is possible to reduce errors on the recorded potentials by using an oscillatory response. There are two types of error: 1) an error proportional to the first derivative of the action potential and 2) an error proportional to the second derivative of the action potential. Due to the absence of discontinuities in the first derivative of the action potential, the error proportional to the second derivative is small relative to the error proportional to the first derivative, and the first derivative error is reduced as effectively by an oscillatory response ($a<1$) as by extension of the frequency range of the feedback amplifier. Thus, when recording action potentials the oscillatory response gives an effective reduction of C_{eq} without extension of the feedback amplifier frequency range.

A high degree of neutralization in a *CF* may be associated with an oscillatory response [3], [12a], *i.e.*, a response with a damping factor of less than 1. The author has calculated how the damping factor depends on the constants of the *CF* when the residual input-ground capacitance C_g is taken into account. This calculation has shown that a given small value of C_5 requires a correspondingly wide frequency range of the feedback path in the *CF* to avoid the oscillatory response. A critically damped response with $C_g \sim 3$ $\mu\mu$f is obtained with a frequency range of several hundred kilocycles, which requires a small cathode-ground capacitance of the input *CF*. This has been obtained by introducing two additional *CF*'s which separate the capacitance to ground of the output cable and of the screen of the input cable from the cathode of the input *CF*.

It has previously been suggested [4]–[6] that measurement of the equivalent input capacitance C_{eq} via a series input resistor may be influenced by the stray parallel capacitance of the resistor. This method gives a value of C_{eq} which is four times too small if C_{eq} is of the order of 1 $\mu\mu$f [18]. This explains why some investigators [5], [7] have found a smaller value of C_{eq} than to be expected from the parameters of their circuits. Among the methods suggested to overcome this difficulty [6], [9], [10] the only suitable one is testing via a small condenser [9]. Testing with a triangular wave to the small condenser charges the interior of the biological specimen. The author has replaced the triangular wave by a rectangular, thereby avoiding a final current through the cell.

When the input capacitance is neutralized the feedback causes an increase in noise of the amplifier input. Hence the summated noise from the amplifier and the electrode is greater than the noise from the electrode alone. The excess noise is present both in the *CF* and in the *NC* and is enhanced by the wide frequency range necessary to reduce C_{eq}. It is, however, possible to reduce the input capacitance to about 1 $\mu\mu$f (at critically damped response) without appreciable excess of noise. With $C_{eq} \sim 1$ $\mu\mu$f, obtained with a low-capacitance, low-noise amplifier input, the noise factor is less than 2, *i.e.*, the noise from the electrode is still larger than that from the amplifier input. The corresponding frequency range of the feedback amplifier is about 1 Mc. A further reduction of C_{eq} by extending the frequency range of the feedback amplifier is associated with a noise factor larger than 2, whereas a C_{eq} of less than 1 $\mu\mu$f, obtained by the use of an oscillatory response, preserves a noise factor of 2.

The grid current can be reduced to 10^{-13} A or less by the use of electrometer tubes [8], [9] or by reduction of the anode current in common tubes to less than 10^{-5} A. The disadvantages of these methods, increased noise and microphony and diminished frequency range of the feedback amplifier, can be largely avoided by compensating the grid current via a large resistor.

Appendix I

Transfer Functions and Time Responses

The transfer functions used in this report are special cases of the general transfer function for the circuit in Fig. 6.

The amplifier has the low-frequency gain A_0 and is characterized by the single time constant T_i, thus

$$A' = \frac{A_0}{1 + sT_i} \cdot \tag{31}$$

There is a single feedback capacitance C_f and a single input-ground capacitance C_g'. The output V_0' is calculated for the following three input voltages present simultaneously:

1) A voltage v_e applied to the electrode R_e with the parallel capacitor C_p.
2) A voltage v_k applied to the capacitor C_k.
3) A voltage e_i in the amplifier input which is an equivalent measure of the noise originating in the amplifier.

The voltage v_i between amplifier input and ground is the resultant of the superposition of the signals 1, 2 and 3. If the Laplace transforms of the above time functions are V_e, V_k, E_i and V_i the current equation for the node at the amplifier input is

$$(V_k - V_i)sC_k + (V_e - V_i)(1/R_e + sC_p) + (-V_i)sC_g' + (V_0' - V_i)sC_f = 0. \tag{32}$$

Furthermore

$$V_0' = A'(V_i + E_i). \tag{33}$$

Eliminating V_i and A by introducing (31) and (33) in (32) and solving for $V_0 = V_0'/A_0$, *i.e.*, the output referred to unity gain, gives

$$V_0 = \frac{V_e + V_e sR_eC_p + V_k sR_eC_k + E_i(1 + sR_e(C_f + C_g))}{1 + saT' + s^2(\frac{1}{2}T')^2} \tag{34}$$

where a and T' in the denominator are given by

$$aT' = R_e(C_g + C_f - A_0C_f) + T_i, \tag{35}$$

$$T' = 2\sqrt{R_e(C_f + C_g)T_i}. \tag{36}$$

C_g in (34)–(36) is

$$C_g = C_g' + C_k + C_p \tag{37}$$

which shows that the neutralization concerns the total input-ground capacitance C_g, which consists of all the capacitances connected to the input whether grounded or connected to an input voltage. Thus for a given set of the circuit constants (A_0, T_i, C_f, C'_g, C_k, C_p and R_e) the value of a and T' applies to outputs from all signals V_e, V_k and E_i.

With the signal V_e applied to a pure resistance R_e as electrode, the transfer function is found from (34) when V_k, E_i, $C_p = 0$,

$$\frac{V_0}{V_e} = \frac{1}{1 + saT' + s^2(\frac{1}{2}T')^2} \cdot \tag{38}$$

The corresponding amplitude frequency characteristic is obtained by replacing s with $j\omega$ and by determining the numerical value

$$\frac{V_0}{V_e} = \left[\frac{1}{1 + (a^2 - \frac{1}{2})(\omega T')^2 + \frac{1}{16}(\omega T')^4}\right]^{1/2} \cdot \tag{39}$$

The frequency characteristics for different values of a with $\omega T'$ as the variable are given in Fig. 2.

When $v_e(t)$ is a step voltage of amplitude A volt, the output $v_0(t)$ in volts is obtained from a table of Laplace transforms (*e.g.*, no. 51 in Pipes [19]). According to the value of a the time course of v_0 takes three different forms:

1) for $a = 1$

$$v_0(t) = \left[1 - \left(1 + \frac{2t}{T'}\right)\exp\left(-\frac{2t}{T'}\right)\right]A, \tag{40}$$

2) for $a > 1$

$$v_0(t) = \left[1 - \tfrac{1}{2}(a^2 - 1)^{-1/2}\left[\frac{1}{u}\exp\left(-u\frac{2t}{T'}\right) - \frac{1}{v}\exp\left(-v\frac{2t}{T'}\right)\right]\right]A,$$

where $u = a - (a^2 - 1)^{1/2}$ and $v = a + (a^2 - 1)^{1/2}$ (41)

3) for $a < 1$

$$v_0(t) = \left[1 - (1 - a^2)^{-1/2} \cdot \sin\left[(1 - a^2)^{1/2}\frac{2t}{T'} + \phi\right]\exp\left(-a\frac{2t}{T'}\right)\right]A,$$

where $\tan \phi = (a^{-2} - 1)^{1/2}$. (42)

The outputs for different values of a as function of t/T' are given in Fig. 3.

When $v_e(t)$ is an impulse voltage the output is the derivative of (40)–(42), since the Laplace transforms of the unit step and the unit impulse functions differ by a factor s. Thus when the impulse voltage has the strength I volt×sec the output is in volts:

1) for $a=1$

$$v_0(t) = \frac{1}{T'}\frac{t}{T'}\exp\left(-a\frac{2t}{T'}\right)I, \tag{43}$$

2) for $a>1$

$$v_0(t) = \frac{1}{T'}(a^2-1)^{-1/2} \cdot\left[\exp\left(-u\frac{2t}{T'}\right)-\exp\left(-v\frac{2t}{T'}\right)\right]I, \tag{44}$$

where u and v is as in (41),

3) for $a<1$

$$v_0(t) = \frac{1}{T'}2(1-a^2)^{-1/2} \cdot\exp\left(-a\frac{2t}{T'}\right)\sin\left[(1-a^2)^{1/2}\frac{2t}{T'}\right]I. \tag{45}$$

The outputs for different values of a as function of t/T' are given in Fig. 4.

Appendix II

The Noise

The output noise originating from the microelectrode and the amplifier is calculated from

$$\overline{e_n^2} = \int_{-\infty}^{+\infty}|H(j2\pi f)|^2S(f)df$$

where $\overline{e_n^2}$ is the mean square of the output noise when an input signal with power density spectrum $S(f)$ is applied to a linear system having a transfer function $H(j2\pi f)$. This expression can be evaluated by means of contour integration. It is convenient to change the variable of integration to the complex frequency s. Hence the above integral can be written as

$$\overline{e_n^2} = \frac{1}{2\pi j}\int_{-j\infty}^{+j\infty}H(s)H(-s)S(s)ds$$

where the path of integration is along the imaginary axis. If

$$s[H(s)H(-s)S(s)]\to 0 \qquad \text{for } s\to\infty \tag{46}$$

the last integral can be replaced by the contour integral

$$\overline{e_n^2} = \frac{1}{2\pi j}\int_C H(s)H(-s)S(s)ds$$

where C is the infinite semicircle of the left half plane. From Cauchy's theorem follows

$$\overline{e_n^2} = \sum \text{Res}\,[H(s)H(-s)S(s)] \tag{47}$$

where the residues are evaluated at the poles in the left half of the complex s plane.

Use of (47) is made in the following cases:

1) The normal noise voltage from a resistance R_1 transmitted through a network consisting of a single time constant T_1. In this case

$$H(s) = 1/(1+sT_1), \qquad S(s) = 2kTR_1, \tag{48}$$

where k is Boltzmann's constant and T is the temperature in °K. Since the condition (46) is fulfilled, the mean-square noise (47) is

$$\overline{e_n^2} = \text{Res}\left[\frac{1}{1+sT_1}\frac{1}{1-sT_1}2kTR_1\right] = \left[\frac{1}{T_1}\frac{1}{1-sT_1}2kTR_1\right]_{s=-1/T_1} = \frac{kTR_1}{T_1}$$

when $kT=4\times10^{-21}$, ($T=290$°K), is

$$\overline{e_{n1}^2} = 2.5\times10^{-20}Rf_1 \tag{49}$$

where $f_1=1/(2\pi T_1)$ is the frequency at the amplitude 3 db down of the frequency response to (48).

2) The effect of neutralization on the mean-square noise of the amplifier input. In this case the appropriate transfer function of the amplifier input noise is obtained from (34) when $V_e=V_k=0$:

$$H(s) = \frac{V_0}{E_i} = \frac{1+sR_e(C_f+C_g)}{(\frac{1}{2}T')^2(s-s_1)(s-s_2)} \tag{50}$$

where s_1 and s_2 are the poles

$$(s_1,s_2) = (2/T')(-a\pm\sqrt{a^2-1}). \tag{51}$$

The appropriate power density spectrum is

$$S(s) = 2kTR_{eq} \tag{52}$$

where R_{eq} is the equivalent noise resistor of the neutralizing input capacitance, found by measuring its mean-square noise $\overline{e_{n1}^2}$ with the input short circuited. When $\overline{e_{n1}^2}$ is measured with an amplifier characterized by a single time constant (48) R_{eq} is obtained from (49).

We thus have the residue (47) from (50) and (52)

$$\text{Res}\,(s_1) = \left[\frac{1-s^2(R_e(C_f+C_g))^2}{(\frac{1}{2}T')^4(s+s_1)(s^2-s_2^2)}2kTR_{eq}\right]_{s=s_1}$$

and the analog residue for $s=s_2$. Summating the residues, inserting s_1 and s_2 from (51) and using (36) gives the mean-square noise

$$\overline{e_{ni}^2} = \frac{kTR_{eq}}{aT'}\left[1+\frac{R_e(C_f+C_g)}{T_i}\right] \tag{53}$$

3) The mean-square noise $\overline{e_{nr}^2}$ from the electrode resistance R_e when the noise is transmitted through the neutralized input capacitance. The proper transfer func-

tion $H(s)$ is (38) and the power spectrum is $S(s)=2kTR_e$. Evaluation of (47) gives

$$\overline{e_{nr}^2} = \frac{kTR_e}{aT'}. \tag{54}$$

By summating (53) and (54) the total noise from the electrode and the neutralized input capacitance is obtained

$$\overline{e_{nt}^2} = \frac{kTR_e}{aT'}\left[1 + \frac{R_{eq}}{R_e} + \frac{R_{eq}(C_f + C_g)}{T_1}\right] \tag{55}$$

where $kT = 4\times 10^{-21}$ at $T = 290°$K.

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Stabilized Wide-Band Potentiometric Preamplifiers*

J. W. MOORE†, SENIOR MEMBER, IRE, AND J. H. GEBHART‡

Summary—The need for a wide-band preamplifier to measure bioelectric potentials from a source with high internal resistance and shunt capacitance has stimulated the development of instruments which should be generally useful. Consideration is given to the simultaneous attainment of such factors as low grid current, linearity, low drift, and improved dynamic response by compensation for input capacitance. An analysis of a simplified circuit is augmented by the use of an analog computer to simulate the system for a preamplifier having either a first- or second-order response.

Some examples of circuits suitable for bioelectric measurements are described. It has been possible to chopper-stabilize such preamplifiers against drift without significant degradation of their high input impedance characteristics.

I. Introduction and Design Requirements

THE REQUIREMENT of a wide-band electrometer, stabilized against drift, arose in the development of a technique for controlling the potential across the membrane of the squid axon[1] as measured with a micropipette electrode. It has been well established that a glass micropipette, drawn to a fine tip and filled with potassium chloride solution, may be used to impale some cells without injury.[2,3] Fig. 1 shows electron microscope photographs of typical micropipette tips. The KCl filled micropipette may be considered an electrolyte bridge from the interior of the cell to some external nonpolarizable metal electrode connected to an amplifier. Most of the resistance of such a salt bridge is concentrated within a few hundredths of a millimeter of the tip[4] and may be as large as 50 to 100 megohms even when a nearly saturated electrolyte (3 molar KCl) solution is used. Intracellular resting and action potentials, both in the neighborhood of 100 mv, have been measured in many excitable tissues with this technique.[2–8]

There are two requirements for low-input stage grid current. In the first place, the current through the electrode must be small enough to avoid an appreciable IR drop in the tip; for such an effect to be less than 0.5 mv in a 50 megohm tip, the grid current must be less than 10^{-11} a. A few millivolts of IR offset might be tolerated were it not for the possibility of errors encountered when the resistance is changed by the tip being plugged or broken; an additional difficulty arises from the fact that the passage of current can change the tip resistance.[9] The second limitation on the grid current is that the current flow into the interior of the cell through the tip should not be enough to introduce an appreciable change in the potential of the membrane under measurement. This depends on both the effective area and permeability of the cell membrane. Some investigators[10,11] have found that currents of only about 10^{-11} a

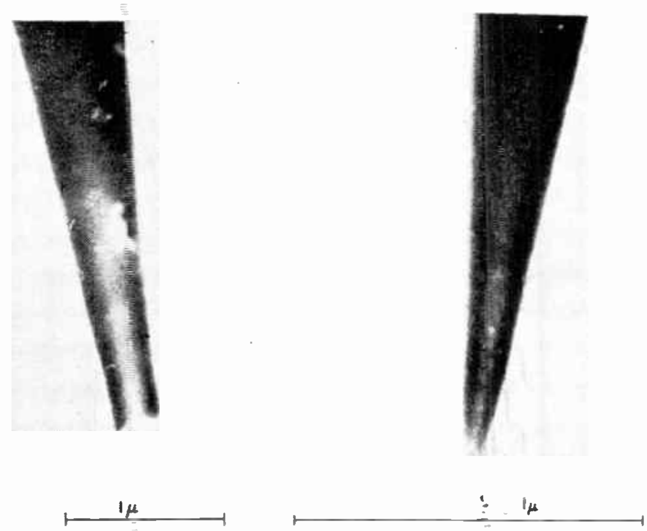

Fig. 1—Electron microscope photographs of ends of glass micropipettes; left, machine drawn tip of about 0.3 micron O.D.; right, hand drawn tip of less than 0.1 micron O.D.

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† Department of Physiology, Duke University Medical Center, Durham, N. C., and a National Neurological Research Foundation Scientist. Formerly with National Institutes of Health, Bethesda, Md.

‡ Wesley Theological Seminary, Washington, D. C. Formerly with National Institutes of Health, Bethesda, Md.

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are large enough to alter the electrical excitability of nervous tissue. For these reasons, it seems necessary to limit the grid current to the 10^{-11} to 10^{-12} ampere range.

The achievement of such a low grid current for an amplifier makes accurate resting potential measurements possible, but the need for faithful reproduction of action potentials (lasting only a few tenths of a millisecond) imposes a further requirement on the measuring system. The capacitance across the glass wall of the micropipette (between the inside KCl solution and the extracellular electrolyte) has been calculated and found to be in the order of 1 pf/mm of tip immersed.[4] Usually the cell surface must be covered with 1–3 mm extracellular electrolyte which is connected to the ground lead of the amplifier. Woodbury[4] showed that the micropipette in Fig. 2(a) could be approximated by the distributed equivalent circuit, Fig. 2(b), or the lumped circuit in Fig. 2(c) with reasonable accuracy. The attenuation of the high frequency components of the signal by the capacitance across the micropipette wall in parallel with the amplifier input capacitance may be equalized by proper high frequency compensation following the output.[4] However, this method depends on a known and constant tip resistance; it has already been shown that this cannot be assured. MacNichol and Wagner[12] pointed out that, even when the dc grid current has been made very small, a capacitive current proportional to any rate of change of the input signal will pass through the tip; this current amounts to 10^{-12} a per picofarad per volt per second. The rate of change of bioelectric action potentials may exceed a kilovolt per second and give rise to a tip current[13] of 10^{-9} a per picofarad of input capacitance. This capacitive current would cause a voltage drop of 100 mv across a 20 megohm tip for an input capacitance of only 5 pf. Not only would such a current give an intolerable distortion but it might also be sufficient to change the membrane potential of the cell under measurement.

Still more stringent requirements must be met for the increasingly used "voltage clamp" experiments.[1,14] The micropipette signal not only monitors the membrane potential but, in addition, it is fed back to a control amplifier so that the membrane is forced to follow potential steps with rise times of a few microseconds. For a 100 mv change in potential to occur in 10 μsec, the tip capacitive current would be increased to 10^{-8} a/pf. This places a premium on the quality of input capacity neutralization. In addition, it is desirable that the response to a voltage step at the micropipette tip be as

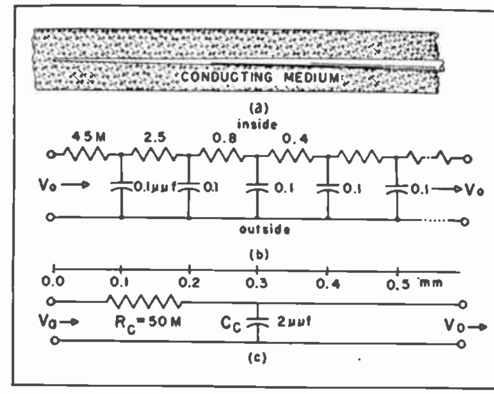


Fig. 2—(a) Scale drawing of microelectrode tip (0.5 mm long) in electrolyte solution. (b) Approximate equivalent circuit for this micropipette filled with 3 m KCl. (c) Lumped circuit approximating the distributed circuit for a 2 mm length in external solution (reproduced, courtesy of J. W. Woodbury[4] and *J. Cellular Comp. Physiol.*).

fast and smooth as possible, so that the over-all control loop rise time can be made small. Furthermore, it is necessary to stabilize the preamplifier because any drift in its baseline would be indistinguishable from a signal and produce an undesired and unknown change in the steady membrane potential.

II. Previous Developments

A number of investigators have tackled the problem of recording faithful reproductions of membrane action potentials from micropipettes and have developed some sophisticated and elegant circuits.

Nastuk and Hodgkin[3] first measured intracellular action potentials by means of a cathode follower with the shield around the input lead driven from the cathode to reduce the effect of the input cable capacity. They used a 6AK5 tube at reduced electrode potentials and selected for low grid current.

Solms, Nastuk and Alexander[15] cascaded two cathode followers to increase the input impedance. Krakauer[16] reduced the "Miller capacity" of the input tube by causing the plate as well as the cathode to follow the voltage changes on the grid in his "Electrometer Triode Follower." Lettvin, Howland and Gesteland[17] have also described rather similar circuits. Bak[18] designed a cathode follower whose gain could be set to unity by adjustment of the in-phase signal fed to the screen grid from an auxiliary amplifier of greater than unity gain. A somewhat similar circuit in which the plate of the input triode is driven with the same excursion as on the

[12] E. F. MacNichol, Jr. and H. G. Wagner, "A High Impedance Input Circuit Suitable for Electrophysiological Recording from Micropipette Electrodes," Nav. Res. Inst., Bethesda, Md., vol. 12, pp. 97–118, Rept. No. 7; 1954.

[13] $I_{tip} = C dv/dt$, where C is the total effective input capacitance and V is the voltage across C.

[14] A. L. Hodgkin, A. F. Huxley, and B. Katz, "Measurement of current-voltage relations in the membrane of the giant axon of Loligo," *J. Physiol.*, vol. 116, pp. 424–448; April, 1952.

[15] S. J. Solms, W. L. Nastuk, and J. T. Alexander, "Development of a high-fidelity preamplifier for use in the recording of bioelectric potentials with intracellular electrodes," *Rev. Sci. Instr.*, vol. 24, pp. 960–967; October, 1953.

[16] S. Krakauer, "Electrometer triode follower," *Rev. Sci. Instr.*, vol. 24, pp. 496–500; July, 1953.

[17] J. Y. Lettvin, B. Howland, and R. C. Gesteland, "Footnotes on a headstage," IRE Trans. on Medical Electronics, vol. ME-10, pp. 26–28; March, 1958.

[18] A. F. Bak, "A unity gain cathode follower," *Electroencephalog. and Clin. Neurophysiol. J.*, vol. 10, pp. 745–748; November, 1958.

grid has also been described by Haapanen and Ottoson.[19] Macdonald[20] has independently developed very high input impedance circuits with wide dynamic range for Hall-effect measurements and high quality audio isolation stages.

Bell[21] developed and described a "negative-capacity amplifier" for use with video signals. He fed back an amplified in-phase signal to the input grid through a small capacitor to neutralize the effect of input capacitance.[22] Several of the designers noted above provided for such neutralization by additional auxiliary amplifiers to boost the signal gain above unity. MacNichol and Wagner[12] have used Bell's circuit configuration with a conventional receiving tube, operating at low electrode potentials, as an input electrometer element followed by a gain package with negative feedback to the input cathode to stabilize the net gain at 3. Woodbury[23] and Amatniek[24] have also developed essentially equivalent circuits.

There has been considerable difficulty and some misunderstanding in establishing an appropriate test and measure of the response of the preamplifier and micropipette system. Some workers have estimated the circuit time constant by applying a square wave at an electrolyte solution into which the microelectrode was dipped. This procedure gives a deceptively fast response because the capacitance of the electrode tip is in parallel with the source resistance as in Fig. 3(a). When the micropipette has been made to just penetrate a cell membrane (the bioelectric potential generator) the outside of most of the glass wall is in contact with the grounded exterior solution and the equivalent circuit is approximately that in Fig. 3(b). Nastuk and Hodgkin[3] recognized this and tested their system response by driving a bath which the microtip just touched but placed a small drop of electrolyte solution (on a grounded loop of silver wire) around the micropipette just above the surface of the bath. This technique is approximately simulated by the circuit of Fig. 3(b) but it is tedious, time consuming, and does not allow the operator to monitor the response during or after penetration. If the tip is partially occluded or broken, the response time will change from the originally measured value.

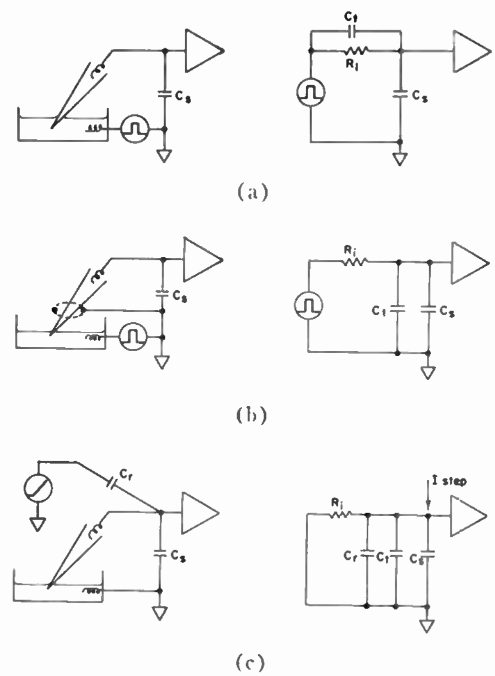


Fig. 3—Methods of testing micropipette—preamplifier response and corresponding equivalent circuits. (a) Test pulse drives solution into which tip dips. (b) Same, but with grounded solution just above tip. (c) Test current pulse injection by means of small capacitor and ramp voltage.

The most effective method for injection of a current step to the grid for monitoring the system response is by application of a short, steep voltage "ramp" (*i.e.*, a voltage which changes linearly with time) to a small capacitor connected to the preamplifier grid as in Fig. 3(c). If the slope of the voltage ramp is considerably in excess of the rate of change of potential on the grid, a step of current will be injected into the system at the grid. The feedback capacitor may be increased to include compensation for the additional "ramp capacitor," which is in parallel with the other capacities to ground. With this technique, the micropipette tip resistance may be continuously monitored and the system response adjusted before, during, and after penetration of the cell membrane, and the response to a test pulse recorded along with each bioelectric response. The superiority of this method of Lettvin, Howland, and Gesteland[17] arises from the fact that a small capacitance can be a very pure element, *i.e.*, its resistance can be essentially infinite, particularly if a small air capacitance is used.[25] In contrast, a high value of resistance is not a pure element because of associated longitudinal and distributed capacitances (to a nearby shield), and attempts to inject current steps through such elements lead to ambiguous results.[12] A photomultiplier "current source" was found to have a minimum dark current of

[19] L. Haapanen and D. Ottoson, "A frequency compensated input unit for recording with microelectrodes," *Acta Physiol. Scand.*, vol. 32, pp. 271–280; 1954.

[20] J. R. Macdonald, "An ac cathode-follower circuit of very high input impedance," *Rev. Sci. Instr.*, vol. 25, pp. 144–147; 1954. Also, J. R. Macdonald, "Some augmented cathode follower circuits," IRE Trans. on Audio, vol. AU-5, pp. 63–70; May–June, 1957.

[21] P. R. Bell, "Cathode-compensated amplifier," in "Waveforms," M.I.T. Rad. Lab. Ser., Cambridge, Mass., vol. 19, Appendix B, 1949.

[22] Although this principle is old and widely employed in "Neutrodyne" radio circuits in the 1920's, Bell's publication has served to point out its usefulness in instrumentation.

[23] J. W. Woodbury, "Recording central nervous activity with intracellular ultramicroelectrodes: Use of negative-capacity amplifier to improve transient response," *Federation Proc.*, vol. 12, p. 159; March, 1953.

[24] E. Amatniek, "Measurement of bioelectric potentials with microelectrodes and neutralized input capacity amplifiers," IRE Trans. on Medical Electronics, vol. ME-10, pp. 3–14; March, 1958.

[25] Guld has suggested that the application of a short voltage pulse to C_r might be less injurious to the cell than the continuing current introduced by a voltage ramp. He gives examples of the output waveforms in response to such test pulses in "Medical Electronics," *Proc. Second Internat'l Conf. on Medical Electronics*, Iliffe and Sons, Ltd., London, England; p. 28, 1959.

10^{-9} a, several orders of magnitude too large for intracellular work.[26]

We have adopted the MacNichol and Wagner circuit configuration for purposes of analysis and further development because of its simplicity in both concept and design. We have been concerned with stabilizing such a preamplifier against drift and improving its speed of response (as far as expedient) for use in active membrane voltage control experiments.

III. Circuit Analysis

Analyses of the response of capacity-neutralized amplifiers have been made,[26,27] assuming lumped input circuit constants and an amplifier with a single time constant response to a low impedance source step. A rather similar analysis will be developed here and includes an input shield capacitance driven by the negative feedback point.

A schematic diagram of the circuit (neglecting dc potentials) to be treated is shown in Fig. 4. The micropipette input resistance is represented by R_i while the capacitances of the signal grid to plate[28] (or screen grid) of the tube, micropipette, wiring, and ramp input are lumped in C_i. The capacity of an input shield which is driven at approximately unity gain from the negative feedback point may be lumped in parallel with the grid-to-cathode capacity C_{gk} of the input tube. Approximate "neutralization" of the input capacitance (or "anti-capacity feedback") may be accomplished by proper adjustment of a feedback capacitor C_{fb}. The negative feedback network is assumed to be composed of resistors whose values are low enough to make the voltage contributions of the currents through the input tube and the C_{gk} capacitor negligible. In addition, the phase shift caused by stray capacitances normally associated with such a network is assumed negligible at frequencies passed by the amplifier.

By Kirchhoff's law, the sum of the currents to the amplifier input grid will be zero when the grid current is negligible and the following equation may be written:

$$\frac{E - V}{R_i} - C_i pV + C_{fb}p(V_0 - V) + C_{gk}p(\beta V_0 - V) = 0, \quad (1)$$

where the operator p is equivalent to d/dt. Upon rearranging,

$$E - V - \tau_i pV + \tau_{fb}p(V_0 - V) + \tau_{gk}p(\beta V_0 - V) = 0,$$

[26] C. C. Yang, J. P. Hervey, and P. E. Smith, "On amplifiers used for microelectrode work," IRE Trans. on Medical Electronics, vol. ME-10, p. 25; March, 1958.

[27] W. H. Freygang, Jr., "A derivation of the input-output relations in a negative capacity preamplifier," *Aux. Pub. Serv.*, Amer. Doc. Inst., Library of Congress, Washington, D. C.

[28] The voltage gain of the electrometer-operated input tube will usually be low (often 4 or less) so that the Miller effect may not be the dominant factor. This capacitor lumping will introduce an error at high frequencies where the plate lags the grid. In several circuits, the plate or screen is driven by a voltage in phase with (and of the same magnitude as) the input signal, and the Miller effect is eliminated.

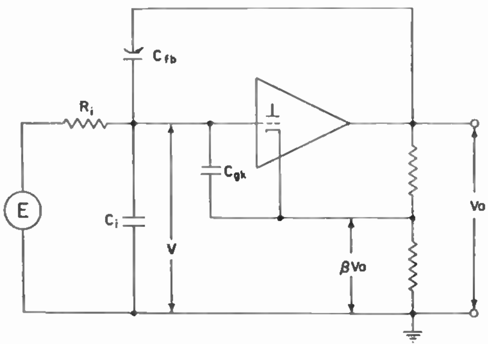


Fig. 4—Schematic circuit diagram of capacitance-neutralizing preamplifier. Conventional negative feedback is used around a gain package symbolized by the triangle; the output has the same polarity as the input signal.

where

$$\tau_i = R_iC_i, \qquad \tau_{fb} = R_iC_{fb}, \qquad \tau_{gk} = R_iC_{gk}.$$

If the response of the amplifier package (with negative feedback having a closed loop gain of A_c at dc) to a zero impedance source potential is assumed to be first order, we have

$$\frac{V_0}{V} = \frac{A_c}{1 + p\tau_a} \quad (2)$$

where τ_a is the closed loop time constant of the amplifier.

Substitution of (2) into (1) gives

$$\frac{V_0}{E} = \frac{A_c}{p^2 + 2\xi\omega_n p + \omega_n^2} \quad (3)$$

where

$$\omega_n^2 = \frac{1}{\tau_a(\tau_i + \tau_{fb} + \tau_{gk})} = \frac{1}{\tau_a R_i(C_i + C_{fb} + C_{gk})} \quad (4)$$

and

$$\xi = \frac{\tau_a + \tau_i + \tau_{fb}(1 - A_c)}{2\sqrt{\tau_a(\tau_i + \tau_{fb} + \tau_{gk})}}, \text{ the damping factor.} \quad (5)$$

Eq. (3) is of the same form as those often encountered in the description of the performance of servomechanisms;[29] the transient response is well known and is usually plotted in terms of dimensionless time ($\omega_n t$) for various values of ξ as in Fig. 5. When $\xi = 1$ the system is critically damped and the solution of (3) is given by

$$\frac{V_0}{E} = A_c[1 - (1 + \omega_n t)e^{-\omega_n t}]. \quad (6)$$

The value of the feedback capacitor C_{fb} to exactly neutralize an input capacitor C_i can be easily determined for the case of a scalar amplifier, *i.e.*, one of infinite bandwidth at constant gain. The input capacitor current, of magnitude $C_i pV$, may be supplied via the feedback capacitor and is given by the expression

[29] G. S. Brown and D. P. Campbell, "Principles of Servomechanisms," John Wiley and Sons, Inc., New York, N. Y.; 1948.

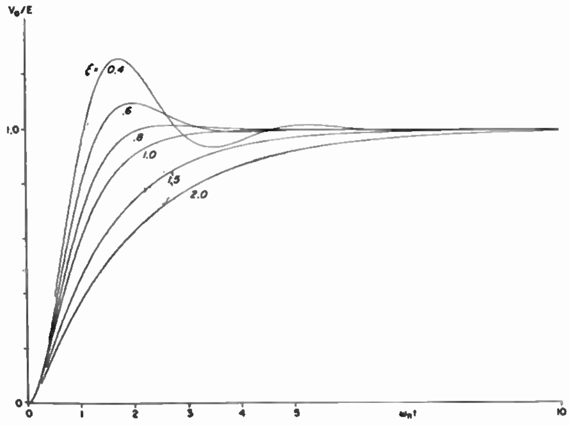


Fig. 5—Family of second-order system responses to a step input for various values of the damping coefficient ξ.

$C_{fb}p(V_0 - V)$. For a scalar amplifier $V_0 = A_c V$ and the proper setting of the feedback capacitor would be

$$C_{fb} = \frac{C_i}{A_c - 1}. \tag{7}$$

In practice, this value of C_{fb} will generally cause a real system in which A has limited bandwidth to ring or oscillate; a somewhat smaller value must be used to obtain a monotonic or a slightly underdamped response. Because C_{fb} is imbedded in the expressions for both ω_n and ξ, it is more convenient to study the effect of C_{fb} on the system response with an analog computer (see Fig. 9) than to develop the analysis further.

Some discussion of the limiting factors in this method of capacitive compensation method seems appropriate at this point. First, it is important to observe that all of the capacitances connected to the grid, including the compensating feedback one, enter with equal weighing into the denominator of the expression for ω_n, the factor which determines the speed of response. Therefore, attention should be given to the problem of minimizing each of these capacitances. The microtip and holder capacity are usually determined by mechanical and physiological considerations and range up to 10 pf or more for deep tissue penetrations. We have seen that the maximum value of the feedback capacity for best compensation varies directly with C_i and inversely with the closed loop amplifier gain. Intracellular recordings, of about 0.1 v full scale, may often be made with satisfactory signal-to-noise ratios without an input shielded cable in areas where strong stray fields have been minimized. If an input shield is required, its capacitance (lumped into the C_{gk} term) limits the maximum speed of response in much the same manner as microtip or stray capacity to ground. For the case where the amplifier time constant τ_a is very small compared to that of the input circuit, it is possible to ground the shield and increase the compensating capacity slightly to give a response which is almost identical to that in which the shield is connected to the cathode (see Fig. 10).

The procedure yielding the minimum total capacitance in the expression for ω_n in situations requiring a shielded input lead makes the coaxial cable capacitance perform the function of C_{fb}. MacNichol and Wagner[12] introduced this scheme and connected the cable shield to the wiper of a potentiometer across the output. The effect of a fixed feedback capacitance driven by an adjustable fraction of the output voltage is equivalent to that of a variable capacitance driven from the full output, as was illustrated in Fig. 4. Another approach to this problem has been to use an integral microelectrode holder and input tube cathode follower connected by flexible cable to the remainder of the circuit. Considerable care must also be exercised with this arrangement because the limited transconductance of electrometer tubes cause the cathode follower to have a relatively high output impedance.

In order to obtain some notion as to the effectiveness of neutralization of the input capacitance for various amplifier characteristics, the appropriate value of the feedback capacitor may be calculated for the critically damped case. If the damping factor ξ as given by (5) is set equal to unity, we have the relation

$$\tau_a[1 + M + NX] = 2\tau_a\sqrt{M + N} \tag{8}$$

where

$$M = \frac{\tau_i}{\tau_a}, \qquad N = \frac{\tau_{fb}}{\tau_a}, \qquad X = (1 - A_c),$$

and τ_{gk} has been assumed equal to zero for simplicity. This equation can be expressed as a quadratic in N and the roots given by

$$N = \frac{2 - X(1 + M) \pm 2\sqrt{A_c(1 - XM)}}{X^2} \tag{9}$$

where the root of interest is that with the negative sign before the radical. Solutions to (9) have been found as a function of the dc closed loop gain of the amplifier for a number of time constants and are summarized in Fig. 6. The ratio t_n/τ_i is plotted as a function of the amplifier gain where τ_i is the time constant R_iC_i, and t_n is the time for the output voltage to reach to within e^{-1} of the steady-state value. It is of interest to note that in order to obtain any appreciable input capacity neutralization, the amplifier time constant must be in the neighborhood of one-tenth that of the input circuit. The critically damped output response time can be further reduced in proportion to the square root of the reduction of the amplifier's time constant (for a given τ_i). This can also be seen easily from the square root-reciprocal relation between ω_n and τ_a in (5) and the fact that ω_n is the factor determining the time scale in the critically damped response expression (6). The speed of the amplifier appears to be much more important than its gain because the improvement in the system response time in going from a gain of 3 to 10 is not appreciable.

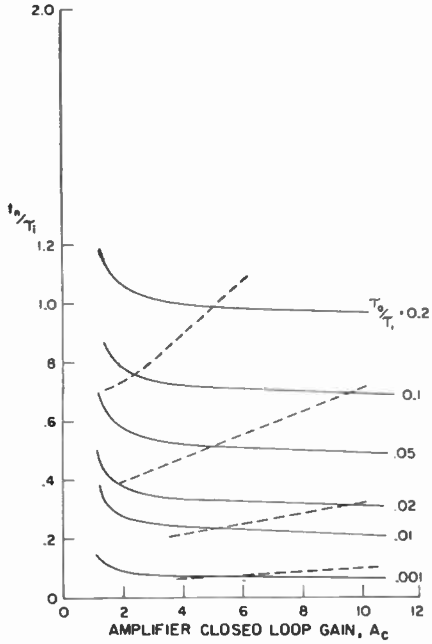


Fig. 6—Ratio of time for preamplifier output to rise to $1/e$ of its final value to input time constant for several values of amplifier gain and time constant. Each dashed line represents an amplifier (with a given figure of merit) and shows the effect of varying the negative feedback ratio to exchange closed loop gain for speed of response.

It might have been expected that more change would be seen in the response time because there would be a corresponding change in the value of C_{fb} from $0.5C_i$ to $0.11C_i$ [from (7)] for a scalar amplifier.

Another way of viewing these results is to plot the value of the feedback capacity as a per cent of its maximum compensating value (for a scalar amplifier) for various combinations of amplifier gain and time constant. Such a normalization also shows, in Fig. 7, that the percentage of maximum compensation used in the critically damped case is rather independent of gain but strongly dependent on the amplifier time constant.

If an amplifier of given open loop gain A_0 having a first-order dynamic response time constant of τ_0 is operated with negative feedback, it can be shown (Appendix I) that the closed loop gain A_c and time constant τ_a are proportional; *i.e.*

$$A_c = \frac{\tau_a}{\tau_0} A_0. \tag{10}$$

The factor A_0/τ_0 may be looked upon as a "figure of merit" for the amplifier package. The effect of varying the negative feedback factor to trade gain for speed is illustrated by the dashed lines in Fig. 6, each line representing an amplifier of a given figure of merit. The choice of closed loop gain for a particular application may be made after consideration of the speed and amplitude requirements for the output signal. For example, the level of bioelectric signals across cell membranes is in the neighborhood of 0.1 v. A gain of 2 to 3 would be quite adequate for driving fast oscilloscopes which are

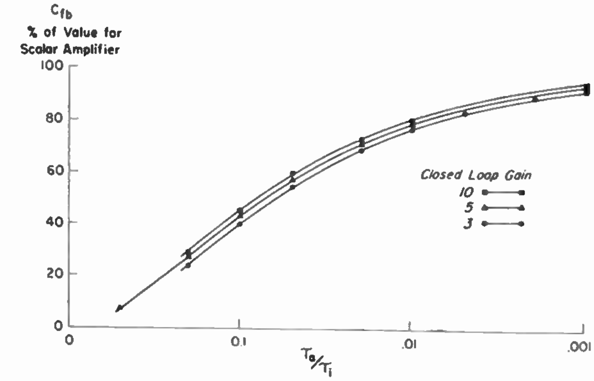


Fig. 7—Feedback capacitance (as per cent of value for scalar amplifier) required for critically damped responses as a function of the amplifier gain and time constant for a fixed source time constant.

available with input sensitivities of 50 mv/cm and would permit the compensation to approach the maximum for a given amplifier.

The results of this analysis are for the rather arbitrary case of critical damping and are probably useful only for showing general trends with parameter variations. An appreciably faster response may be obtained by use of a damping coefficient of 0.6, with a 10 per cent overshoot which is acceptable for many applications. The assumption of a first-order amplifier package also limits the usefulness of this analysis. In general, a multistage amplifier will have a higher-order dynamic response unless carefully designed so that the over-all phase shift and amplitude cutoff is dominated by a single time constant which is much longer than any of the others in the amplifier. The omission of the capacitor C_{gk} in (9) to simplify the study of the roots also introduced some errors in the analytic results.

IV. Circuit Simulation of an Analog Computer

Because of these limitations in the analysis and the complexity and tedium of handling more realistic equations, it appeared attractive to study the system on an analog computer because of the

1) Speed and convenience of varying design parameters and plotting out the associated transient responses on an X-Y plotter.
2) Ease of handling a multiple time constant amplifier.
3) Ease of studying the effect of positive feedback.
4) Possibility of investigating alternate methods of input capacity compensation.

It is possible to program such a computer to solve for V_0 directly from (1). However, because both C_{fb} and C_{gk} are imbedded in the coefficients of V and V_0, the effect of variation of *either* of these for a real amplifier system would require a change of two potentiometers on the computer. Furthermore, the amount of this change would depend on the values of other parameters in the circuit. It seemed more convenient to simulate the circuit after a fashion in which the amplifier source impedance, amplifier gain and phase, shield and feedback capacitors, etc., could be directly associated with components in the computer circuit.

An analog computer (Donner 30) was programmed to simulate the lumped tip parameters and a preamplifier of the MacNichol and Wagner configuration (see Fig. 4). The machine time scale was chosen slow enough so that a servo driven X-Y plotter could accurately follow the output. The use of a ratio for machine-to-real time of 10^5, and input resistors in a one-to-one correspondence, meant that a capacity of 1 μf on the simulator represented a real input capacity of 10 pf; therefore, stray wiring capacity in the simulator could be neglected. In addition, a scalar amplifier could be approximated because the flat frequency response of the analog operational amplifier exceeds the highest frequencies of interest in machine time by several decades, or an amplifier described by first- or second-order equations could also be simulated by addition of passive networks. With analog simulation, it is also possible to have all signals of interest in the real amplifier, including the capacitor currents, available for recording at low impedance levels.

A schematic of a simplified simulator circuit is shown in Fig. 8. Lumped constants were used for the microtip and input tube; it is also feasible to arrange for distributed constants if desired. The triangles represent operational amplifiers[30] which force the input grid to remain at or near ground potential by negative feedback from the output.

In order to gain some understanding of the operation of the circuit, let us consider first what happens when both C_{gk} and C_{fb} are set at zero. A positive step function of voltage at E will cause the output of operational amplifier no. 1 to generate a voltage $-V$ which will approach a steady value in an exponential fashion with a time constant of R_iC_i. This signal is changed in sign to $+V$ by the unity gain inverter no. 2; V corresponds to the potential on the input grid of the preamplifier being simulated. The sum of this potential and the negative of that on the input cathode $-\beta V_0$ is amplified by operational amplifier no. 3 which simulates an amplifier with an open loop dc gain of A_0; its time constant is set by C_a. If the closed loop amplifier response τ_a is fast compared to τ_i, the output voltage response time constant will approximately equal τ_i.

If C_{gk} is now set to a value corresponding to the sum of the grid-to-cathode capacity and the driven shield capacity, a small additional current equal to $C_{gk}p(\beta V_0 - V)$ will enter the summing point of amplifier no. 1; the term $(\beta V_0 - V)$ is generated by amplifier no. 4. Because the magnitude of the grid potential V normally exceeds βV_0 by a small amount, the whole term will be negative and is net effect will be to slow somewhat any transient change at V.

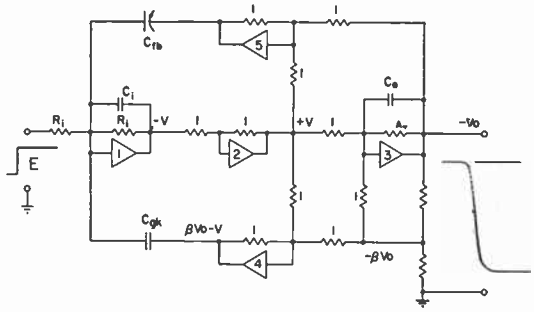


Fig. 8—Diagram of simplified analog circuit for simulation of capacitance-neutralizing preamplifier. The triangles represent the operational amplifiers of the analog computer.

Neutralization of input capacity may be obtained by introducing the proper value of capacitance between output and the input grid of the real amplifier. In the analog, the voltage difference between the output and the input grid is generated by operational amplifier no. 5. The current that flows to the summing junction of amplifier no. 1 will be $C_{fb}p(V_0 - V)$. This can supply most of the current for C_i (the lumped tip and grid to ground capacity) and cause V to rise much more rapidly than when C_{fb} was at zero.

The amplifier characteristics first used in this simplified analog simulator were taken from the MacNichol and Wagner[12] circuit: dc open loop gain $A_0 = 57$ and $\tau_0 = 8.4$ μsec. A negative feedback factor β of 0.2 was normally used so that the net gain was 4.6 and the closed loop time constant, τ_a, was 0.68 μsec. The input circuit usually simulated consisted of a 20 megohm input resistance shunted by 10 pf. Thus τ_i/τ_a would have a value of

$$M = \frac{200}{0.68} \simeq 300.$$

Transient responses recorded from the simulator under these conditions are shown in Fig. 9 for several values of the compensating feedback capacitor C_{fb}. The value of t_n for $C_{fd} = 2.3$ pf is 28 μsec, and this gives a t_n/τ_i ratio of 0.14, in agreement with the analysis as illustrated in Fig. 6. The system rings badly for values of C_{fb} which are appropriate for a scalar amplifier as given by (7). Fig. 10 indicates the slowing of the over-all response by the addition of the capacitance of an input cable whose shield is either connected to the negative feedback point (curve 2) or to ground (curve 3). The neutralizing capacitor has been adjusted to give nearly identical responses for these two cases.

The effect of changing the time constant of the amplifier at a fixed gain is illustrated in Fig. 11. The feedback capacitor was adjusted in each case for the fastest rise with no more than about a 1 per cent overshoot. The

[30] When the operational amplifier grid current is small, the sum of currents entering the grid lead (or summing point) via input impedances must leave through the feedback impedance. From this, the ratio of the output (V) to input (E) voltage can be derived and shown closely to approximate Z_f/Z_i for a high gain amplifier where Z_i and Z_f are the input and feedback impedances respectively. A standard text on analog computers may be consulted for a more detailed discussion, e.g., C. L. Johnson, "Analog Computer Techniques," McGraw-Hill Book Co., Inc., New York, N. Y.; 1956.

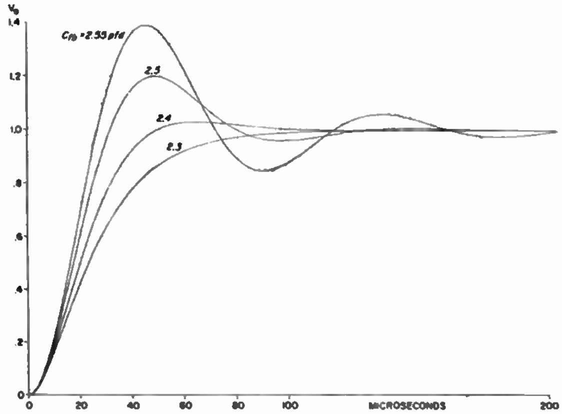


Fig. 9—Analog response to step input voltage for several values of the feedback capacitor C_{fb}. The closed loop gain is 4.6; the appropriate values of C_{fb} for a scalar amplifier of this gain would be 2.78 pfd.

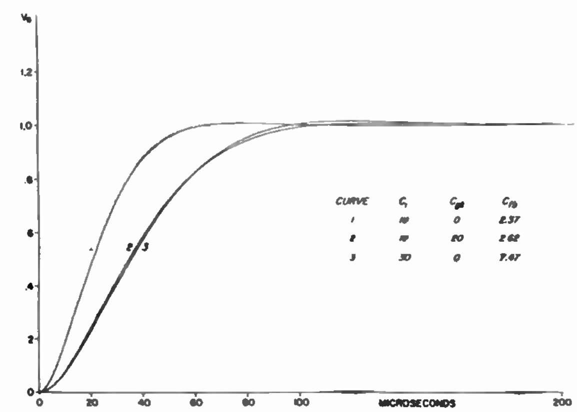


Fig. 10—Analog transient responses showing the effect of the addition of a shielded input lead capacitance of 20 pfd (equivalent to several inches of coaxial cable). Curve 1—No shield. Curve 2—Shield tied to negative feedback point. Curve 3—Shield tied to ground.

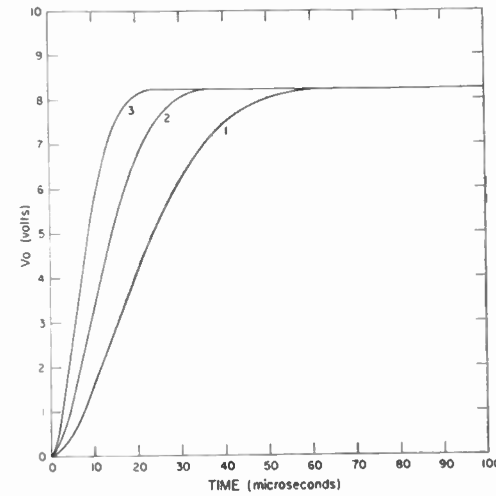


Fig. 11—Analog records showing the effect of the amplifier bandwidth on the over-all transient response with the standard 20 meg, 10 pfd source. The open loop gain is 57 and the open loop time constants are 8 μsec (curve 1), 3 μsec (curve 2), and 1 μsec (curve 3).

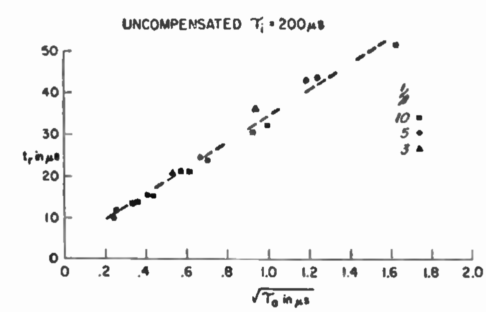


Fig. 12—A summary of analog results in which the compensated rise time (0.1 to 0.9 of final value) is plotted as a function of the square root of the closed loop time constant for three values of β.

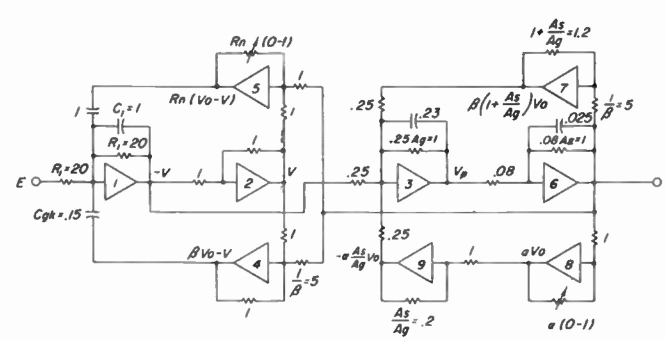
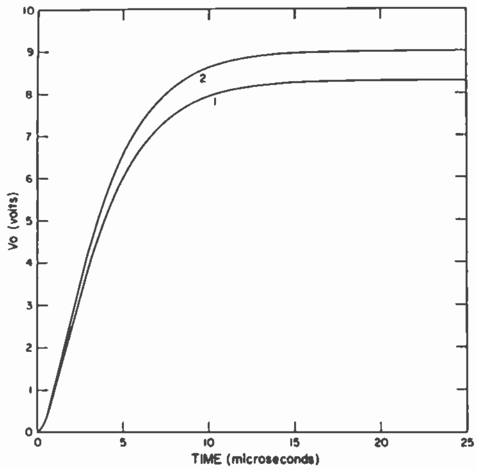


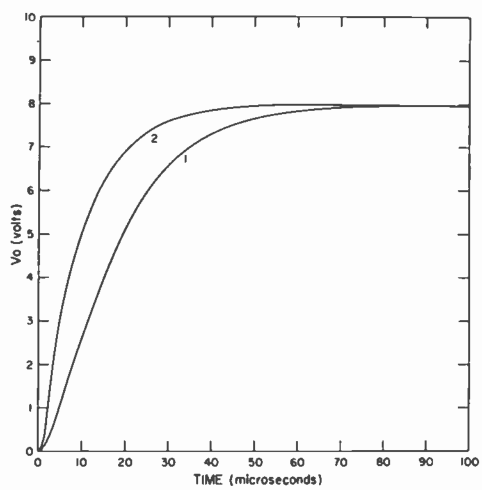
Fig. 13—Circuit diagram of analog simulator for second-order amplifier with positive feedback to the input screen grid.

results of a number of such simulator runs are summarized in Fig. 12; in general, the over-all picture is much the same as was obtained analytically for the case with critical damping.

A more complex circuit was also developed to simulate a second-order amplifier and is shown in Fig. 13. It included provision for simulation of the type of positive feedback which we were planning to add to the MacNichol-Wagner[12] preamplifier design (see next section). For these purposes, it was convenient to simulate the input stage as a first-order device and to follow this with a first-order gain package representing the balance of the amplifier. The equivalent circuit for such a system and its equation (11) are given in Section V. The left half of Fig. 13 generates a voltage simulating that on the input grid of the preamplifier and is essentially unchanged from the original circuit (Fig. 8). The output of the analog operational amplifier no. 3 represents the potential on the plate of the input tube. The voltage gain for signals applied the control grid (A_g) of the 5879 input tube is about 4 under the conditions used by MacNichol and Wagner.[12] The corresponding voltage gain for signals on the screen grid (A_s) of this tube is about 0.8. The ratio A_s/A_g was therefore taken as 0.2. The negative feedback ratio was fixed at 0.2 for a nominal net gain of 5. The positive feedback fraction α could be varied from 0 to 1. The variable feedback capacitor in the original simulator has been replaced by an equivalent circuit consisting of a fixed capacitor driven by a variable fraction of the potential difference between the input grid and the output.



(a)



(b)

Fig. 14—Transient responses obtained with the simulator of Fig. 13. (a) Effect of increasing open loop gain (first order) from 57 (curve 1) to infinity (curve 2) by positive feedback; 1 megohm source resistance. (b) Illustrates compensation for a 20 megohm source with first-order amplifier (curve 1) and the improvement obtained by peaking its high frequency response (curve 2).

Two results obtained with this simulator are illustrated in Fig. 14. An increase of open loop gain from the original value of 57 to "infinity" increased the output voltage to precisely 5 times the input but gave essentially no change in the compensated rise time [Fig. 14(a)]. When the amplifier time constants were adjusted to give a 10 per cent overshoot in response to a step from a low resistance source, the compensated rise time for a high resistance source was appreciably shortened over that for the amplifier with a first-order response [Fig. 14(b)].

V. Improvements in Preamplifier Design

Over a period of some years we have constructed, modified, and used several amplifiers of the MacNichol-Wagner design. The need for a system which was faster than the original and incorporated a stable baseline for experiments in which the membrane potential was controlled by feedback became urgent, and we sought ways to improve on their effective and straightforward design. MacNichol and Wagner chose the value of the plate load resistor of the input tube to be 560 kΩ which, in combination with an estimated interstage capacity of 10 pf, would give a time constant of about 5 μsec, much longer than any other time constant in the system. Actual measurements, as given in their paper, show a cutoff frequency of about 19 kc, corresponding roughly to a time constant of 8 μsec for a first-order response. Precision wirewound resistors were used in the negative feedback ratio network in the original design and mica capacitors had been added in shunt with them to suppress a tendency for oscillation at several hundred kc. We first substituted lower value precision metal film resistors in the feedback ratio network and removed the capacitors when they were found to be no longer needed (probably because of the lower inductance of the film resistors). The power supply potentials were increased and the input plate resistors were decreased slightly. These modifications more than doubled the open loop frequency response (43 kc cutoff); the amplitude and phase characteristics are shown as curves 1 and 2, respectively, in Fig. 15. With a closed loop gain of 5, this amplifier exhibits a bandwidth of nearly 1 Mc, as shown by curves 3 and 4 of Fig. 15.

These modifications, of course, produced little change in the open loop gain from the value of 57 for the original design. A much larger value is preferable, so that the net closed loop gain is dependent only on the ratio of precision resistors. For example, the net gain of the original design is 2.85 instead of the value of 3 which would be obtained for a high open loop gain with the same feedback ratio of 1/3. For a feedback ratio of 0.2 (which we wanted to use) the corresponding net gain would be only 4.6 instead of 5. Although it is possible to alter the feedback ratio slightly so that the desired net gain is obtained, the alteration factor will still be dependent on the open loop gain which may vary appreciably with tube age, etc.

Other important limitations of such a low value of open loop gain can be appreciated by the following considerations. The cathode of the input stage does not precisely follow the grid potential. The voltage difference between these tube elements would be about 5 and 8 per cent of the input signal for net gains of 3 and 5, respectively. This means that the effect of the capacity between the input grid and cathode is not reduced as much as is desirable. This would be particularly troublesome with the large C_{gk} which results from tying an input shield to the cathode. For purposes of stabilizing against drift, it is necessary to insert a resistance network between the grid and cathode. This would cause a reduction of the input resistance and increase the current flow in the input circuit unless the cathode is made to follow the grid potential precisely.

For these reasons, it seemed necessary to increase the open loop gain of the circuit. Although it is possible to

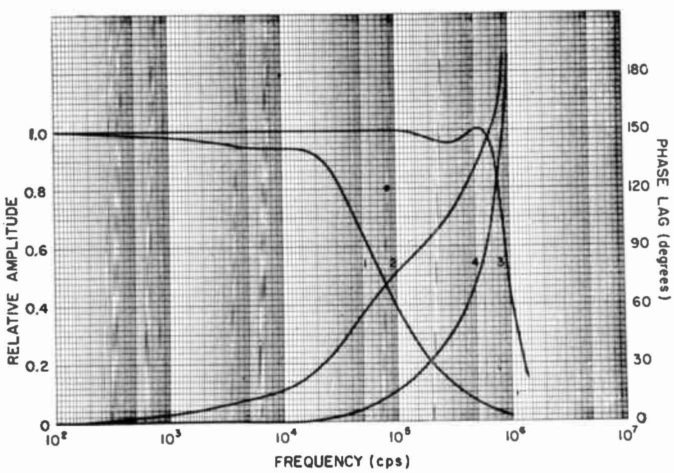
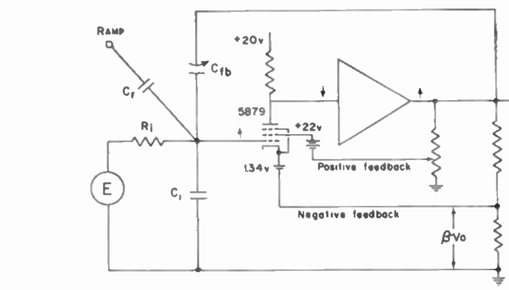


Fig. 15—Open loop gain (curve 1) and phase (curve 2) characteristics of the modified MacNichol-Wagner amplifier. The corresponding characteristics for this amplifier with a closed loop gain of about 5 are given by curves 3 and 4.

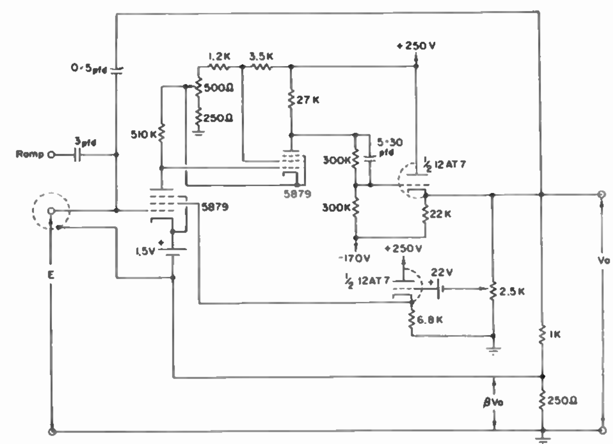
increase the open loop gain by additional stages of amplification, we chose the alternative procedure of using positive feedback because of the concomitant advantage of reduction of the input capacity when an in-phase signal is applied to the screen grid. As a matter of fact, the amount of in-phase swing on the screen grid required to produce the desired open loop gain was usually slightly in excess of the signal amplitude on the input grid. This of itself, therefore, added some input capacitance neutralization.

A block diagram of this configuration is given in Fig. 16(a) and a circuit schematic of this preamplifier is shown in Fig. 16(b). It uses the same complement of tubes as the original MacNichol-Wagner configuration and their functions are retained with the exception of using one of the original pair of output triodes for positive feedback. A cathode follower is used to drive the input screen grid with the sum of a dc potential from a battery and a fraction of the output voltage taken from a low impedance potentiometer. The output impedance was not increased by the sacrifice of one of the output triodes because the quiescent current through the remaining cathode follower was increased beyond that of the original pair with an accompanying increase in g_m. The supply voltage for the plate of the input tube was varied to set the dc output level, as in the original design. The filaments as well as the positive and negative voltages were supplied by highly regulated dc from line operated power supplies. The net equivalent ripple at the input of the preamplifiers was usually a fraction of a millivolt peak-to-peak and was quite satisfactory for bioelectric measurements where full scale values were ± 150 mv or so.

For purposes of analysis, the positive feedback system used in this amplifier may be represented by the linear equivalent circuit of Fig. 17(a), where the input stage characteristics are treated in detail and the remainder of the amplifier lumped into $-A_2$. The small loading of the

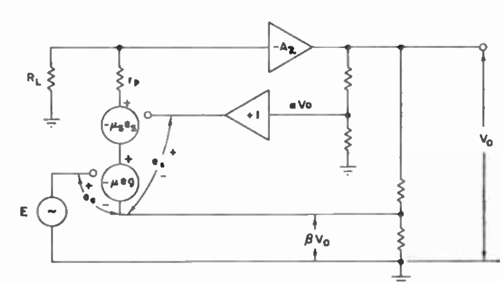


(a)

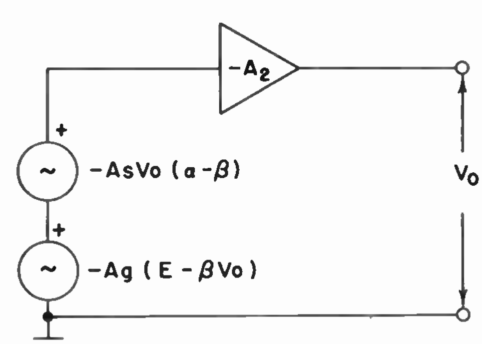


(b)

Fig. 16—Amplifier with positive feedback to the screen grid. (a) Block diagram. (b) Circuit schematic.



(a)



(b)

Fig. 17—(a) Linear equivalent circuit of amplifier in Fig. 16. (b) Simplified equivalent circuit.

input stage on the low impedance negative feedback network may be neglected and the equivalent circuit simplified as in Fig. 17(b). The gain of the first stage signal grid is represented by A_g and that of the screen grid by A_s. Then we may solve for the closed loop gain G as follows:

$$G = \frac{V_0}{E} = \frac{A_g A_2}{1 + \beta A_g A_2 - A_s A_2(\alpha - \beta)} \cdot \qquad (11)$$

For the case of $\beta = 0$ (no negative feedback),

$$G \to \infty \quad \text{when} \quad \alpha = \frac{1}{A_s A_2} \cdot \qquad (12)$$

The positive feedback fraction α which will give a gain which is dependent only on the negative feedback fraction β is given by the expression

$$\alpha = \beta + \frac{1}{A_s A_2} \cdot$$

For the real amplifier, this condition was met by setting the fraction of positive feedback signal to the point where the *difference* between an input signal (of about 100 mv) and the feedback voltage (βE_0) was zero, as observed on an oscilloscope with a differential input. This procedure makes the open loop gain approach "infinity" in effect and increases the input impedance, as well as causing the closed loop gain to be determined by the feedback ratio alone. For the amplifier shown in Fig. 16(b), a low impedance precision resistor network with a β of 0.2 was used to give a net gain of 5.

Some concern has been expressed about the susceptibility of such a positive feedback circuit to ringing or oscillations. If the proper phase and amplitude response has been originally established for negative feedback around the amplifier to give a smooth response to a step input from a low impedance source, the addition of a positive feedback signal, to increase the open loop gain, will not appreciably affect the closed loop transient response because most of the increase in gain is at the lower frequencies. Photographs of the response of the preamplifier to a step function input from a low impedance source with and without the positive feedback is shown in Fig. 18; the rise time for the normally used "infinite" open loop gain setting is 0.38 μsec. Several circuits of this general type have been constructed in our laboratory and none has shown any more tendency to ring than this unit, even with the magnitude of positive feedback increased substantially past the appropriate value for "infinite gain." Thus, there is small likelihood of an increase in the normal forward open loop gain of the amplifier causing the amplifier to ring. Actually, some ringing would not be a disadvantage because, as already shown on the simulator, a somewhat better capacitance neutralization can be obtained with an amplifier which has a small amount of ringing or overshoot in its response to a step input from a low impedance source. In the Appendix we show that the major effect of a small increase in positive feedback over that amount required for "infinite gain" is a degradation of performance, similar to that observed upon a decrease in forward gain, but of opposite sign.

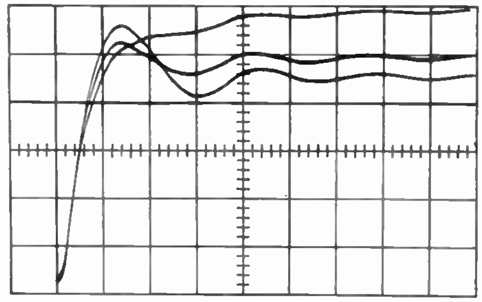


Fig. 18—CRO photographs of the transient response of the amplifier to input step functions from a low impedance source. The increasing steady-state outputs correspond to positive feedback settings of zero, "infinite gain," and maximum. The sweep rate is 0.5 μsec/major division.

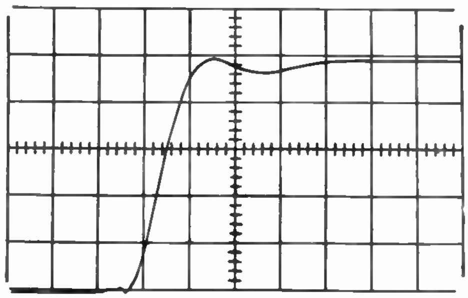


Fig. 19—Source preamplifier system dynamic response tested by injection of a current step via the "ramp capacitor." The source resistance is 20 megohms and enough external capacitance has been added to give an uncompensated time constant of 200 μsec (or a rise time of 440 μsec). The sweep rate is 20 μsec/major division.

Fig. 19 illustrates the transient performance of this amplifier when connected to a 20 megohm source resistance and enough additional external capacity from grid-to-ground to give an over-all uncompensated time constant 200 μsec or a rise time (0.1 to 0.9) of 440 μsec. A step of current was injected at the grid through the ramp capacitance and the neutralization adjusted to give a smooth output response with a rise time of about 22 μsec. Thus, the response time has been reduced to 0.05 that of the input circuit. Although τ_n is not defined for this amplifier because it does not have a pure first-order response and the curves of Fig. 6 are not directly applicable, the ratio of the rise times for the amplifier and source used in this example is about 0.0008 and can apparently be used in place of the time constant ratio in Fig. 6 to give a rough estimate of the available compensation.

A study of the effect of positive feedback on the constancy of the closed loop gain has been made by analyzing its sensitivity to variations in stage gains and feedback ratios. Expressions have been derived and representative values calculated for the present circuit as well as for several generalized cases and are tabulated in Appendix II. The results may be summarized here by noting that positive feedback around one of the stages can make A_c insensitive to small variations in the gain of the other stage. With positive feedback around each stage, A_c is independent of small changes of gain in either stage, but not of simultaneous changes in both stages. If the positive feedback is applied around both stages together, as in our amplifier, there is no improvement in closed loop gain stability.

The effect of positive feedback on the output impedance was also derived and shown in Fig. 21 in Appendix II. For the amplifier of Fig. 16(b), the output impedance was increased from 10.5 Ω for no positive feedback to 11.5 Ω when the positive feedback was adjusted for "infinite gain." Although it is clear that positive feedback is no substitute for a normal high forward gain as far as output impedance reduction is concerned, this impedance level is low enough to be insignificant in normal instrumentation practice.

VI. Base Line Stabilization

The need to control, by feedback, the potential across a bioelectric membrane[1] as measured by such a preamplifier adds a further requirement that the base line be stable. Drift in the preamplifier is indistinguishable from signal and not only introduces an error into the measured potential, but also forces the membrane potential away from the command value by the amount of the drift. In an excitable biological tissue, a millivolt deviation or change in the steady-state potential can make significant changes in its response to constant voltage pulses. An accumulated drift in the order of a few millivolts may cause such drastic changes in the condition of an excitable membrane that they will mask those introduced by the experimental procedure. Therefore, it was both desirable and necessary to stabilize the electrometer preamplifier as well as the standard operational amplifiers in the rest of the system.[1]

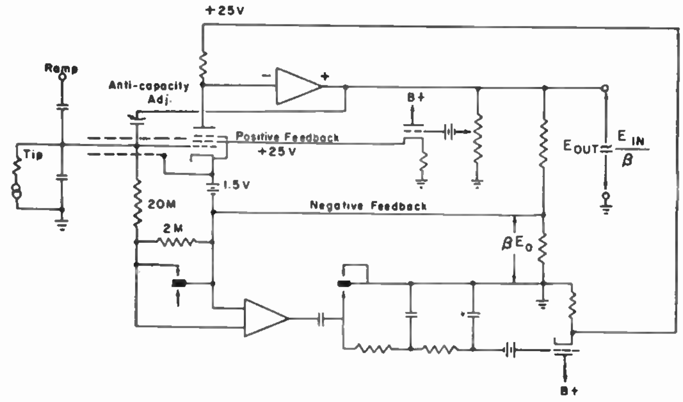
Goldberg[31] has developed a method for stabilizing operational amplifiers by means of a chopper, auxiliary ac amplifier, rectifier, and filter; this technique has been rather generally adopted. We have applied the Goldberg approach to the more difficult problem of stabilizing our potentiometric type of amplifier against drift.

Although a large amount of negative feedback reduces the effect of drifts introduced in later stages, it cannot compensate for the main source of drift which is the variation in the "contact potential" between the grid and cathode of the input stage caused by changes in electron emission velocity with heater voltage variation. This is equivalent to an additional potential in series with the signal and, for oxide-coated cathodes, amounts to approximately 0.2 v for a 20 per cent change in the heater voltage.[32] The effect of heater voltage fluctuation can be reduced to a few millivolts by use of any one of a number of commercially available units which regulate the heater voltage to within about ±0.1 per cent for ±10 per cent live voltage variations. This does not afford a sufficient reduction in drift for long term measurements where the maximum signal is only a few millivolts. In order further to stabilize an amplifier of the potentiometric type against baseline drift, the dc difference in potential between the grid and cathode must be measured and controlled to a constant value. In addition, the auxiliary monitoring circuit must not degrade the high impedance of the amplifier nor introduce an effective grid current. Fortunately, the normal procedure of using negative feedback provides a voltage which follows the signal input more or less precisely. The impedance of a network inserted between the negative feedback point and the signal grid will be raised in effect by a factor approximately equal to $A_0\beta$. If the open loop gain is made very large, the network current goes toward zero and its effective impedance approaches infinity.

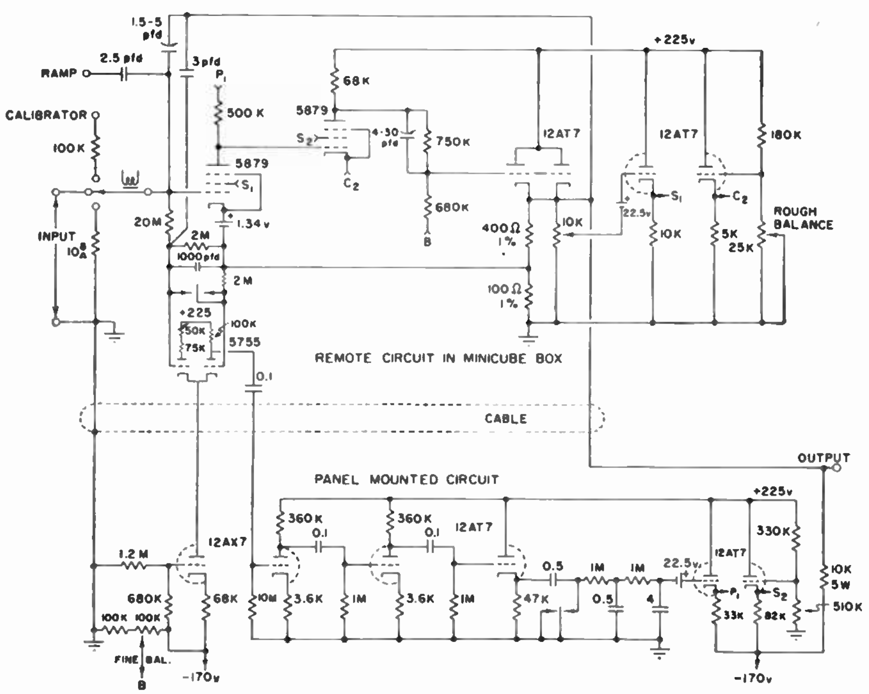
Of the several possible devices considered for conversion of dc to an ac signal, the photo converter seems to be the ideal. However, these were not well developed at the initiation of this work. The available magnetic modulators were of far too low current sensitivities. The remaining choice was that of the electromechanical chopper which is also widely used with operational amplifier stabilization. To avoid shorting the grid-to-cathode feedback point when the chopper was closed and to keep from introducing chopper noise into the input, a voltage divider was inserted between the grid and cathode feedback as shown in Fig. 20(a). The chopper shorts the resistor nearest the feedback point for about half of each drive cycle. An ac amplifier with high impedance differential input boosts the chopped signal level by more than 80 db. The output of this amplifier is synchronously rectified by a second chopper and is well filtered with a time constant in the order of one second. The output of the filter is added to a constant potential from a battery in a cathode follower whose output supplies the plate voltage for the input tube. The sense of the polarities is arranged to change the potential on the input plate in such a direction as to minimize the chopped signal at the input to the auxiliary ac amplifier. This, in turn, means that the dc level of the feedback point automatically and accurately tracks the signal grid. An auxiliary manual dc level control is used initially to set the output to zero when the signal lead is grounded. Under these conditions, the amount of the stabilizing control signal on the first plate is minimized; it normally remains less than ±1 volt over long periods of operation. Consequently, the difference in the dc potential between the signal grid and feedback point

[31] E. A. Goldberg, "Stabilization of wideband dc amplifiers for zero and gain," *RCA Rev.*, vol. 11, pp. 296–300; June, 1950.

[32] G. E. Valley and H. Wallman, "Vacuum Tube Amplifiers," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., vol. 18, p. 421; 1948.



(a)



(b)

Fig. 20—Circuit for chopper stabilization of potentiometric amplifier. (a) Block diagram. (b) Circuit schematic. Not shown, 10Ω in series with input heater.

was held to a fraction of a millivolt. The dc current which would flow between these two points for an offset of 0.1 mv would be 5×10^{-12} a. This is equivalent to an additional grid current but is within the specifications originally outlined.

The choppers were driven at 60 cps in our initial designs. Care was taken to keep the high-level chopped signal output well shielded from the high resistance input. These signals are volts for millivolts of dc offset and have high rates of rise so that capacity coupling must be minimized. Initial attempts at putting the lower side of the output filter a few hundred ohms off ground so that the dc resetting signal could be sent to the cathode or screen grid were always unsuccessful because of the few millivolts of ripple that were introduced into the main amplifier. Consequently, we adopted the quieter design in which the lower or reference side of the filter was grounded and the reset signal fed into a lower gain point further from the input. In our present design shown in Fig. 20(b), the rebalancing signal is introduced via a cathode follower at the input tube plate supply. A battery (22.5 v) in series with the filter output sets the approximate operating point to which rebalancing signals are added (or subtracted). Even with the disadvantage of a 10:1 attenuator at the input chopper, and feeding back the reset signal to the first plate, a stabilization ratio of greater than ten is easily obtained. By "stabilization ratio" is meant the ratio of the dc offset without the automatic resetting circuit to the offset with this circuit operating for a small change in the heater voltage or balance potentiometer. When the heater voltage has been regulated to within about 0.1 per cent for line changes, it has been possible to hold the drift of the amplifier to within ± 200 μv referred to the input over indefinite periods of time after a short warmup. These limits can be squeezed somewhat tighter by more ac amplifier gain or reduction of the input chopper attenuator with a slight reduction in the quality of the input impedance.

Appendix I

Gain and Time Constant Relations

For a potentiometric amplifier with negative feedback as in Fig. 4, the input/output ratio may be written as

$$\frac{V_0}{E} = \frac{\mu}{1 + \mu\beta} \tag{13}$$

where μ is the amplifier gain. If the amplifier has a first-order open loop characteristic, μ may be written as $\mu_0/(1+p\tau_0)$ where μ_0 is the low frequency gain, τ_0 is the open loop time constant, and p is the operator d/dt. For a particular amplifier, μ_0 and τ_0 may be taken as given constants and the expression for μ substituted into (13). The closed loop time constant τ_c can be found to be a function of the negative feedback ratio as follows:

$$\tau_c = \frac{\tau_0}{1 + \mu_0\beta}. \tag{14}$$

The closed loop gain at low frequencies is given by (13) and may be called G. By substitution we obtain the relation

$$\frac{\tau_c}{\tau_0} = \frac{G}{\mu_0} \quad \text{or} \quad \frac{\tau_c}{G} = \frac{\tau_0}{\mu_0}. \tag{15}$$

This shows that, for a given amplifier with a first-order response, the ratio of time constant to gain for the closed loop is a constant. Thus the designer may purchase a faster response at the expense of a reduction in net gain.

Appendix II

Effects of Positive Feedback Inside a Negative Feedback Loop

The sensitivity of the net closed loop gain to variations in the stage gains and feedback ratios has been analyzed for a number of circuit arrangements. The sensitivity of the closed loop gain G to the variation of a parameter K is symbolized by $S_K{}^G$ and is taken as the ratio of the fractional change in closed loop gain to the fraction change in K;

$$S_K^G = \frac{G}{K}\frac{dG}{dK}. \tag{16}$$

Schematic diagrams of some possible circuit configurations are shown in Fig. 21 along with a tabulation of the sensitivity to various parameter variations.

In general, the effect of the positive feedback is to reduce the denominator D so that the closed loop gain increases and approaches the reciprocal of β. If the positive feedback is set so that the denominator is precisely $A_1A_2\beta$, the open loop gain is infinite. Something less than this amount of positive feedback will leave an output error which is characteristic of insufficient forward gain. Positive feedback in excess of that for infinite gain will cause a slightly excessive output; *i.e.*, the error has a sign opposite that for insufficient forward gain. If the transient response of the negative feedback loop has no tendency toward ringing, it is possible to use considerable positive feedback without getting into difficulties with the transient response. Therefore, it is practical to use an amount of positive feedback which initially sets the open loop gain at infinity and to expect to have only minor degradation in the closed loop performance resulting from open loop gain decreases or increases (caused by tube aging or replacement, etc.). Some diagrams of circuits exemplifying the structures shown in Fig. 21 are given in Valley and Wallman.[33]

The output impedance with feedback is also calculated and given in Fig. 21 as a fraction of the open circuit output impedance.

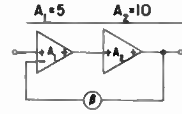
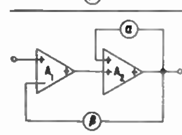
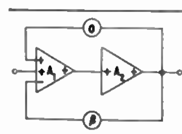
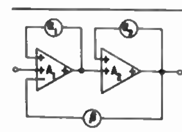
CIRCUIT STRUCTURE $A_1=5$ $A_2=10$	α	β	D	G	$S^G_{A_1}$	$S^G_{A_2}$	S^G_α	S^G_β	$\frac{Z_2}{Z_0}$
		0.18	$1+A_1A_2\beta$	$\frac{A_1A_2}{D}$ 5	$\frac{1}{D}$ 0.1	$\frac{1}{D}$		$\frac{-A_1A_2\beta}{D}$ -09	$\frac{1}{D}$ 0.1
	0.1	0.2	$1-A_2\alpha+A_1A_2\beta$	$\frac{A_1A_2}{D}$ 5	$\frac{1-A_2\alpha}{D}$ 0	$\frac{1}{D}$ 0.1	$\frac{A_1\alpha}{D}$ 0.1	$\frac{-A_1A_2\beta}{D}$ -1.0	$\frac{1}{D}$ 0.1
	0.02	0.2	$1-A_1A_2\alpha+A_1A_2\beta$	$\frac{A_1A_2}{D}$ 5	$\frac{1}{D}$ 0.1	$\frac{1}{D}$ 0.1	$\frac{A_1A_2\alpha}{D}$ 0.1	$\frac{-A_1A_2\beta}{D}$ -1	$\frac{1}{D}$ 0.1
	α_1=0.2 α_2=0.1	0.2	$(1-A_1\alpha_1)(1-A_2\alpha_2)+A_1A_2\beta$	$\frac{A_1A_2}{D}$ 5	$\frac{1-A_2\alpha_2}{D}$ 0	$\frac{1-A_1\alpha_1}{D}$ 0	$\frac{A_1\alpha_1}{D}$ 0.1 $\frac{A_2\alpha_2}{D}$ 0.1	$\frac{-A_1A_2\beta}{D}$ -1	$\frac{1}{D}$ 0.1

Fig. 21—Summary chart showing sensitivity of gain to variations of several parameters. The last column gives the ratio of the closed to open loop output impedance.

Acknowledgment

We are grateful for the suggestions and discussions of a number of colleagues: E. Amatniek, A. Bak, J. Coombs, K. Frank, J. Hervey, J. Y. Lettvin, J. R. Macdonald, E. F. MacNichol, R. Schoenfeld and M. Wolbarsht.

[33] *Ibid.*, p. 475.

Bandwidth Limits for Neutralized Input Capacity Amplifiers*

ROBERT L. SCHOENFELD†, MEMBER, IRE

Summary—**Criteria for bandwidth limits for neutralized input capacity amplifiers are developed. The behavior of different neutralization schemes is analyzed in terms of the root locus of the system gain function. This technique permits quantitative design and evaluation of the different circuits that have been used. It makes it possible to judge the effectiveness of new approaches to the problem.**

It is shown that second-order amplifiers may achieve inherently faster response than amplifiers with a single time constant. Using a critically damped criterion, the maximum bandwidth of the second-order amplifier is equal to the cube root of the product of the input circuit bandwidth times the square of the amplifier bandwidth with the input circuit removed. A single time constant amplifier has a maximum bandwidth equal to the square root of the product of the input circuit bandwidth times that of the amplifier alone.

It is shown also that one may have to choose between speed of response and excess noise. The noise figure of these systems may increase markedly with bandwidth and increases to a lesser degree with the system complexity.

INTRODUCTION

NEUTRALIZED input capacity amplifiers are used to improve the bandwidth of biolelectric signals recorded with electrolyte-filled micropipette electrodes.[1] Such electrodes constitute a very high resistance in series with the source of bioelectric potential. The input capacitance at the amplifier is augmented by contributions from the electrodes, cables and feedback circuits. The electrode resistance and the total input capacitance comprise a low-pass filter with a narrow pass band that limits the fidelity of reproduction of the source waveform.

This situation is idealized in Fig. 1 which illustrates a simple lumped parameter model for the input circuit and the general scheme used to improve the bandwidth. The resistance R_1 represents the electrode resistance which may vary between 5 and 50 megohms. The input capacitance C_1 is of the order of several picofarads. The amplifier A provides positive feedback through the capacitance C_2. If the amplifier is assumed to have infinite bandwidth, the effective input capacitance $C_1+C_2(1-A)$ can be adjusted to zero. Under these assumptions, complete input capacity neutralization is possible.

Guld has pointed out that the degree of neutralization is limited fundamentally because of the inadequacy of the lumped parameter electrode model.[2] The deeper the penetration of the electrode in the tissue, the more important becomes the effect of distributed capacitance along the electrode.

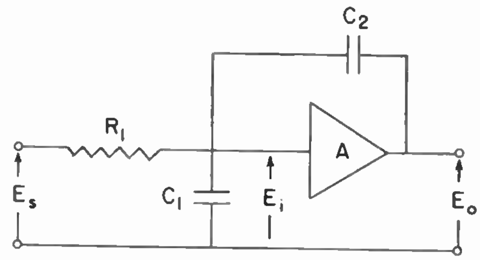


Fig. 1—Circuit model for input capacity neutralization. R_1 = electrode resistance, C_1 = input capacity, C_2 = feedback capacity, A = amplifier gain, E_s = signal voltage, E_i = amplifier input voltage, E_o = amplifier output voltage.

However, accepting the simple lumped parameter model for purposes of analysis, it is necessary to take into account the amplifier frequency response characteristics. The simplest realistic model represents the amplifier as a single time constant, low-pass filter with a cutoff angular frequency ω_2 much larger than that of the input circuit; *i.e.*,

$$\omega_2 \gg \omega_1 \quad \text{where} \quad \omega_1 = \frac{1}{R_1(C_1 + C_2)}.^{3}$$

Moore has reported a negative capacitance amplifier which uses combined negative and positive feedback within the block labeled A in Fig. 1.[4] This amplifier behaves as an underdamped low-pass filter of second order; *i.e.*, its amplitude falls off at the rate of 12 db per octave at high frequencies. Moore claims that better capacity neutralization can be accomplished with this system than can be obtained with an amplifier possessing primarily a 6-db octave cutoff at high frequencies.

The purpose of the present paper is to carry out an analysis of the model of Fig. 1 and to determine bandwidth limits for the system for different amplifier response characteristics. This analysis is carried out by studying the root locus of the system transfer function. It tends to substantiate Moore's claim that wider

* Received September 29, 1961; revised manuscript received June 25, 1962.
† The Rockefeller Institute, New York, N. Y.

[1] E. Amatniek, "Measurement of bioelectric potentials with microelectrodes and neutralized input capacity amplifiers," IRE TRANS. ON MEDICAL ELECTRONICS, vol. ME-10, pp. 3–14; March, 1958. Contains bibliography on this subject. This issue contains papers presented at a symposium on this subject at the U. of Penn., June, 1956.

[2] C. Guld, "Factors limiting capacitance neutralizing in microelectrode amplifiers," *Proc. 2nd Internat'l Conf. on Medical Electronics*, Paris, France, June, 1959, pp. 28–32.

[3] C. C. Yang, J. P. Hervey, and P. F. Smith, "On amplifiers used for microelectrode work," IRE TRANS. ON MEDICAL ELECTRONICS, vol. ME-10, p. 25; March, 1958.

[4] J. Moore and J. H. Gebhart, "Stabilized wide-band potentiometric amplifiers," PROC. IRE, vol. 50, pp.; August, 1962.

bandwidth can be obtained with a second-order amplifier. Moreover, the technique permits specification of maximum bandwidth criteria for different types of amplifier response and permits insight into the performance observed in the practical application of such amplifiers.

First-Order Amplifier

Assuming that the amplifier labeled A in Fig. 1 has infinite input impedance, zero output impedance, and summing currents at the input node,

$$\frac{E_s - E_i}{R_1} = pC_1E_i + pC_2(E_i - E_o) \tag{1}$$

and

$$E_i = E_o/A, \tag{2}$$

where p is the complex Laplace transform variable. Solving (1) and (2) for the closed-loop gain $G(p)$,

$$G(p) = \frac{E_o}{E_s} = \frac{\dfrac{A}{R_1(C_1+C_2)\left(p + \dfrac{1}{R_1(C_1+C_2)}\right)}}{1 - \dfrac{AC_2p}{(C_1+C_2)\left(p + \dfrac{1}{R_1(C_1+C_2)}\right)}}. \tag{3}$$

For first-order amplifier,

$$A = \frac{A_0\omega_2}{p + \omega_2} \tag{4}$$

where A_0 is the gain at zero frequency and ω_2 is the cutoff angular frequency. Let $\omega_1 = 1/R_1(C_1+C_2)$ and and $K = C_2A_0/C_1+C_2$ where, as will be seen, K is the constant which determines the degree of neutralization. Eq. (3) becomes

$$G(p) = \frac{\dfrac{A_0\omega_1\omega_2}{(p+\omega_1)(p+\omega_2)}}{1 - \dfrac{K\omega_2 p}{(p+\omega_1)(p+\omega_2)}}. \tag{5}$$

A greater degree of generalization can be obtained by writing (5) in normalized form. Let $r = \omega_1/\omega_2$ and $q = p/\omega_2$. The ratio r relates the uncompensated bandwidth ω_1 to that of the amplifier alone, ω_2. By substituting $q = p/\omega_2$, the complex frequency variable is scaled to the amplifier intrinsic bandwidth. Eq. (5) becomes

$$G(q) = \frac{\dfrac{A_0 r}{(q+1)(q+r)}}{1 - \dfrac{Kq}{(q+1)(q+r)}} \tag{6}$$

which may be written

$$G(q) = \frac{A_0 r}{q^2 + (1 + r - K)q + r}. \tag{7}$$

A considerable amount of information may be gleaned from (7) without further analysis. The gain at dc is A_0. Eq. (7) represents a second-order closed-loop system with undamped angular frequency equal to $\sqrt{r}$. The damping may be adjusted by varying $1+r-K$, with K the adjustable parameter.

The behavior of the circuit can be understood better in terms of the loci of the poles of (7) in the complex q plane illustrated in Fig. 2.[5] Eq. (6) is written in a form suitable for sketching the root locus. As explained in Truxal,[5] the solution of

$$\frac{Kq}{(q+r)(q+1)} = 1$$

as K varies, can be sketched by inspection, even in cases when the explicit solution is difficult because the left-hand side involves high-order polynomials in the complex q variable. It will be expedient to review the principles used in sketching the root locus in Fig. 2 even though the loci can be solved directly from the denominator of (7).

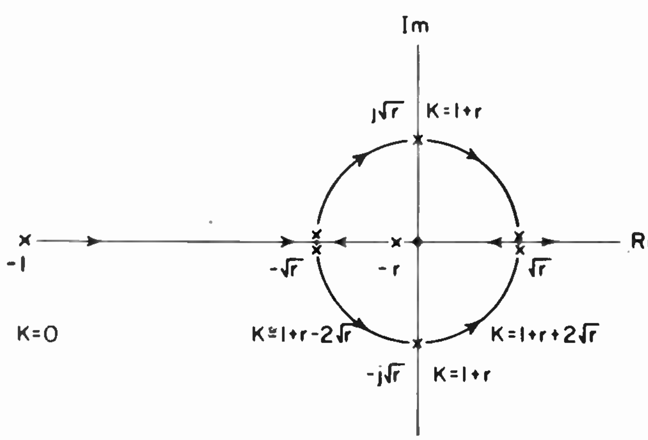


Fig. 2—Root locus in the complex q plane of neutralization system with first-order amplifier. *Abscissa:* Re (real part of q), *Ordinate:* Im (imaginary part of q). Values of K under abscissa of corresponding root values.

The function

$$\frac{q}{(q+r)(q+1)}$$

multiplying K in the denominator of (6) plays the role of an open-loop gain in a feedback system. The poles of this function are the poles of the closed-loop system when K is zero. The following rules for the root loci may be derived on the basis of the algebra of polynomial equations of a complex variable with real coefficients.

[5] J. G. Truxal, "Automatic Feedback Control System Synthesis," McGraw-Hill Book Co., Inc., New York, N. Y., ch. 4; 1955.

1) As K approaches infinity the poles of the closed-loop function terminate on the open-loop zeroes. In (6) there is an explicit zero of

$$\frac{q}{(q+r)(q+1)} \text{ at } q = 0$$

and an implicit zero at infinity.

2) The locus of the poles for a positive feedback system comprise portions of the real axis of the q plane to the left of an even number of poles and zeros and the entire positive real axis.

3) As K approaches infinity the root loci approach straight line asymptotes given by the roots of the equation $q^{n_p - n_z} - K = 0$ where n_p is the number of poles and n_z the number of zeros of the open-loop gain function. In the present case, as K approaches infinity, there is a single asymptote approaching plus infinity along the real axis. If $n_p - n_z = 2$, both positive and negative real axes are asymptotes. For $n_p - n_z = 3$, there are three asymptotes; *i.e.*, the negative real axis and two straight lines making angles of ± 60 degrees to the positive real axis. These asymptotes originate on the real axis at the algebraic centroid of the real part of the poles and zeros.

4) Complex poles occur in complex conjugate pairs so that the root loci are always symmetrical about the real axis of the q plane.

The above four rules will suffice for our purpose. Truxal's book should be consulted for a justification of these rules and for a more complete treatment of the root-locus technique.

In the case of the first-order negative capacitance amplifier illustrated by the root locus in Fig. 2, the open-loop poles are located at $-r$ and -1 and there is a zero at the origin; *i.e.*, at $q = 0$. As K is increased from zero the poles move together until they coalesce at $-\sqrt{r}$, where $K = 1 + r - 2\sqrt{r}$. The poles then travel on a circle in the complex plane reaching the imaginary axis at $K = 1 + r$. At $K = 1 + r + 2\sqrt{r}$ they reach the positive real axis. One travels toward $q = 0$ to cancel the zero at that location and the other approaches infinity along the positive real axis.

This figure correlates well with experimental observations of the behavior of these amplifiers. As K increases, the closed-loop bandwidth improves. Beyond a critical value of K, the transient response becomes underdamped and eventually the circuit breaks into spontaneous oscillation. It can be seen that the maximum bandwidth is equal to $r^{-1/2}$ times the uncompensated bandwidth. Depending on the object of the bioelectric investigation, the parameter K may be adjusted to give critical damping or a slightly underdamped condition.

The results are more striking if typical numbers are substituted for the parameters of the figure. If

$$R_1 = 10^7 \text{ ohms}$$
$$C_1 + C_2 = 20 \times 10^{-12} \text{ farad}$$
$$\omega_2 = 2 \times 10^6 \text{ rps}$$

then

$$\omega_1 = 5 \times 10^3 \text{ rps}$$
$$r = 2.5 \times 10^{-3}$$
$$\sqrt{r} = 0.05.$$

Consequently, a 20-to-1 increase in the uncompensated bandwidth is possible. Using the formula given by Elmore for the rise time of a filter or an amplifier with n identical isolated stages,[6] each with a bandwidth w_b.

$$T_R = \frac{\sqrt{2\pi n}}{\omega_b}; \qquad (8)$$

the uncompensated rise time will be 500 μsec, and the compensated rise time will be 35.3 μsec compared to 1.3 μsec for the amplifier alone. For the critically damped adjustment, the compensated system behaves as two identical isolated stages. For a first-order amplifier the rise time of the system equals $\sqrt{2r}$ times the uncompensated rise time.

The change in K producing varying degrees of neutralization is especially significant. When the two poles coalesce, the system is critically damped and $K = 0.946$. When the two poles touch the imaginary axis, the system reaches the borderline of stability. Then $K = 1.0025$, a change of only 6 per cent from the critically damped condition. These figures emphasize the need to stabilize the capacitance ratio and the zero frequency gain A_0. The variation in K may be obtained by means of a precision potentiometer at the output of the amplifier.

Second-Order Amplifier

An amplifier with more than one stage of amplification or one utilizing a low-pass filter as an interstage coupling network may have a gain function involving at least the square of the complex variable p. A generic form for the gain of a second-order amplifier is

$$A = \frac{A_0\omega_n^2}{p^2 + 2\delta\omega_n p + \omega_n^2}. \qquad (9)$$

The form of (9) can be realized using negative feedback around two stages, each with a 6-db high-frequency cutoff characteristic. Alternatively, (9) can be realized, as discussed in Appendix I, by combined positive feedback around one stage and negative feedback around two stages. The form of (9) includes all of the response possibilities for a second-order amplifier. For $\delta > 1$,

[6] W. C. Elmore and M. L. Sands, "Electronics—Experimental Techniques," McGraw-Hill Book Co., Inc., New York, N. Y.; 1949.

the amplifier is overdamped and for $\delta=1$, critically damped.

Substituting (9) in (3) with $\omega_1=1/R_1(C_1+C_2)$ and $K=C_2A_0/C_1+C_2$, as before,

$$G(p) = \frac{\dfrac{A_0\omega_1\omega_n^2}{(p+\omega_1)(p^2+2\delta\omega_n p+\omega_n^2)}}{1-\dfrac{K\omega_n^2 p}{(p+\omega_1)(p^2+2\delta\omega_n p+\omega_n^2)}}. \tag{10}$$

Eq. (10) can be put into normalized form by letting $q=p/\omega_n$ and $r=\omega_1/\omega_n$. Then,

$$G(q) = \frac{\dfrac{A_0 r}{(q+r)(q^2+2\delta q+1)}}{1-\dfrac{Kq}{(q+r)(q^2+2\delta q+1)}}. \tag{11}$$

The root locus for the poles of $G(q)$ in (11) is more complicated than for (6) because of the cubic in the denominator and the variable parameter δ.

As shown in Appendix II, the maximum bandwidth with the system critically damped is obtained for a particular value of δ; namely,

$$2\delta = 3r^{1/3} - r. \tag{12}$$

Since with $r<1$, δ is less than unity, this adjustment corresponds to an underdamped second-order amplifier.

The root locus of the poles of (11) with K varying is shown in Fig. 3 for the value of δ determined in (12). For $K=0$, the "open-loop gain" has a zero at the origin, a pole at $-r$ and a pair of complex conjugate poles at $-\delta\pm j\sqrt{1-\delta^2}$. As K is increased from zero, the two conjugate poles approach the negative real axis and coalesce with the pole moving to the left from $-r$. The three poles meet at $-r^{1/3}$ for

$$K = 1 + 2\delta r - \frac{(r+2\delta)^3}{2}$$

with δ taking on the value given by (12). As K is increased further, two of the poles become complex and approach the imaginary axis along the curved path shown in the figure. The third pole continues to move to the left along the real axis. Eventually the complex poles cross over into the right-hand half plane, and coalesce on the positive real axis. One goes to the left toward zero, the other to the right toward positive infinity. Meanwhile, as K increases, the negative pole moves to the left along the real axis toward negative infinity.

The system is critically damped when K is adjusted to give a triple pole at $-r^{1/3}$. The system can also be adjusted for an underdamped condition by increasing K. In the case of a second-order amplifier, the path of the poles in the complex plane is no longer circular so that

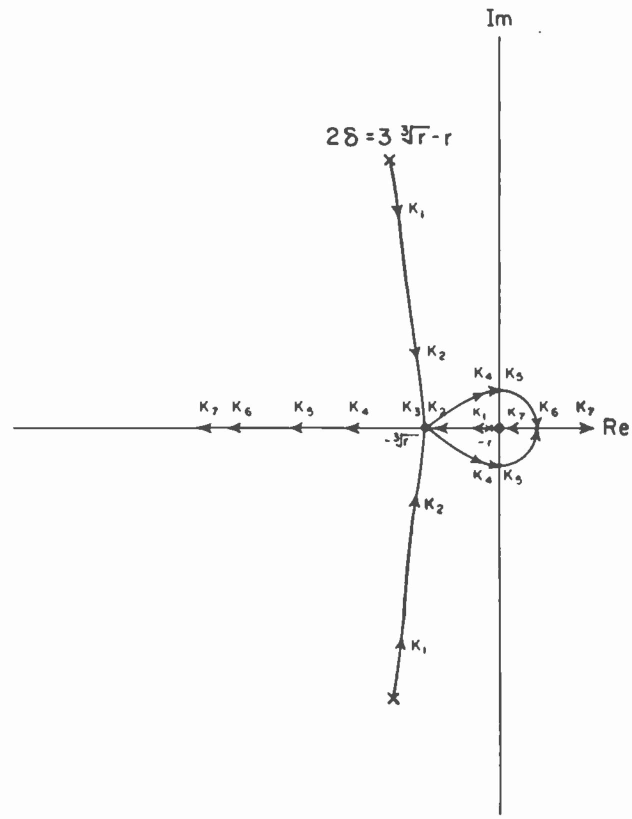


Fig. 3—Root locus of neutralization system with second-order amplifier. Same axes as in Fig. 2. Successive values of K above locus of corresponding root values. Arrows in direction of pole migration with K increasing.

the underdamped response has lower bandwidth than for a critically damped adjustment.

The results of this analysis indicate a significant increase in bandwidth and decrease in rise time over that possible with a first-order amplifier. The maximum bandwidth is $r^{-2/3}$ times the bandwidth of the uncompensated input network as compared with $r^{-1/2}$ times for the first-order amplifier. Using the figure given in the example of the last section, the bandwidth is increased by a factor of 54 instead of 20. The improvement can be evaluated more realistically by considering the decrease in rise time. At the critical value of K, the system corresponds to a third-order amplifier with three identical isolated stages. Using (8), the rise time is $3^{1/2}r^{2/3}$ times the rise time of the uncompensated system compared with $2^{1/2}r^{1/2}$ times for the amplifier of the previous section. Thus, for an uncompensated rise time of 500 μsec, the compensated system has a rise time of 35.3 μsec if a first-order amplifier is used, and a rise time of 16.0 μsec if a second-order amplifier is used. The first-order amplifier alone has a rise time of 1.3 μsec as compared with 1.84 μsec for the second-order amplifier with the input network removed. These figures may be conservative because it should be possible to attain higher bandwidth for a second-order amplifier with the same over-all gain as a first-order amplifier.

The variation in K necessary to adjust the system from the maximally flat condition to the borderline of stability is comparable to that of the first-order amplifier. In the example given, $K=0.946$ for maximal flatness and $K=0.995$ for undamped oscillations, a change of 5 per cent.

Noise

Neutralized input capacity amplifiers have an extremely large noise figure.[1,2] The use of amplifiers with more than a 20 db per decade slope in their high-frequency cutoff characteristic may enhance the noise. One writer has suggested the use of a filter following the amplifier as a possible way to reduce the extra noise.[1] A quantitative noise analysis is needed to assess these possibilities.

The circuit of Fig. 4 represents a model of the system including the predominant noise voltage sources. The voltage e_1 represents the root-mean-square noise voltage associated with the resistance R_1. The input tube noise is represented by an equivalent resistance R_{eq} and source e_2 in series with the input line. The amplifier symbol labeled B represents a filter cascaded with the system.

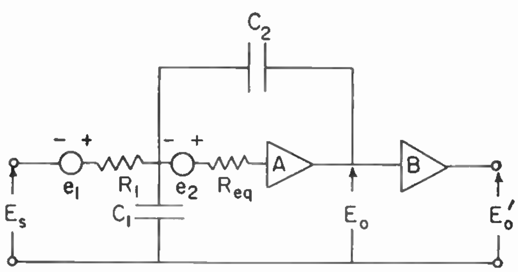


Fig. 4—Circuit model for analysis of noise in neutralized input capacity system. Symbols as in Fig. 1 except R_{eq} = amplifier noise equivalent resistance, e_1 = noise voltage source associated with R_1, e_2 = noise voltage source associated with R_{eq}, B = filter transfer function, E_o' = filter output voltage.

The circuit of Fig. 4 can represent a number of different systems depending on the complexity of the amplifier and filter. As a basis of comparison we shall assume that all of these can be adjusted to have a critically damped response. The poles of the system gain function are coincident of order $m+\alpha+1$ where m and α respectively are the orders of the amplifier and filter cutoff characteristics.

Under these conditions, the rise time of the system is identical with that of $m+\alpha+1$ identical amplifier stages and is given by (8).[6] It is then possible to specify the rise time and to compare the noise figure of the systems as a function of their complexity and the relative role of the amplifier and filter.

The noise voltage e_N referred to the input of the system of Fig. 4 is[7]

$$\frac{e_N^2}{4KT} = R_1\Delta f_1 + R_{eq}\Delta f_2 \qquad (13)$$

[7] M. Schwartz, "Information Transmission Modulation and Noise," McGraw-Hill Book Co., Inc., New York, N. Y.; 1959.

where

K is the Boltzmann constant
T is the absolute temperature

and

$$\Delta f = \int_0^\infty \left|\frac{G(f)}{G(0)}\right|^2 df. \qquad (14)$$

The gain $G(f)$ is defined with respect to the noise source voltage and the output of the system. $G(0)$ is the zero frequency value of the gain, assuming that these amplifiers are dc coupled. When the filter is included, its transfer function $B(f)$ will be multiplied by $G(f)$ to obtain the over-all gain. The amplification of the noise source e_1 is the same as was derived for the signal E_s in the previous sections. The Laplace Transform of the gain function with respect to the tube noise source e_2 is

$$G_2(p) = \frac{A(p+\omega_1)}{p+\omega_1 - \dfrac{AC_2}{C_1+C_2}p}. \qquad (15)$$

The numerator term in (15) indicates the effect of positive feedback and produces a rising frequency characteristic. It is convenient to normalize the terms in (14) and (15); *i.e.*, let

$$r = \frac{\omega_1}{\omega_2}$$

for

$$p = j\omega$$

let

$$\omega = (\omega_1\omega_2^n)^{(1/(m+1))}x = \omega_1 r^{(-m/(m+1))}x \qquad (16)$$

then

$$df = \frac{d\omega}{2\pi} = \frac{\omega_1}{2\pi} r^{(-m/(m+1))}dx.$$

The following changes in notation from the preceding sections are made. Let

$\omega_2^V = \omega_2$ for $V = 1$ (first-order amplifier)
$\omega_2^V = \omega_n^2$ for $V = 2$ (second-order amplifier)
$\omega_2^V = \omega_2^3$ for $V = 3$ (third-order amplifier).

In order to adjust all the possible systems for critical damping, the cutoff value of the high-frequency asymptote is made equal to

$$\omega_1 r^{(-m/(m+1))}.$$

The order of the composite system excluding the filter is $m+1$, and the filter is assumed to have a cutoff characteristic of 20α db per decade. With these assumptions and the normalizations given, adjusting A_oC_2/C_1+C_2 in (5) and (15) to give critical damping, one obtains

$$\left|\frac{G_1(x)}{G_1(0)}\right|^2 = \frac{1}{(1+x^2)^{m+1}}$$

$$\left|\frac{G_2(x)}{G_2(0)}\right|^2 = \frac{1 + x^2 r^{(-2m/(m+1))}}{(1+x^2)^{m+1}}$$

$$\left|\frac{B(x)}{B(0)}\right|^2 = \frac{1}{(1+x^2)^{\alpha}}. \tag{17}$$

The integrals obtained when equations (17) are substituted in (14) are standard.[8] Consequently

$$\Delta f_1 = \omega_1 r^{(-m/(m+1))} \left[\frac{[2(m+\alpha)]!}{4^{m+\alpha+1}[(m+\alpha)!]^2}\right]$$

$$\Delta f_2 = \Delta f_1 (1 + r^{(-2m/(m+1))})$$

$$\cong \Delta f_1 r^{(-2m/(m+1))} \qquad \text{for } r \ll 1. \tag{18}$$

The rise time of the critically damped systems may be expressed,

$$\frac{\omega_1 T_R}{\sqrt{2\pi}} = \tau = (m+\alpha+1)^{1/2} r^{(m/m+1)}. \tag{19}$$

Solving (19) for r, substituting for r in (18) and then substituting the results in (13), an expression is obtained relating the noise voltage e_N to the rise time τ; *i.e.*,

$$e_N = \frac{K_{m+\alpha}}{\left(\frac{C_T}{4KT}\right)^{1/2}} \tau^{-1/2}[1 + (m+\alpha+1)\Re\tau^{-2}]^{1/2} \tag{20}$$

where

$$C_T = C_1 + C_2$$

$$\Re = \frac{R_{eq}}{R_1}$$

and

$m+\alpha$	$K_{m+\alpha}$
1	0.423
2	0.403
3	0.398
4	0.393

Eq. (20) summarizes the performance of the critically damped system with respect to noise. For $m+\alpha=1$, (20) is the same equation presented by Guld, but expressed differently.[2] The variation of $K_{m+\alpha}$ is small enough (8 per cent) to be neglected for all practical purposes. In Fig. 5, we have plotted

$$20 \log_{10} \frac{e_N}{K_{m+\alpha}} \left(\frac{C_T}{4KT}\right)^{1/2}$$

vs τ^{-1} to illustrate the behavior.

[8] B. O. Pierce and R. M. Foster, "A Short Table of Integrals," Ginn and Co., Boston, Mass., 4th ed., No. 53—p. 9, No. 59—p. 10, No. 495—p. 67; 1956.

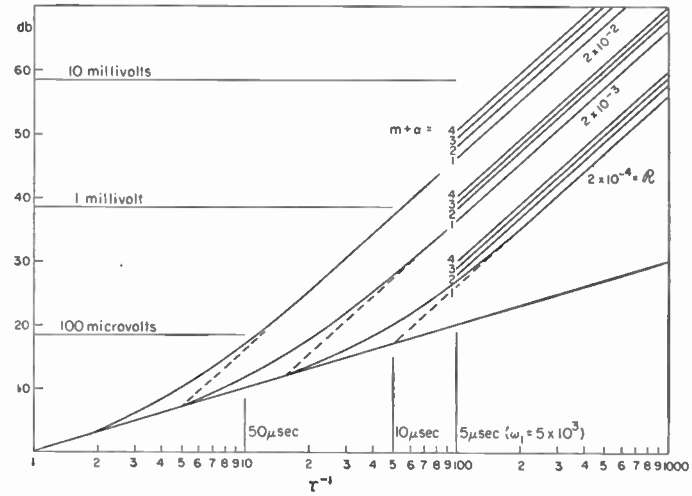


Fig. 5—Family of curves giving relative system noise in db vs the reciprocal of rise time τ^{-1}. Plot of (20). *Ordinate:*

$$\text{db} = 20 \log_{10} \frac{e_N}{K_{m+\alpha}} \left(\frac{C_T}{4KT}\right)^{1/2}.$$

Abscissa: τ^{-1}. Excess noise becomes predominant when $(m+\alpha+1)\Re = \tau^2$, the intersection of asymptotes shown by the dotted lines. $\Re = R_{eq}/R_1$, m = cutoff slope of A, α = cutoff slope of B. Labeled voltages and rise times for $C_T = 20\times10^{-12}$ f, $T = 293°$K, $\omega_1 = 5\times10^3$ rps, $K_{m+\alpha} = K_1 = 0.423$.

The labeled noise voltages and rise times refer to the example cited in a previous section; *i.e.*,

$$C_T = 20 \times 10^{-12} f$$

$$m + \alpha = 1$$

$$\omega_1 = 5 \times 10^3 \text{ rps.}$$

With $R_{eq} = 20K\Omega$, $\Re = 2\times10^{-3}$. The conditions are then the same as those discussed by Guld.[2] It should be noted that the behavior can be approximated by the intersection of two semilogarithmic straight lines which intersect at $(m+\alpha+1)\Re$ equal to τ^2. For values of τ^{-1} less than this, the noise amplitude rises at 10 db per decade. In this region the noise is contributed by the microelectrode resistance alone. Beyond this junction point, the noise amplitude increases at 30 db per decade and the increase in rate reflects the system noise figure which is equal to $1+(m+\alpha+1)\Re\tau^{-2}$.

For the example, the noise voltage is 48 μv, at a rise time of 33 μsec (at the junction point of the asymptotes). The noise rises to 1.05 mv for a rise time of 3.3 μsec. If the same rise time (3.3 μsec) were achieved by reducing the physical capacitance in the circuit, the noise voltage would be 105 μv. The neutralization system brings about a 10-fold increase in noise over that of a passive circuit.

The price of system complexity $(m+\alpha)$ is increased noise. The greater the reduction in rise time, the higher is the excess noise. The filter B is indistinguishable from a system with a higher-order amplifier, and brings about an increase in noise for the same effective rise time achieved by a simpler system. The only reason one might go to a higher-order system is to achieve a lower rise time than is otherwise possible. The increase in noise for a unit increase in system complexity is relatively moderate as seen in the figure. In certain applications one may be able to tolerate the noise in exchange for a reduction in distortion. It is clear from the analysis that the best design from the signal-to-noise point of view is the simplest system adjusted to have the longest rise time consistent with waveform fidelity.

Discussion

Criteria for the performance of negative capacitance amplifiers have been developed in this paper on the basis of idealized circuit models and the root locus of the system gain function. Many of the neutralized input capacity amplifiers in use can be reduced to the two basic forms we have considered. More complex amplifiers are amenable to analysis by studying the root locus of the system gain function, although it may be difficult to formulate maximum bandwidth criteria for them. A preliminary analysis has shown that a shunt-peaked amplifier will yield the same bandwidth as the second-order amplifier discussed. The problem of neutralizing the input capacity of a measurement system for bioelectric waveforms is seen to be similar in principle to the problem of compensating the system function of a servomechanism.[5]

We have shown that markedly faster response is possible for a second-order amplifier than for a first-order amplifier. For the best simultaneous adjustment of the damping constant δ and the compensation constant K, a second-order amplifier may achieve a bandwidth $r^{-2/3}$ times the uncompensated bandwidth of the input circuit as compared with $r^{-1/2}$ for the first-order amplifier.

Both system responses are sensitive to small variations in K, a 5 per cent change being sufficient to carry the system from a critically damped adjustment to instability. In Appendix I, it will be pointed out that it is difficult in practice to vary both δ and K independently to achieve the best adjustment for a second-order amplifier.

It is believed that the root-locus technique provides a unified conceptual framework for studying input capacity neutralization circuits. It helps provide design criteria and a theoretical basis for evaluating experimental observations on the performance of practical circuits.

A study of the noise contributed in neutralized input capacity amplifiers, adjusted for critical damping, indicates that the noise voltage amplitude increases with the system complexity. The noise figure of such systems is $1+(m+\alpha+1)\mathcal{R}\tau^{-2}$. In this expression, the crucial terms are contributed by the rise time of the system and the equivalent noise resistance R_{eq} of the input tube. The addition of a filter cascaded with the amplifier increases the noise, through the factor $m+\alpha+1$, for the same over-all system rise time.

Appendix I
A Second-Order Amplifier

Moore describes an amplifier with combined positive and negative feedback similar to the one illustrated in Fig. 6.[4] The gain A of that amplifier is

$$A = \frac{A_1 A_2}{1 - \alpha A_1 + \beta A_1 A_2}. \tag{21}$$

For simplicity, assume

$$A_1 = A_2 = \frac{A_{12}\omega_2}{p + \omega_2}. \tag{22}$$

Then

$$A = \frac{A_{12}^2\omega_2^2}{p^2 + (2 - \alpha A_{12})\omega_2 p + (1 - \alpha A_{12} + \beta A_{12}^2)\omega_2^2}. \tag{23}$$

Let

$$\omega_n^2 = \omega_2^2(1 - \alpha A_{12} + \beta A_{12}^2)$$

$$2\delta = \frac{2 - \alpha A_{12}}{\sqrt{1 - \alpha A_{12} + \beta A_{12}^2}}$$

$$A_0 = \frac{A_{12}^2}{1 - \alpha A_{12} + \beta A_{12}^2}. \tag{24}$$

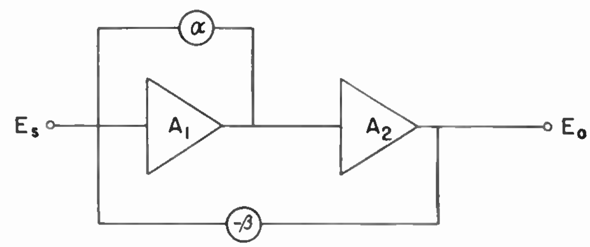


Fig. 6—Second-order amplifier with combined positive and negative feedback. A signal flow diagram.

Eq. (23) becomes identical with (9) in the text.

In principle, it is possible to adjust δ and A_0 by varying α and β in the amplifier. However, it is desirable to adjust $\alpha A_{12} = 1$ and $\beta A_{12}^2 \gg 1$ to make the gain with feedback insensitive to variations in the gain of the separate stages. If this is done, (24) becomes

$$\omega_n^2 = \omega_2^2 \beta A_{12}^2$$

$$2\delta = \frac{1}{\sqrt{\beta A_{12}^2}}$$

$$A_0 = 1/\beta. \tag{25}$$

Moreover,

$$K = \frac{C_2 A_0}{C_1 + C_2} = \frac{C_2}{(C_1 + C_2)\beta}. \tag{26}$$

In practice, even if K is varied by means of a potentiometer outside the feedback loops, the range of this control is changed as δ is adjusted. Moreover, ω_n is dependent on the setting of the feedback factor β.

It seems evident that the proper procedure is to adjust $\alpha A_{12}=1$ and to permit a range of control for β to allow the amplifier response with the input circuit removed to go from a critically damped condition to a markedly underdamped one. Then, with the input circuit connected, both δ and K may be varied by adjusting β and the external control to obtain best response.

For the example in the text, $2\delta=0.4$ and $\beta A_{12}^2=6.25$.

Appendix II

Root Locus with δ as a Parameter

In Fig. 7, the root locus of the system characterized by (11) is illustrated with δ as a parameter. Basically this sketch is a solution of the paths of the roots of the cubic

$$q^3+(r+2\delta)q^2+(1+2\delta r-K)q+r=0 \qquad (27)$$

with δ as a parameter and K as a variable.

Generally, these paths follow the rules described in the text. The case illustrated in Fig. 3 corresponds to δ_2 in Fig. 7. However, it remains to show the form for the loci for values of δ greater than or less than δ_2 and to prove that maximum bandwidth is achieved for $\delta=\delta_2$.

The figure indicates that for $\delta<\delta_2$, the paths of the complex roots do not touch the negative real axis but cross the imaginary axis into the right-hand half plane. For $\delta>\delta_2$ two double roots exist in the left-hand half plane. The complex roots coalesce at a point more negative than $-r^{1/3}$, but the root moving to the left from $-r$ does not reach $-r^{1/3}$. The two real roots move in different directions and the one moving to the right coalesces with the root moving to the left from $-r$. These two roots traverse a curved path in the complex plane illustrated by the small circle in the figure. Consequently, for $\delta=\delta_2$, all three roots are simultaneously as far to the left in the complex plane as is possible. This is the condition for maximum bandwidth. In the other cases, a smaller absolute magnitude for one root corresponds to a lower cutoff frequency.

The nature of the roots of the cubic is determined by whether or not the discriminant

$$\frac{(r+2\delta-x)^2}{2} \gtreqless \frac{r}{x} \qquad (28)$$

where x is the negative of the real root of the cubic and $r+2\delta-x$ is the negative of the sum of the real parts of the remaining two roots. Fig. 8 is a plot of the equation

$$y=x\,\frac{(r+2\delta-x)^2}{2}-r. \qquad (29)$$

On the basis of the conditions of (28), for $y\geq 0$ there are three real roots of the cubic. For $y<0$ a pair of complex conjugate roots exist and double or triple real

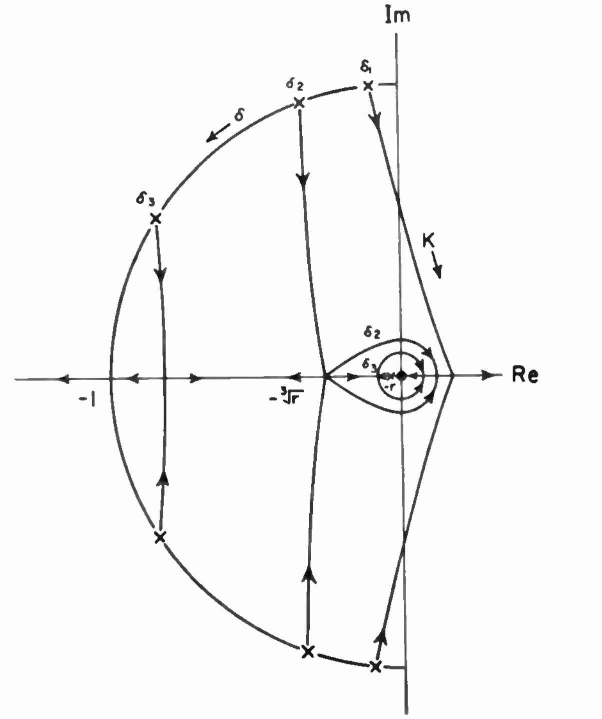


Fig. 7—Family of root loci of the second-order system with δ as a parameter. Axes as in Fig. 2.

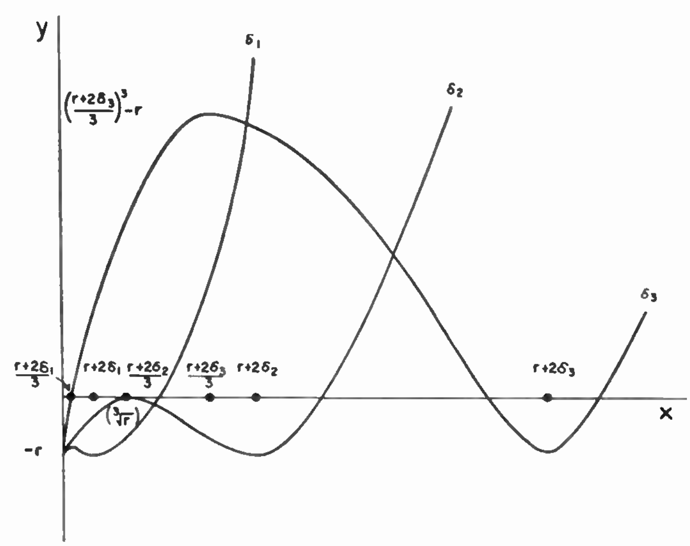


Fig. 8—Family of discriminant equations to the cubic with δ as a parameter. *Abscissa:* The amplitude of the negative of the real root x. *Ordinate:* Positive for three real roots, negative for a pair of complex conjugate roots, zero for multiple roots y. Plot of the equation

$$y=\frac{x(r+2\delta-x)^2}{2}-r.$$

roots occur for $y=0$. Eq. (29) has extrema at

$$x = \frac{(r+2\delta)}{3} \quad \text{and} \quad x = r + 2\delta$$

so that it is necessary for

$$y > 0 \text{ at } x = \frac{r+2\delta}{3}$$

for three real roots to occur.

From this figure, it is seen that

$$\frac{r+2\delta_2}{3} = r^{1/3}$$

provides the boundary for the existence of more than one double root for (27). For $\delta<\delta_2$, a double root exists with $x>r+2\delta$, but this root has a positive real part equal to

$$\frac{-(r+2\delta-x)}{2}.$$

For $\delta=\delta_2$ a triple root exists at $x=(r+2\delta)/3=r^{1/3}$. For $\delta>\delta_2$ three double roots exist. Two of these are in the left-hand q plane. However, from (27) it is seen that the product of the real roots must equal r independently of the value of δ and K. Obviously, for $\delta>\delta_2$, one of the roots must always be less in magnitude than $r^{1/3}$, the magnitude of the triple root when $\delta=\delta_2$. The conclusions have been verified by computation of sample points along the curves for δ_1, δ_2, and δ_3 in Fig. 7. The illustrations are distorted in relative scale for clarity in the presentation.

On the Reception of Quasi-Monochromatic, Partially Polarized Radio Waves*

H. C. KO†, SENIOR MEMBER, IRE

Summary—**The response of an elliptically polarized antenna to a quasi-monochromatic, partially polarized radio wave is treated from the standpoint of coherence theory. A general formula is derived for the available power at the terminals of a receiving antenna in terms of appropriately chosen coherency matrices for the antenna and the incident wave. It is shown that the result is formally identical with the basic interference law of partially coherent, quasi-monochromatic wave fields. Conditions for the maximum available power are discussed, and a geometrical interpretation of the result is given. The coherency matrix for the antenna is determined uniquely in terms of the transmitting properties of the antenna.**

It is shown that the coherency matrix formalism for the interaction of an antenna with an incident wave fits very we l with those used in the modern theory of optics and quantum mechanics.

A new definition for the antenna effective aperture using the coherency matrix is suggested, which takes into account the polarization property of the antenna.

Introduction

IN THE ANALYSIS of radio antennas one normally assumes that the incident radio wave is completely polarized (*i.e.*, elliptically polarized). The response of an elliptically polarized receiving antenna to an elliptically polarized radio wave incident upon the antenna has been thoroughly discussed in the antenna literature [1]–[8]. A completely polarized wave is a limiting case of a more general type of wave. that is, a partially polarized wave. However, the treatment on the response of an antenna to partially polarized radio waves has been neglected until recently. The use of radio antennas for the reception of partially polarized radio waves occurs in radio astronomy, microwave plasma diagnostics, passive radar mapping, etc. Thus, a detailed analysis of the interaction of an elliptically polarized antenna with a quasi-monochromatic, partially polarized wave is of interest both from theoretical and practical points of view.

The purpose of this paper is to treat the response of an elliptically polarized antenna to a partially polarized incident radio wave from the standpoint of coherence theory. Although this problem may be treated in different ways [9]–[11], there are several reasons why an analysis within the framework of coherence theory is desirable. During the last few years, the concept of partial coherence and its attendant theory have become increasingly important in many branches of physics, particularly in optics. Space-time correlation functions were introduced by Wolf [12] to formulate a large branch of optics in terms of observable quantities. The coherence theory [13] has since been successfully used to formulate theories of partial coherence, partial polariza-

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† Radio Observatory, Department of Electrical Engineering, The Ohio State University, Columbus, Ohio.

tion, and the interaction of certain optical devices. At present, when microwave engineering has been extended to the near optical wavelengths, the formulation of an antenna theory consistent with that of the modern optical theory and the quantum mechanical treatment of photons will lead to a better understanding of the antenna itself as well as its relation to other physical devices.

Quasi-Monochromatic Partially Polarized Waves

It is well known that a rigorously monochromatic electromagnetic wave is completely polarized. A completely polarized wave belongs to, in general, an elliptically polarized type which includes linear and circular polarizations as special cases. The problem of determining the state of polarization of a completely polarized wave is familiar to radio physicists and engineers and is thoroughly discussed in radio and optical literature. For a rigorously monochromatic wave, the end point of the electric field vector traces out an ellipse, which may reduce, in special cases, to a straight line or to a circle. Thus the wave may be elliptically, linearly or circularly polarized.

However, if one considers a more realistic case of a beam of electromagnetic waves due to natural radiation, it is never rigorously monochromatic. Even the sharpest spectral line from a real physical source has a finite frequency bandwidth. We shall restrict ourselves to the case of a quasi-monochromatic plane wave, *i.e.*, a wave whose spectral components cover a frequency range Δf which is very small compared with the mean frequency f (*i.e.*, $\Delta f \ll f$). Such a quasi-monochromatic wave may result from the superposition of a large number of randomly timed statistically independent pulses with the same mean frequency of oscillations. The concept of partial polarization becomes significant when dealing with such statistical radiation. The end point of the electric field vector of a quasi-monochromatic wave traces out, in general, an ellipse whose shape changes continuously. When the ellipse maintains a constant orientation, axial ratio, and the sense in which the ellipse is described, in spite of continuous fluctuations of its size, the wave is said to be completely polarized. On the other hand, the end point of the field vector may move completely irregularly, and we may say that the wave is randomly polarized. Between these two extreme cases, we have a partially polarized wave which shows neither completely regular nor completely irregular variation in the trace of the end point of the electric field vector.

The first systematic investigation of partially polarized light was made by Sir George Stokes in 1852. Stokes [14] has shown that a partially polarized electromagnetic wave may be completely characterized by four parameters which are today known as the Stokes parameters. These parameters may be derived from the electric field associated with the wave. A new treatment of the partial polarization was later originated by Wiener [15] and Perrin [16] who employed the concept of correlation functions between the two components of the electric field vector in two mutually perpendicular directions at right angles to the direction of the wave propagation. Wiener showed that the state of wave polarization may be completely characterized by a 2×2 matrix which he called the coherency matrix. Landau and Lifshitz [17] showed that a quasi-monochromatic partially polarized wave may be characterized by the quadratic functions of its field components in a tensor form.

The most complete and systematic analysis on the coherence theory of quasi-monochromatic, partially polarized electromagnetic radiation is due to Wolf [18]. More recently, Parrent and Roman [19] formulated the effect of the interaction of optical devices with a partially polarized quasi-monochromatic plane wave from the standpoint of coherence theory.

Let us consider a quasi-monochromatic plane wave in the frequency interval $(f\pm\Delta f/2)$ and let $E_x(t)$ and $E_y(t)$ represent, in the usual rectangular coordinates, the components of the electric field vector $\boldsymbol{E}$ in the two mutually orthogonal directions at right angle to the direction of wave propagation (positive z-direction). Although we are dealing with a physical field which is real, it is convenient to employ a complex representation. Following Wolf [13], we shall use Gabor's analytic signal [20] as the complex representation of a real signal. The analytic signal representation is a generalization of the conventional use of $\exp(j\omega t)$ in the place of $\cos \omega t$ or $\sin \omega t$ familiar in radio physics and engineering. The analytic signal associated with a real function is uniquely defined so that its real part is equal to the real function, while its imaginary part is in quadrature with its real part. The real and imaginary parts of an analytic signal are conjugate functions and may be shown to be Hilbert transforms of each other [20]. The analytic signal for a real, quasi-monochromatic wave may be uniquely represented in the form

$$A(t)e^{j[\omega t+\alpha(t)]} \tag{1}$$

where $A(t)\geq 0$ and $\alpha(t)$ are both real. The spectral components of the wave will be appreciable only near the mean angular frequency ω. $A(t)$ and $\alpha(t)$ will vary slowly, at a rate set by the bandwidth Δf, in comparison, with $\cos \omega t$ and $\sin \omega t$. Thus, we write

$$\boldsymbol{E}(t) = E_x(t)\boldsymbol{u}_x + E_y(t)\boldsymbol{u}_y \tag{2}$$

where

$$E_x(t) = A_1(t)e^{j[\omega t-kz+\alpha_1(t)]}$$
$$E_y(t) = A_2(t)e^{j[\omega t-kz+\alpha_2(t)]}$$

$\boldsymbol{u}_x$ and $\boldsymbol{u}_y$ are unit vector in the x and y directions, respectively. We assume that the direction of wave propagation is along the positive z direction. For a strictly monochromatic wave, A's and α's become constants.

The degree of correlation between the orthogonal components $E_x(t)$ and $E_y(t)$ determines the state of wave polarization. The degree of correlation may be represented by a quantity called the complex correlation factor μ_{xy} defined by [18]

$$\mu_{xy} = \frac{\langle E_x(t)E_y^*(t)\rangle}{\sqrt{\langle E_x(t)E_x^*(t)\rangle}\sqrt{\langle E_y(t)E_y^*(t)\rangle}} \tag{3}$$

where the symbol * represents the complex conjugate and the angular brackets $\langle \cdots \rangle$ represent the time averaging operation defined by

$$\langle \cdots \rangle = \lim_{T\to\infty} \frac{1}{2T}\int_{-T}^{+T} \cdots dt \tag{4}$$

μ_{xy} is in general a complex quantity. It can be shown from Schwartz' inequality that $|\mu_{xy}| \le 1$. The absolute value of μ_{xy} is a measure of the degree of correlation between the x and y components of the electric field vector, while the phase angle of μ_{xy} is a measure of their effective phase difference. Thus, the wave is completely polarized when $|\mu_{xy}|$ is unity; and it is completely unpolarized (or randomly polarized) when $|\mu_{xy}|$ is zero and furthermore $\langle E_xE_x^*\rangle = \langle E_yE_y^*\rangle$. The wave is said to be partially polarized when $|\mu_{xy}|$ is between zero and unity. The phase angle of μ_{xy} is related to the shape and the sense in which the ellipse is traced. As shown by Wolf [18], this complex correlation factor has a significance similar to the complex degree of coherence in the optical theory of interference and diffraction.

The state of wave polarization may be completely characterized by the following 2×2 matrix.

$$\begin{bmatrix} \langle E_xE_x^*\rangle & \langle E_xE_y^*\rangle \\ \langle E_x^*E_y\rangle & \langle E_yE_y^*\rangle \end{bmatrix} \tag{5}$$

Wolf [18] called it the coherency matrix and proved that the elements of this matrix are observables of the radiation field.

We shall write this matrix in a slightly different form as

$$P[\rho_{ij}] = P\begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix} = \frac{1}{2Z_0}\begin{bmatrix} \langle E_xE_x^*\rangle & \langle E_xE_y^*\rangle \\ \langle E_x^*E_y\rangle & \langle E_yE_y^*\rangle \end{bmatrix} \tag{6}$$

where P = power density of the wave (w/m$_2$ in mks units)

$$P = \frac{1}{2Z_0}\langle \boldsymbol{E}\cdot\boldsymbol{E}^*\rangle = \frac{1}{2Z_0}\{\langle E_xE_x^*\rangle + \langle E_yE_y^*\rangle\}\cdot$$

Z_0 = intrinsic impedance of free space $= \sqrt{\mu_0/\epsilon_0} = 120\pi$ ohms. From (6) it is obvious that

$$\begin{aligned} \rho_{11} &= \langle E_xE_x^*\rangle/2Z_0P, & \rho_{12} &= \langle E_xE_y^*\rangle/2Z_0P \\ \rho_{21} &= \langle E_x^*E_y\rangle/2Z_0P, & \rho_{22} &= \langle E_yE_y^*\rangle/2Z_0P. \end{aligned} \tag{7}$$

It can be shown easily from (7) that the elements of the matrix have the following properties:

$$\rho_{11} \ge 0,\ \rho_{22} \ge 0,\ \rho_{12} = \rho_{21}^*,\ \rho_{11} + \rho_{22} = 1 \tag{7a}$$

$$\rho_{11}\rho_{22} - \rho_{12}\rho_{21} \ge 0. \tag{7b}$$

The equal sign in (7b) can hold when and only when the wave is completely polarized. The matrix $|\rho_{ij}|$ is Hermittian since $\rho_{ij} = \rho_{ji}^*$ for all i and j, and is normalized in the sense that the trace of the matrix (*i.e.*, the sum of the principal diagonal terms) is unity.

It is easy to see that the following matrices

$$\begin{bmatrix} \frac{1}{2} & 0 \\ 0 & \frac{1}{2} \end{bmatrix},\ \begin{bmatrix} \frac{1}{2} & +j\frac{1}{2} \\ -j\frac{1}{2} & \frac{1}{2} \end{bmatrix},\ \begin{bmatrix} \frac{1}{2} & -j\frac{1}{2} \\ +j\frac{1}{2} & \frac{1}{2} \end{bmatrix},$$

represent a randomly polarized wave, a right handed circularly polarized wave and a left handed circularly polarized wave, respectively.

It should be noted that the elements of the coherency matrix will change under rotation of the axes of coordinates. However, both the trace of the matrix and its associated determinant are invariant under rotation of the axes.

In defining the coherency matrix (6) we have so far referred to a wave expressed in (2), which propagates in the positive z direction. If the wave is propagating in the negative z direction, the appropriate matrix for the wave will be $P[\tilde{\rho}_{ij}]$, the transpose of $P[\rho_{ij}]$. The reason is as follows: we distinguish two kinds of polarization, right-handed or left-handed, according to whether the rotation of the end of the electric field vector and the direction of wave propagation form a right-handed or left-handed screw (it should be mentioned here that this definition adopted by radio physicists and engineers is just the opposite of the definition traditionally used in optics). Thus, if the coherency matrix $P[\rho_{ij}]$ represents a right-handed polarization for a wave propagating in positive z direction, the same matrix represents a left-handed polarization for this wave propagating in the opposite direction. The transpose of the matrix requires the interchange of the off-diagonal terms, *i.e.*, ρ_{12} and ρ_{21}. Since $\rho_{12} = \rho_{21}^*$, this results in the change of signs in the phase angle of the off-diagonal terms only. The effect of the sign changes is only to reverse the sense in which the polarization ellipse is described.

Reception of Quasi-Monochromatic, Partially Polarized Radio Waves

Let us consider a receiving antenna located at the origin of the usual spherical coordinates $(r,\ \theta,\ \Phi)$. A quasi-monochromatic, partially polarized plane wave is incident upon the antenna from the direction (θ, Φ). Let $\boldsymbol{E}_i(t)$ represent the electric field of the incident wave and we shall write it in the following form:

$$\boldsymbol{E}_i = E_\theta\, \boldsymbol{u}_\theta + E_\Phi\, \boldsymbol{u}_\Phi \tag{8}$$

where

$$\begin{aligned} E_\theta &= A_1(t)e^{j[\omega t+kr+\alpha_1(t)]} \\ E_\Phi &= A_2(t)e^{j[\omega t+kr+\alpha_2(t)]}. \end{aligned}$$

E_θ and E_Φ are the analytic signal representations of the θ- and Φ-components of the incident electric field, while $\boldsymbol{u}_\theta$ and $\boldsymbol{u}_\Phi$ are unit vectors along θ and Φ coordinates.

The open circuit voltage at the terminals of a receiving antenna due to the incident wave may be derived using the Reciprocity Theorem. The formulation becomes quite elegant when the concept of the complex vector effective height of antenna due to Sinclair [3] is used. Thus, the open circuit voltage V may be written as

$$V(t) = \boldsymbol{E}_i \cdot \boldsymbol{h} = E_\theta h_\theta + E_\Phi h_\Phi \tag{9}$$

where $\boldsymbol{h} = h_\theta \boldsymbol{u}_\theta + h_\Phi \boldsymbol{u}_\Phi$ is the complex vector effective height of the antenna. This quantity is directly related to the transmitting property of the antenna by

$$\boldsymbol{E}_t(r, \theta, \Phi, t) = E_\theta' \boldsymbol{u}_\theta + E_\Phi' \boldsymbol{u}_\Phi = \frac{-j\omega\mu_0 I_i \boldsymbol{h}(\theta, \Phi)}{4\pi r} \cdot e^{j(\omega t - kr)} \tag{10}$$

where

$$E_\theta' = B_1 e^{j(\omega t - kr + \beta_1)}$$

$$E_\Phi' = B_2 e^{j(\omega t - kr + \beta_2)}.$$

$\boldsymbol{E}_t$ is the far-zone field of the antenna at the angular frequency ω when the antenna is used as a transmitting antenna with an input terminal current I_i. B_1, B_2, β_1 and β_2 are time independent constants since the antenna is assumed to transmit a monochromatic wave, and hence $\boldsymbol{E}_t$ is in general elliptically polarized. It can be seen from (10) that the complex vector effective height $\boldsymbol{h}$ is a function only of θ and Φ, and is independent of r and t. We shall assume that $\boldsymbol{h}$ is independent of frequency over the small bandwidth Δf.

For our later use, we shall express the gain $G(\theta, \Phi)$ and the radiation resistance R_a of the antenna in terms of $\boldsymbol{h}(\theta, \Phi)$ as follows:

$$G(\theta, \Phi) = \frac{\boldsymbol{E}_t \cdot \boldsymbol{E}_t^*}{\frac{1}{4\pi}\iint_{4\pi} (\boldsymbol{E}_t \cdot \boldsymbol{E}_t^*) d\Omega} = \frac{\boldsymbol{h} \cdot \boldsymbol{h}^*}{\frac{1}{4\pi}\iint_{4\pi} (\boldsymbol{h} \cdot \boldsymbol{h}^*) d\Omega} \tag{11}$$

and

$$R_a = \frac{2W_t}{|I_i|^2} = \frac{Z_0}{4\lambda^2}\iint_{4\pi} (\boldsymbol{h} \cdot \boldsymbol{h}^*) d\Omega \tag{12}$$

where W_t is the total power transmitted from the antenna, and λ is the wavelength, and $d\Omega = \sin\theta \, d\theta \, d\Phi$.

The power available from the receiving antenna due to the incident wave is

$$W = \frac{\langle V \cdot V^* \rangle}{8R_a}. \tag{13}$$

Substituting (9), one obtains

$$\begin{aligned} W &= \frac{\langle (\boldsymbol{E}_i \cdot \boldsymbol{h})(\boldsymbol{E}_i \cdot \boldsymbol{h})^* \rangle}{8R_a} \\ &= \frac{1}{8R_a} \langle (E_\theta h_\theta + E_\Phi h_\Phi)(E_\theta^* h_\theta^* + E_\Phi^* h_\Phi^*) \rangle \\ &= \frac{1}{8R_a} \{ h_\theta h_\theta^* \langle E_\theta E_\theta^* \rangle + h_\theta h_\Phi^* \langle E_\theta E_\Phi^* \rangle \\ &\quad + h_\theta^* h_\Phi \langle E_\theta^* E_\Phi \rangle + h_\Phi h_\Phi^* \langle E_\Phi E_\Phi^* \rangle \}. \end{aligned} \tag{14}$$

Substituting (11) and (12) into (14), we obtain

$$W = \frac{\lambda^2 G}{4\pi} P \frac{\{ h_\theta h_\theta^* \langle E_\theta E_\theta^* \rangle + h_\theta h_\Phi^* \langle E_\theta E_\Phi^* \rangle + h_\theta^* h_\Phi \langle E_\theta^* E_\Phi \rangle + h_\Phi h_\Phi^* \langle E_\Phi E_\Phi^* \rangle \}}{\{ \langle E_\theta E_\theta^* \rangle + \langle E_\Phi E_\Phi^* \rangle \} (h_\theta h_\theta^* + h_\Phi h_\Phi^*)}. \tag{15}$$

Let us define the coherency matrix for the receiving antenna by

$$A_e[\rho_{ij}'] = \frac{\lambda^2 G}{4\pi} \begin{bmatrix} \rho_{11}' & \rho_{12}' \\ \rho_{21}' & \rho_{22}' \end{bmatrix} \tag{16}$$

where

$$\begin{aligned} \rho_{11}' &= \frac{h_\theta h_\theta^*}{h_\theta h_\theta^* + h_\Phi h_\Phi^*} = \frac{E_\theta' E_\theta'^*}{E_\theta' E_\theta'^* + E_\Phi' E_\Phi'^*} = \frac{B_1^2}{B_1^2 + B_2^2} \\ \rho_{12}' &= \frac{h_\theta h_\Phi^*}{h_\theta h_\theta^* + h_\Phi h_\Phi^*} = \frac{E_\theta' E_\Phi'^*}{E_\theta' E_\theta'^* + E_\Phi' E_\Phi'^*} \\ &= \frac{B_1 B_1 e^{j(\beta_1 - \beta_2)}}{B_1^2 + B_2^2} \\ \rho_{21}' &= \frac{h_\theta^* h_\Phi}{h_\theta h_\theta^* + h_\Phi h_\Phi^*} = \frac{E_\theta'^* E_\Phi'}{E_\theta' E_\theta'^* + E_\Phi' E_\Phi'^*} \\ &= \frac{B_1 B_2 e^{-j(\beta_1 - \beta_2)}}{B_1^2 + B_2^2} \\ \rho_{22}' &= \frac{h_\Phi h_\Phi^*}{h_\theta h_\theta^* + h_\Phi h_\Phi^*} = \frac{E_\Phi' E_\Phi'^*}{E_\theta' E_\theta'^* + E_\Phi' E_\Phi'^*} \\ &= \frac{B_2^2}{B_1^2 + B_2^2}. \end{aligned} \tag{16a}$$

Time averages are not necessary for ρ_{ij}' since B_1, B_2, β_1 and β_2 are time independent constants. In (16), A_e represents the antenna aperture as defined by the IRE Standards, *i.e.*, $A_e = \lambda^2 G / 4\pi$. Hence, $A_e[\rho_{ij}']$ may be called the antenna aperture matrix. The matrix is uniquely determined by the transmitting properties of the antenna.

Similarly, we define the coherency matrix for the incident wave by

$$P[\rho_{ij}] = P \begin{bmatrix} \rho_{11} & \rho_{12} \\ \rho_{21} & \rho_{22} \end{bmatrix} \tag{17}$$

where

$$\left.\begin{aligned}
\rho_{11} &= \frac{\langle E_\theta E_\theta^*\rangle}{\langle E_\theta E_\theta^*\rangle + \langle E_\Phi E_\Phi^*\rangle} = \frac{\langle A_1^2\rangle}{\langle A_1^2\rangle + \langle A_2^2\rangle}\\
\rho_{12} &= \frac{\langle E_\theta^* E_\Phi\rangle}{\langle E_\theta E_\theta^*\rangle + \langle E_\Phi E_\Phi^*\rangle} = \frac{\langle A_1 A_2 e^{-j(\alpha_1-\alpha_2)}\rangle}{\langle A_1^2\rangle + \langle A_2^2\rangle}\\
\rho_{21} &= \frac{\langle E_\theta E_\Phi^*\rangle}{\langle E_\theta E_\theta^*\rangle + \langle E_\Phi E_\Phi^*\rangle} = \frac{\langle A_1 A_2 e^{j(\alpha_1-\alpha_2)}\rangle}{\langle A_1^2\rangle + \langle A_2^2\rangle}\\
\rho_{22} &= \frac{\langle E_\Phi E_\Phi^*\rangle}{\langle E_\theta E_\theta^*\rangle + \langle E_\Phi E_\Phi^*\rangle} = \frac{\langle A_2^2\rangle}{\langle A_1^2\rangle + \langle A_2^2\rangle}
\end{aligned}\right\} \cdot \quad (17a)$$

The difference in the definitions of ρ_{12} and ρ_{12}' (also ρ_{21} and ρ_{21}') should be noted. This is related to the fact that $\boldsymbol{E}_i$ and $\boldsymbol{E}_t$ are propagating in opposite directions.

Then (15) may be written in a concise form using (16), (16a), (17) and (17a).

$$\begin{aligned}
W &= A_e P(\rho_{11}\rho_{11}' + \rho_{12}\rho_{21}' + \rho_{21}\rho_{12}' + \rho_{22}\rho_{22}')\\
&= \text{Trace}\,\{A_e[\rho_{ij}'] \times P[\rho_{ij}]\}. \qquad (18)
\end{aligned}$$

Eq. (18) shows that the available power from the receiving antenna may be formulated in a concise matrix representation.

We shall now show that (18) is formally identical with the basic interference law of partially coherent, quasi-monochromatic wave fields. We shall first express the elements of the antenna coherency matrix in a new form by introducing a new parameter defined by

$$\tan\gamma' = \sqrt{\rho_{22}'/\rho_{11}'} \quad \text{with} \quad 0 \le \gamma' \le \pi/2. \qquad (19)$$

Using (19) and (7a), we have

$$\rho_{11}' = \cos^2\gamma' \quad \text{and} \quad \rho_{22}' = \sin^2\gamma'. \qquad (19a)$$

Since $\boldsymbol{E}_t$ represents a completely polarized wave, we have $\rho_{11}'\rho_{22}' = \rho_{12}'\rho_{21}'$ from which one obtains

$$\begin{aligned}
\rho_{12}' &= \sqrt{\rho_{11}'}\sqrt{\rho_{22}'}\,e^{j(\beta_1-\beta_2)} = \sin\gamma'\cos\gamma' e^{j\xi'}\\
\rho_{21}' &= \rho_{12}^{*\prime} = \sin\gamma'\cos\gamma' e^{-j\xi'} \qquad (19b)
\end{aligned}$$

where $\xi' = \beta_1 - \beta_2$. From (19a) and (19b), one observes that the elements ρ_{ij}' may be determined completely in terms of two angles, γ' and ξ' for a completely polarized wave. Substituting (19a) and (19b) into (18), one obtains

$$\begin{aligned}
W = A_e P\{&\rho_{11}\cos^2\gamma' + \rho_{22}\sin^2\gamma'\\
&+ 2\sqrt{\rho_{11}}\sqrt{\rho_{22}}\sin\gamma'\cos\gamma'\,|\mu_{12}|\,\cos(\xi-\xi')\} \qquad (20)
\end{aligned}$$

where

$$\mu_{12} = \frac{\rho_{12}}{\sqrt{\rho_{11}}\sqrt{\rho_{22}}} = |\mu_{12}|\,e^{j\xi}. \qquad (20a)$$

Eq. (20) may be compared with the basic interference law of partial coherent fields. In the theory of partially coherent light the intensity I due to the interference of two quasi-monochromatic beams of intensities I_1 and I_2 when a phase difference ξ' is introduced between the interfering beams is given by [13]

$$I = I_1 + I_2 + 2\sqrt{I_1}\sqrt{I_2}\,|\mu|\,\cos(\xi-\xi') \qquad (21)$$

where μ is the complex degree of coherence for the two partially coherent beams and ξ is the phase of μ. It is easily seen that (20) is formally identical with (21) if we set $I_1 = A_e P\rho_{11}\cos^2\gamma'$ and $I_2 = A_e P\rho_{22}\sin^2\gamma'$.

The Maximum Available Power

In this section we shall examine how the power available from a receiving antenna will vary as the polarization property of the antenna is changed, the gain of the antenna being held constant. This problem is of interest since one often wishes to design an antenna to obtain the maximum power from an incident wave.

It is well known [13], [14] that a partially polarized wave may be considered as a mixture of a randomly polarized wave and a completely polarized wave which is independent of the former. Moreover, this resolution into two components is unique. In terms of the coherency matrix of the incident wave, this means that the matrix may be uniquely expressed as the sum of two matrices, each representing the randomly and completely polarized components, respectively. The coherency matrix for the incident wave may be written in the following form

$$P\begin{bmatrix}\rho_{11} & \rho_{12}\\ \rho_{21} & \rho_{22}\end{bmatrix} = \tfrac{1}{2}(1-p)P\begin{bmatrix}1 & 0\\ 0 & 1\end{bmatrix} + pP\begin{bmatrix}q_{11} & q_{12}\\ q_{21} & q_{22}\end{bmatrix} \qquad (22)$$

where

$$p = \sqrt{1 - 4(\rho_{11}\rho_{22} - \rho_{12}\rho_{21})} = \text{degree of polarization}$$

$$\left.\begin{aligned}
q_{11} &= \frac{1}{p}\left[\rho_{11} - \tfrac{1}{2}(1-p)\right], & q_{22} &= \frac{1}{p}\left[\rho_{22} - \tfrac{1}{2}(1-p)\right]\\
q_{12} &= \frac{1}{p}\rho_{12}, & q_{21} &= \frac{1}{p}\rho_{21}.
\end{aligned}\right\} \quad (22a)$$

The first term in (22) represents the randomly polarized part, while the second term represents the completely polarized part. The constant p is called the degree of polarization which is the ratio of the power density of the completely polarized part to that of the total power density of the incident wave. Thus, $0 \le p \le 1$. p is independent of the particular choice of the coordinate axes.

Since the matrix $[q_{ij}]$ in (22) represents a completely polarized wave, we have $q_{11}q_{22} = q_{12}q_{21}$. Thus, in accordance with (19), (19a) and (19b), we may write it in the following form

$$\begin{bmatrix}q_{11} & q_{12}\\ q_{21} & q_{22}\end{bmatrix} = \begin{bmatrix}\cos^2\gamma & \sin\gamma\cos\gamma e^{j\xi}\\ \sin\gamma\cos\gamma e^{-j\xi} & \sin^2\gamma\end{bmatrix} \qquad (23)$$

where

$$\tan\gamma = \sqrt{\frac{q_{22}}{q_{11}}}, \qquad \text{with } 0 \le \gamma \le \frac{\pi}{2}$$

and ξ is the phase angle of ρ_{12}.

The power available from the antenna may be written as follows, when (22) is substituted in (18),

$$W = \text{Trace}\{A_e[\rho_{ij}'] \times P[\rho_{ij}]\}$$
$$= \tfrac{1}{2}(1-p)PA_e + pPA_e(q_{11}\rho_{11}' + q_{12}\rho_{21}' + q_{21}\rho_{12}' + q_{22}\rho_{22}'). \quad (24)$$

Substituting (19a), (19b) and (23) into (24), one obtains

$$W = \tfrac{1}{2}(1-p)PA_e + pPA_e\cos^2\left(\frac{\delta}{2}\right) \quad (25)$$

and

$$\cos^2\left(\frac{\delta}{2}\right) = \cos^2\gamma\cos^2\gamma' + \sin^2\gamma\sin^2\gamma' + 2\sin\gamma\cos\gamma\sin\gamma'\cos\gamma'\cos(\xi-\xi') \quad (26)$$

with $0 \leq \delta \leq \pi$. Eq. (25) shows that the power available from the antenna consists of the sum of the contributions from the randomly polarized component and the completely polarized part. The first term in (25), which is due to the randomly polarized part, is constant for an antenna of a given gain and is independent of the polarization of the antenna. Only the second term in (25), which is due to the completely polarized part, is dependent on the antenna polarization. This term varies between 0 to pPA_e as the angle δ varies between π to 0. The significance of δ may be seen by rewriting (26) as

$$\cos\delta = \cos 2\gamma\cos 2\gamma' + \sin 2\gamma\sin 2\gamma'\cos(\xi-\xi'). \quad (27)$$

Eq. (27) is a classical formula of spherical trigonometry. It shows that the angle δ is the angular distance between two points Q and Q' on a sphere of unit radius, where Q represents the state of polarization of the completely polarized part of the incident wave and Q' that of the receiving antenna. The location of Q on the sphere is determined by two angles $(2\gamma, \xi)$ which are specified by the elements of the incident wave matrix $[q_{ij}]$. Similarly, Q' is determined by two angles $(2\gamma', \xi')$ which are specified by the elements of the antenna matrix $[\rho_{ij}']$. The geometrical representations of Q and Q' are illustrated in Fig. 1. The representation of the polarization of a completely polarized wave as a point on the sphere was originated by Poincaré [21] and the sphere is often called the Poincaré sphere. Thus, the coherency matrix for a completely polarized wave may be uniquely represented by a point on the Poincaré sphere. The Poincaré sphere has been used by Deschamps [7] in solving antenna problems.

From (25), it is seen that the available power from the antenna becomes maximum when δ is zero (*i.e.*, the points Q and Q' on the Poincaré sphere coincide). This can occur when and only when $\rho_{11}'=q_{11}$, $\rho_{12}'=q_{12}$, $\rho_{21}'=q_{21}$ and $\rho_{22}'=q_{22}$. The maximum available power then becomes

$$W_m = \tfrac{1}{2}(1-p)PA_e + p\,PA_e = \tfrac{1}{2}(1+p)PA_e. \quad (28)$$

When the incident wave is completely polarized, $p=1$ and (28) becomes

$$W_m = PA_e \quad (29)$$

This is a familiar relation in the monochromatic theory of radio antennas.

In Fig. 2, the available power (in units of PA_e) is plotted as a function of p and δ using (25). This figure illustrates several interesting points. For a given degree of polarization p, there is a maximum and a minimum in the available power from the antenna as the polarization of the antenna is changed. For any p, a maximum value is obtained when $\delta=0$ and a minimum when $\delta=180°$. The range between the maximum and the minimum values of the available power increases as the

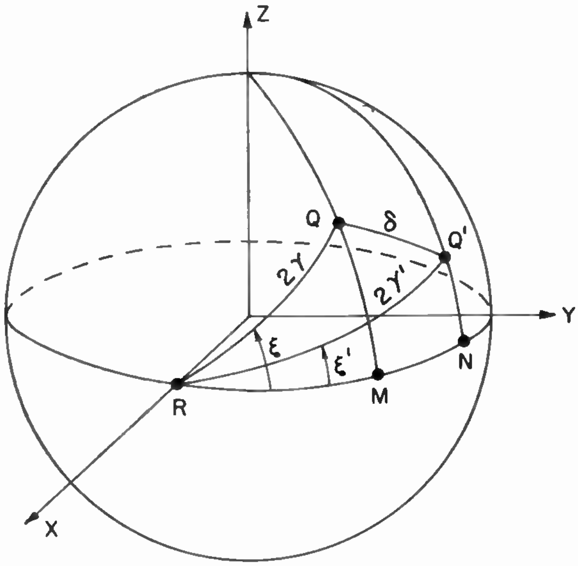

Fig. 1—Geometrical representation of the coherency matrix on the Poincaré sphere. Angular distance $QR=2\gamma$, angular distance $Q'R=2\gamma'$, angle $QRM=\xi$, angle $Q'RN=\xi'$, and angular distance $QQ'=\delta=$ polarization mismatch angle.

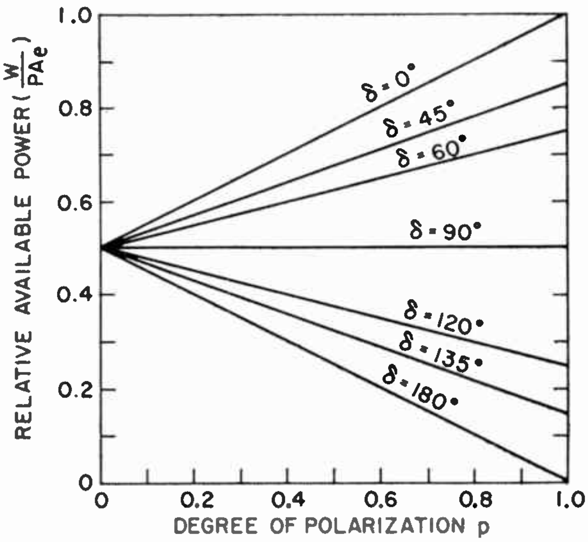

Fig. 2—The relative available power from antenna as a function of the degree of polarization and of the polarization mismatch angle.

degree of polarization of the incident wave is increased. Thus, when $p=1$, which corresponds to a completely polarized incident wave, the available power ranges between the maximum value PA_e to the minimum value 0. When $p=0$, which corresponds to a randomly polarized incident wave, the available power becomes independent of the polarization property of the receiving antenna and is equal to $\frac{1}{2}PA_e$. It is interesting to note that, when $\delta=90°$, the available power from the antenna becomes independent of the degree of polarization and is equal to $\frac{1}{2}PA_e$.

Discussion

In the above discussion, we have seen that the coherency matrix representation provides a concise theoretical tool for calculating the response of a radio antenna to a quasi-monochromatic, partially polarized wave. The formulation is equally valid for a rigorously monochromatic wave. In (18), $A_e[\rho_{ij}']$ may be considered as an operator which depends only on the characteristics of the antenna and independent of the incident field. The operator is determined by the transmitting property of the antenna (*i.e.*, $\boldsymbol{E}_t$) only, as shown in (16). On the other hand, the state of the incident field is described completely by the coherency matrix $P[\rho_{ij}]$. The interaction of the antenna with the incident wave is then given by the concise matrix expression shown in (18).

It has been pointed out by Tai [22] that the present IRE definition of the antenna effective aperture does not take into account the polarization matching effect between the antenna and the incident wave. Eq. (18) suggests that $A_e[\rho_{ij}']$ may be used as a definition of the antenna effective aperture, as pointed out in a previous communication [23] by the author. The proposed definition has an added advantage in that A_e corresponds to the current IRE definition, while $[\rho_{ij}']$ serves to define the polarization of the antenna.

It is interesting to note that the coherency matrix formalism as shown in (18) shows close resemblance with those used in the modern theory of optics and the quantum mechanical treatment of photons and elementary particles. Thus, in the modern theory of optics, the response of a light detector to a light beam incident upon it is given by [24], [25]

$$\mathrm{Trace}\ (QI) \tag{30}$$

where Q and I are the coherency matrices of the optical detector and the incident light beam. Similar forms may be developed for various optical devices such as compensator, absorber, rotator or polarizer, as demonstrated by Parrent and Roman [19].

In quantum statistical mechanics, the ensemble average of the expectation value of a physical observable F is given by

$$\overline{\overline{F}} = \mathrm{Trace}\ (\underline{F}\rho) \tag{31}$$

where $\overline{\overline{F}}$ is the ensemble average of the expectation value of an observable F, $\underline{F}$ is a corresponding matrix operator, and ρ is the density matrix of the quantum mechanical system. For more details on these applications, the reader should refer to excellent papers by Fano [24] and McMaster [25], where further literature is also quoted.

During recent years, the domain of microwave engineering has been extended to the near optical wavelengths due to the development of maser, laser, and submillimeter techniques. Thus the formulation of the antenna theory consistent with that of optics and the quantum mechanical treatment of the polarization of photons will lead to a better understanding of the antenna itself as well as its relation to other physical devices.

The analysis presented in this paper may also be formulated in terms of the Stokes parameters as shown in previous communications by the author [9], [10].

The experimental methods for determining the state of polarization of partially polarized radio waves are not discussed in the present paper. Various polarization measuring schemes have been developed by radio astronomers [26]–[28]. An excellent summary of measurement techniques has been given by Cohen [29] who discussed various methods in terms of Stokes parameters. It should be noted that the coherence theory of partial polarization suggests new ways for the measurement of polarization using correlation techniques. To measure the polarization, it is sufficient to determine four elements of the coherency matrix which are auto- and cross-correlation coefficients of the two mutually perpendicular components of the electric field vector. Polarization measurement techniques in terms of coherence theory will be discussed in a later paper.

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Coherent FDM/FM Telephone Communication*

JEAN A. DEVELET, JR.†, MEMBER, IRE

***Summary*—This paper analyzes telephone communication as applied to a satellite repeater system. Particular emphasis is placed on a method of coherent reception which is important to our emerging communication satellite systems. This reception technique is not new to the field of space communications and telemetry; however, it is new to the field of common carrier telephony. As a consequence, for the class of signals utilized in common carrier telephony, an attempt is made to place on a quantitative footing the design of FDM/FM satellite communication systems. The interrelation among such quantities as sensitivity, bandwidth occupancy, and channel quality is presented for a simply realized second-order receiving system. In addition, the maximum sensitivity achievable with the optimum receiving system is shown. It is anticipated that these two situations will bound the performance of the majority of systems.**

I. INTRODUCTION

TELEPHONE COMMUNICATION by means of active satellite repeaters is currently at hand. Our first spacecraft for this purpose will of necessity be limited in transmitter power output capability. For reliability reasons a considerable time may lapse before available spacecraft transmitter power becomes of little importance to the designer. Until such time, the communication system design must center around achieving the maximum information flow per watt of satellite power even at the expense of another precious commodity, bandwidth.

The most common modulation method utilized in ground microwave relays has the capability of trading bandwidth for transmitter power if this is desired. This modulation method consists of frequency modulating a carrier wave with a single-sideband frequency-division-multiplex (FDM) of telephone channels. Upon traversing the communication link from ground transmitter to spacecraft and thence to the ground receiver, corrupting noise can be considered as added for the most part in two places—the satellite receiver and the ground receiver. Coherent demodulation techniques may then be applied to extract the desired signal from the noise with less required received signal power than standard FM discrimination.

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† Space Technology Laboratories, Inc., Los Angeles, Calif.

This paper will be concerned only with the above mentioned sources of *thermal noise* and no attempt will be made to treat other sources of interference such as intermodulation noise due to link nonlinearities, direct adjacent channel crosstalk, co-channel interference, etc., each of which constitutes an extensive study in its own right. For more information on these topics the reader is referred to [1].

Throughout this paper the standards of the International Radio Consultative Committee (CCIR) are adhered to in order to make the results useful to those familiar with these international guides to radio relay system design. In particular considerable reliance on [2] and [3] was necessary.

II. Telephone Channel Quality

Prior to discussing receiving system sensitivity with coherent demodulation, it will be necessary to develop relationships among the various parameters of an FDM/FM system to determine the expected quality of an individual channel. Consider the satellite system of Fig. 1. Regardless of which FM reception technique is utilized, standard discrimination or coherent demodulation with a frequency-following receiver, the same performance formulations hold above threshold. Thus, we may discuss this topic independently of receiver sensitivity, realizing that the extra sensitivity which may be provided by coherent demodulation will allow us to increase frequency deviations and obtain the same system performance at lower received signal levels.

Consider the voltage input to the system of Fig. 1 to be that resulting from a composite of single-sideband channels extending from F_1 to F_2 cps whose equivalent power-spectral density is shown in Fig. 2. The CCIR has determined that a multichannel FDM signal can be represented during the busy hour by white Gaussian noise extending from F_1 to F_2 cps where specific values of F_1 and F_2 depend on the channel arrangement [4]. The power level of this equivalent signal is given by the CCIR as [4]:

$$
\begin{aligned}
P_{eq} &= -1 + 4 \log_{10} (N_c), && \text{dbm0 } 12 \le N_c < 240 \\
P_{eq} &= -15 + 10 \log_{10} (N_c), && \text{dbm0 } 240 \le N_c
\end{aligned}
\tag{1}
$$

where

N_c number of channels in the system.

Dbm0 in (1) is power in dbm referred to a point of zero relative level in the communication system. The zero relative level concept is convenient in that one may talk of absolute power with no ambiguity.

As a direct result of (1), a system design may now be carried out without reference to the intricate statistics of a number of actual talkers.

Since the signal described by (1) must be passed by amplifiers and filters in the FM system, it is convenient to define a peak factor for the noise signal. In this paper the peak of the noise signal is defined as being 13 db above the rms value. This corresponds to a probability of overload less than 10^{-5} which the author has assumed adequate for satellite communication.

The signal and noise performance of the communication system may be found by considering the output of the receiver demodulator. Note that this point may not

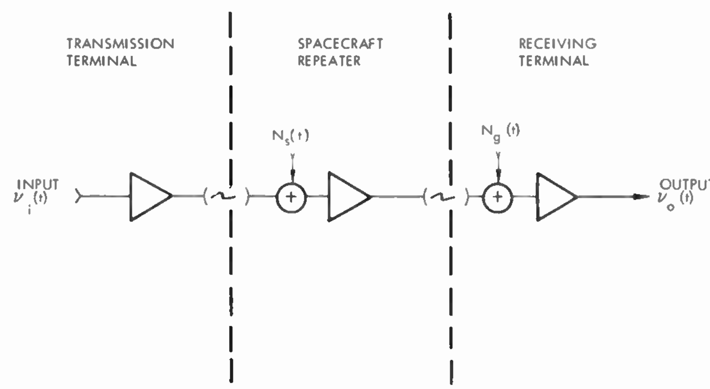

Fig. 1—Active communication satellite system.

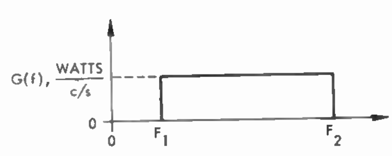

Fig. 2—Equivalent power-spectral density of the multichannel signal.

be at zero relative level. Let the signal power output at the receiver demodulator in Fig. 1 be

$$S_0 = kF^2_{\mathrm{drms}},\ \mathrm{mw} \tag{2}$$

where

S_0 = sinusoidal output power, milliwatts
k = demodulator constant, mw/cps^2
F_{drms} = rms deviation due to an 800-cps test tone of 0 dbm0 without pre-emphasis (CCIR Test Tone for Telephony Systems).

Familiar FM theory [5] gives the *one-sided* noise spectral density in the top channel of the radio system as

$$\Phi_0 = k\left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]F_2^2,\ \mathrm{mw/cps} \tag{3}$$

where

Φ_0 = top channel *one-sided* spectral density mw/cps
k = demodulator constant, mw/cps^2
Φ_s, Φ_g = *one-sided* power spectral density of the additive spacecraft and ground receiver noise perturbations respectively, mw/cps
S_s, S_g = sinusoidal carrier power received at the spacecraft and ground receiver respectively, mw
F_2 = highest channel frequency, cps.

The psophometrically weighted [6] noise power at the demodulator output in the top channel may be found by multiplying (3) by the noise bandwidth of the psophometric weighting filter. Ref. [3] gives this noise bandwidth as

$$B = 3100 \times 10^{-0.25},\ \mathrm{cps}. \tag{4}$$

Division of (2) by the product of (3) and (4) yields the test tone to noise power ratio in the top channel without pre-emphasis as follows:

$$\frac{S_0}{N_0} = \frac{10^{0.25}F^2_{\mathrm{drms}}}{3100\left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]F_2^2} \tag{5}$$

where

S_0/N_0 = test tone to noise power ratio in the top channel of the satellite communication system.

Considering F_{drms} is caused by a 0 dbm0 test tone one may express the psophometrically weighted top channel noise power in picowatts [7] referred to zero relative level. It is easily shown

$$N_{\mathrm{pw}} = \frac{3.1\left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]F_2^2 \times 10^{12}}{10^{0.25}F^2_{\mathrm{drms}}},\ \mathrm{pw\ (psoph)}. \tag{6}$$

Since (3) demonstrates that top channel performance is worse than the remaining channels, pre-emphasis is usually applied to equalize the noise in all the channels. If just thermal noise were present, 6 db/octave pre-emphasis would achieve equal noise in all channels. However, certain intermodulation products due to link nonlinearities, terminal noise, etc., prevent the use of this ideal pre-emphasis. Through study and practical experience with radio relay systems CCIR has recommended a pre-emphasis curve [8]. This curve yields approximately 4-db improvement in top channel tone to noise power ratio instead of 4.8 db which is obtainable by use of 6 db/octave pre-emphasis. Recommended CCIR pre-emphasis and 6 db/octave pre-emphasis result in so nearly the same radio relay system performance that in Section IV of this paper the latter will be used because of its analytic simplicity.

Finally, (6) may be written

$$N_{\mathrm{pw}} = \frac{3.1\left\{\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right\}F_2^2 \times 10^{12}}{10^{0.25}IF^2_{\mathrm{drms}}},\ \mathrm{pw\ (psoph)} \tag{7}$$

where

I = numerical improvement achievable by use of pre-emphasis, *e.g.*, $I=3$ for 6 db/octave and $F_2 \gg F_1$
F_{drms} = test tone deviation without pre-emphasis, cps.

Eq. (7) is the principal result of this section. It establishes the psophometrically weighted noise in any telephone channel (assuming pre-emphasis equalizes this quantity) vs the rms deviation of a 0 dbm0 800-cps test tone, the spacecraft and ground receiver noise perturbations, the highest equivalent baseband frequency, and the received carrier powers at the spacecraft and ground receiver.

The value for N_{pw} in satellite communications has not yet been specified by the CCIR. For purposes of the performance curves in this paper, however, a nominal value of 10,000 pw (psoph) is chosen. This is consistent with present CCIR *total* noise objectives in a 2500-km link [8]. Enough latitude either side of nominal is displayed on the curves of Figs. 5–12 in Section V to provide for most eventualities.

III. Radio Frequency Bandwidth Occupancy

The purpose of this section is to present formulations for estimating RF bandwidths necessary for communication with FDM/FM. Since the actual bandwidth utilized is a matter of considerable engineering judgment and depends on the number of repeaters in tandem, only a simplified bandwidth "occupancy" will be considered, given by the usual rule of thumb

$$B_{\mathrm{rf}} = 2F_2 + F_{\mathrm{pp}},\ \mathrm{cps} \tag{8}$$

where

B_{rf} = bandwidth "occupancy" of the signal, cps
F_2 = highest equivalent baseband frequency, cps
F_{pp} = peak-peak deviation of the multichannel signal, cps.

For more detailed treatment of bandwidths required for FDM/FM signals the reader is directed to the work of Medhurst.[9]

Referring back to the equivalent multichannel loading P_{eq}, equation (1), and the 13-db peak factor, one may calculate F_{pp} in terms of F_{drms}, the deviation of the 0 dbm0 800-cps test tone without pre-emphasis. A simple analysis yields

$$F_{pp} = 2(10^{1.3}P_{eq})^{1/2}F_{drms}, \text{ cps.} \tag{9}$$

Eq. (9) is not affected by pre-emphasis, since a rule of pre-emphasis is to keep the rms value of the total frequency deviation the same with and without pre-emphasis.

Substitution of (9) in (8) yields the principal result of this section.

$$B_{rf} = 2\{F_2 + 10^{0.65}P_{eq}{}^{0.5}F_{drms}\}, \text{ cps.} \tag{10}$$

IV. Coherent Receiver Sensitivity

A coherent receiver is one in which a "replica" of the received signal is generated locally to assist in more optimal demodulation of that signal.

There are many realizations of ths type of reception. One realization is sometimes referred to as "FM feedback" reception. This technique was invented by J. G. Chaffee of Bell Telephone Laboratories [10]. Another realization is a modulation tracking phase-lock receiver. This reception technique is a variation of the type used to obtain horizontal synchronization in television receivers. Either realization of the receiver, and there are others such as exalted carrier techniques [11], possess certain dynamic properties when driven by an FM or PM signal.

In addition, either realization possesses similar *noise-induced* threshold properties. That is, coherent receivers cease to function when a "clean" local replica can no longer be generated. Dr. L. H. Enloe and C. L. Ruthroff of Bell Telephone Laboratories [12], [13], found that the FM feedback realization had a threshold close to the point where the rms value of the phase error due to *noise* alone between the local replica andthe received signal became greater than 0.32 radian. C. R. Woods and E. M. Robinson of General Electric found a similar threshold point for the phase-lock receiver [14]. Their data indicates 0.354 radian maximum rms noise error is tolerable prior to onset of loss of lock.

The effect of *modulation* error manifests itself in a different manner in phase-lock reception than in "FM feedback" reception. In phase-lock reception of phase-encoded FDM/FM signals in the presence of additive white Gaussian noise, the sum total of the mean-square modulation plus noise error must remain small to prevent nonlinear operation of the loop-phase detector. It will be assumed for purposes of this paper that phase-lock loop threshold occurs when the sum total mean-square error due to modulation *and* noise equals 1/8 rad². In "FM feedback" reception utilizing simple low-order transfer functions, Dr. L. H. Enloe has indicated lower thresholds are possible than the phase-lock receiver due to the fact larger *modulation* phase errors are tolerable prior to development of loop nonlinearities.

In the case of FDM/FM signals which can be represented by white Gaussian noise as in Fig. 2, Yovits and Jackson [15] have shown that high-order, actually infinite-order, transfer functions are necessary for optimal demodulators. The optimization performed was that of finding the transfer function which yields minimum total mean-square error. For phase-lock reception this minimization also yields maximum receiver sensitivity.

Study of the optimal transfer function derived by Yovits and Jackson shows that for the situation of high channel quality (large frequency deviations due to the signal), the modulation error becomes insignificant compared to the noise error.[1] Thus for the special case of FDM/FM signals and use of the corresponding optimal phase-lock transfer function at toll level qualities, threshold is determined primarily by noise error only. As a consequence of the similar *noise-induced* threshold property, use of the Yovits and Jackson filter in either "FM feedback" *or* phase-lock reception will yield similar lower bounds on receiver sensitivity.

An "FM feedback" second-order receiver can be designed to be more sensitive than a second-order phase-lock receiver at high deviations due to the larger allowable modulation error; however, by using more complex transfer functions to approach maximum sensitivity, one will find less and less difference between the two techniques until finally they both converge to similar performance with the Yovits and Jackson filter.

This section will now treat two situations which will bound the performance of most coherent FDM/FM receivers designed for maximum sensitivity be they of the phase-lock configuration or the "FM feedback" con-

[1] Proof of this fact can most easily be accomplished by forming the ratio of the modulation error in (18) to the total error (13) and taking the limit as $\Phi_m/\Phi_n \to \infty$. Substitution of the Yovits and Jackson optimal filter function in (18) yields the following modulation error:

$$\sigma_m{}^2 = \frac{\Phi_m\Phi_n}{\Phi_m + \Phi_n}(F_2 - F_1).$$

The ratio of the above modulation error to total error given by (13) thus becomes $R = [1+\Phi_n/\Phi_m) \log_e (1+\Phi_m/\Phi_n)]^{-1}$. It is clear $R \to 0$ as $\Phi_m/\Phi_n \to \infty$.

figuration. First, the sensitivity of a receiver utilizing the optimal transfer function as derived by Yovits and Jackson will be treated followed by derivation of the sensitivity of a simple second-order transfer function utilized in a phase-lock configuration.

The latter case is most important in practice. The reason for the importance of the second-order loop is its simplicity. At the wide base-bandwidths necessary for 300, 600 and 1800 channels of telephony, it is difficult to realize much more than a second-order loop because of stability considerations.

A. The Optimal Receiver

Consider the filter (or receiver) of Fig. 3 as postulated by Yovits and Jackson [16]. The input to the receiver consists of the white Gaussian phase variable $\phi_m(t)$ proportional to the telephone multiplex signal input to the communication link.[2] $\phi_n(t)$ is the corrupting phase noise spectral density due to the ground receiver *and* spacecraft. Deleting proportionately constants which

Yovits and Jackson [17] have shown that if one uses the optimum closed-loop transfer function in Fig. 3, limited only by physical realizability, the minimum mean-square error $\overline{\epsilon(t)^2}$ for a white signal and noise spectral density is given by [18]

$$\overline{\epsilon(t)^2}_{\min} = \Phi_n(F_2 - F_1)\log_e\left\{1 + \frac{\Phi_m}{\Phi_n}\right\}. \tag{13}$$

Substitution of (7), (11), and (12), in (13) yields the fundamental relation

$$\epsilon^2_{\min} = \left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right](F_2 - F_1) \cdot \log_e\left\{1 + \frac{3.1P_{eq} \times 10^{12}}{N_{pw}\cdot 10^{0.25}(F_2 - F_1)}\right\}. \tag{14}$$

Eq. (14) is fundamental in that given a channel quality N_{pw} and a maximum $\epsilon_{\min}^2$ (1/8 rad²) the maximum

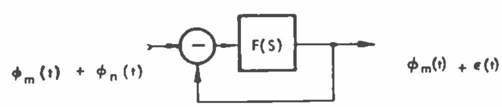

Fig. 3—Yovits and Jackson filter.

are identical for both signal and noise at any point in the system, it can be shown

$$\Phi_n = \left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right], \ \text{rad}^2/\text{cps} \tag{11}$$

where

Φ_n = *one-sided* phase noise spectral density due to spacecraft and ground receiver, rad²/cps.

By noting F_{drms} is the rms frequency deviation due to a 0 dbm0 800-cps test tone inserted at a point where the deviation with and without pre-emphasis is the same, one may easily derive the ***one-sided phase*** spectral density of the signal Φ_m. The result is

$$\Phi_m = \frac{P_{eq} I F^2_{drms}}{(F_2 - F_1)F_2^2}, \ \text{rad}^2/\text{cps} \qquad F_1 \le F \le F_2 \tag{12}$$

$$0 \ \text{elsewhere}$$

where P_{eq} is given by (1)

$$I = \frac{3}{1 + \frac{F_1}{F_2} + \left(\frac{F_1}{F_2}\right)^2} \ \text{the improvement factor}$$

attainable with 6 db/octave pre-emphasis.

[2] 6 db/octave pre-emphasis is assumed here for analytic simplicity. The resultant *performance* is within 1 db of an actual pre-emphasis schedule, however, the resulting *threshold* is within a fraction of a db.

value of $\Phi_s/S_s + \Phi_g/S_g$ is determined. This is exactly the threshold relation desired.

Defining $\alpha = S_g\Phi_s/S_s\Phi_g$ as the fractional contribution of the ground to spacecraft link to over-all system noise and letting $\epsilon_{\min}^2 = 1/8$, one can obtain from (14) the simple threshold criteria for the optimal loop

$$S_g = 4\{1 + \alpha\}\Phi_g B \tag{15}$$

where

$$B = 2(F_2 - F_1) \cdot \log_e\left\{1 + \frac{3.1P_{eq} \times 10^{12}}{N_{pw}10^{0.25}(F_2 - F_1)}\right\}, \ \text{cps}. \tag{16}$$

Thus the received signal-to-noise power ratio in a bandwidth B, cps must be 6 db or more (depending on ground-spacecraft contribution) for proper demodulation.

Note that quantity B is not a strict ***noise*** bandwidth since it also includes the effects of modulation error. B is essentially a coefficient whose dimensions are cps, which, when multiplied by the ***two-sided*** phase noise spectral density input, yields the ***total*** mean-square loop error due to both noise ***and*** modulation.

In any particular receiver realization, one could compare the necessary received signal power S_g to that determined by (15) and (16). This would give insight as to the efficiency of the design for FDM/FM signals.

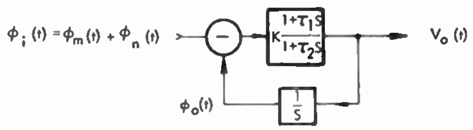

Fig. 4—Second-order receiver.

Section V accomplishes this comparison for the second-order loop.

B. The Second-Order Receiver

This section will utilize the terminology of the phase-lock realization of this receiver since much documentation exists in this area [19]–[22].

Consider the block diagram of Fig. 4. After Gruen [23], Martin [24], and others [25], [26], the loop was linearized and as can be seen is identical to the "filter" of Yovits and Jackson.

The definitions of $\phi_m(t)$ and $\phi_n(t)$ remain the same as those in Section IV-A. The phase detector in the phase-lock loop has been replaced by its linearized equivalent, the subtractor. The voltage controlled oscillator is a perfect integrator of baseband voltage to RF phase. The loop filter is a simply realized lag network. K represents the over-all loop gain including all elements at the signal level of concern. After the work of Gruen [27] the following definitions are established.

$$\omega_n^2 = \frac{K}{\tau_2}, \ (\text{rad/sec})^2$$

$$2\xi\omega_n = \frac{1 + K\tau_1}{\tau_2}, \ \text{rad/sec}$$

where

ω_n = loop undamped natural frequency, rad/sec

ξ = ratio of actual to critical loop damping

τ_1, τ_2 = time constants of the lead lag compensation networks, seconds.

The loop transfer function as determined by Gruen [28] with the above definitions is the following:

$$\frac{\phi_o}{\phi_i} = \frac{\omega_n^2\left\{1 + \left(\frac{2\xi}{\omega_n} - \frac{1}{K}\right)s\right\}}{\omega_n^2 + 2\xi\omega_n s + s^2}. \tag{17}$$

The error function which describes the faithfulness of tracking is easily established from the work of Yovits and Jackson [29] as follows:

$$\overline{\epsilon(t)^2} = \underbrace{\int_0^\infty \left|\frac{\phi_o}{\phi_i}\right|^2 \Phi_n df}_{\textit{Noise error}} + \underbrace{\int_0^\infty \left|1 - \frac{\phi_o}{\phi_i}\right|^2 \Phi_m df}_{\textit{Modulation error}}; \ \text{rad}^2. \tag{18}$$

The first portion of (18) has been evaluated by Gruen [28] and is repeated here as a slight modification of his (36) in order to utilize our *one-sided* spectral densities. In addition, the following assumes small ω_n/K which is normally the case in practice. The first portion of (18) now becomes

$$\overline{\epsilon_1^2} = \Phi_n\omega_n\left(\frac{1 + 4\xi^2}{8\xi}\right), \ \text{radians}^2. \tag{19}$$

For CCIR channel arrangements and the channel qualities normally encountered, the second half of (18) can be approximated as follows to an excellent degree for $\xi \cong 1/\sqrt{2}$ and $\omega_n/K \to 0$:[3]

$$\overline{\epsilon_2^2} \cong \frac{(2\pi)^4\Phi_m}{\omega_n^4}\int_{F_1}^{F_2} f^4 df$$

or

$$\overline{\epsilon_2^2} \cong \frac{\Phi_m(2\pi)^4}{5\omega_n^4}\left[F_2^5 - F_1^5\right]. \tag{20}$$

Combining (20) and (19) yields the total loop error.

$$\overline{\epsilon^2} = \Phi_n\omega_n\left\{\frac{1 + 4\xi^2}{8\xi}\right\} + \frac{(2\pi)^4\Phi_m(F_2^5 - F_1^5)}{5\omega_n^4}, \ \text{rad}^2. \tag{21}$$

Substitution of Φ_n (11) and Φ_m (12) in (21) finally gives

$$\overline{\epsilon^2} = \left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]\omega_n\left(\frac{1 + 4\xi^2}{8\xi}\right) + \frac{(2\pi)^4 P_{\text{eq}} I F^2_{\text{d rms}}(F_2^5 - F_1^5)}{5(F_2 - F_1)F_2^2\omega_n^4}, \ \text{rad}^2. \tag{22}$$

Utilizing (7) we obtain

$$\overline{\epsilon^2} = \left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]\left\{\omega_n\left(\frac{1 + 4\xi^2}{8\xi}\right) + \frac{(2\pi)^4 P_{\text{eq}}(F_2^5 - F_1^5)3.1 \times 10^{12}}{5N_{\text{pw}}10^{0.25}\omega_n^4(F_2 - F_1)}\right\}, \ \text{rad.}^2 \tag{23}$$

Considering all other parameters fixed except ω_n, the natural frequency of the receiver loop, $\overline{\epsilon^2}$ may be minimized with respect to this quantity. The optimum ω_n becomes for a damping of $\xi = 1/\sqrt{2}$:

$$\omega_{n\ \text{opt}} = 2\left\{\frac{3.1(2\pi)^4 P_{\text{eq}}(F_2^5 - F_1^5) \times 10^{12}}{15\sqrt{2}\ 10^{0.25}N_{\text{pw}}(F_2 - F_1)}\right\}^{1/5}, \ \text{rad/sec}. \tag{24}$$

[3] These are reasonable values for most loop designs. Since Yovits and Jackson have shown the second-order loop is not optimum for FDM/FM signals, the author does not feel justified in using other than normally encountered values for ξ and ω_n/K.

Substitution of (24) back in (23) yields the minimum error attainable contingent on the assumptions. One obtains for $\xi = 1/\sqrt{2}$:

$$\overline{\epsilon^2} = \frac{1}{2}\left[\frac{\Phi_s}{S_s} + \frac{\Phi_g}{S_g}\right]\left[\frac{5}{4} B_N\right], \text{ rad}^2 \qquad (25)$$

where

$$B_N = \frac{3\omega_{n\ \mathrm{opt}}}{2\sqrt{2}} \cdot \text{cps}.$$

B_N is the conventional definition of *two-sided* noise bandwidth in cps for a second-order loop of damping $\xi = 1/\sqrt{2}$ [30]. The factor 5/4 takes into account the effects of modulation error. To be consistent with the results of Section IV-A, a quantity $B = 5/4\ B_N$, cps will be defined as the noise coefficient, which when multiplied by the *two-sided* phase noise spectral density, will yield the *total* mean-square loop error due to the effects of both modulation *and* noise.

For a damping ratio of $\xi = 1/\sqrt{2}$, it can also be shown $B_N = 3.24\ F_{3\mathrm{db}}$, cps where $F_{3\mathrm{db}}$ is the closed loop *baseband* 3-db bandwidth in cps. Similarly, the newly defined noise coefficient, $B = 4.05 F_{3\mathrm{db}}$, cps for the second-order loop with $\xi = 1/\sqrt{2}$.

Applying the identical threshold criteria as in Section IV-A, $\epsilon^2 \leq \frac{1}{8}$ rad^2, (25) becomes

$$S_g = 4(1 + \alpha)\Phi_g B \qquad (26)$$

where

$$B = \frac{5}{4} B_N, \text{ cps}.$$

Thus as in the optimum loop, the received signal-to-noise power ratio in a bandwidth B, cps must be 6 db or more (depending on ground-spacecraft contribution) for proper demodulation.

V. Summary of Results

The previous sections have derived very important relationships for communication system design utilizing FDM/FM and coherent reception based on CCIR practice. Bounds on the performance expected by utilizing coherent reception have been established on the one hand by a second-order transfer function and on the other by the optimal transfer function derived by Yovits and Jackson [31].

Table I summarizes the most important relationships under the *key assumption that the desired quality of performance is achieved at threshold.* In order to make the maximum use of spacecraft transmitter power and thereby allow every decibel of received power above threshold to represent true margin, that is both performance *and* threshold margin, this assumption should be made.

TABLE I
Formula Summary*

Quantity	Type of Demodulator	Formulation at Threshold
0 dbm0 test tone deviation	Optimal and second-order	$F_{\mathrm{drms}} = F_2 \left\{ \frac{3.1}{4 \times 10^{0.25} I N_{\mathrm{pw}} B} \right\}^{1/2} \times 10^6$, cps
Noise coefficient	Optimal	$B = 2(F_2 - F_1) \log_e \left\{ 1 + \frac{3.1\ P_{\mathrm{eq}} \times 10^{12}}{N_{\mathrm{pw}} 10^{0.25}(F_2 - F_1)} \right\}$, cps
	Second-order	$B = \frac{15}{4\sqrt{2}} \left\{ \frac{3.1(2\pi)^4 P_{\mathrm{eq}}(F_2^5 - F_1^5) \times 10^{12}}{15\sqrt{2}\ 10^{0.25} N_{\mathrm{pw}}(F_2 - F_1)} \right\}^{1/5}$, cps
Threshold criteria	Optimal and second-order	$S_g = 4(1 + \alpha)\Phi_g B$
Bandwidth occupancy	Optimal and second-order	$B_{\mathrm{rf}} = 2\{F_2 + 10^{0.65} P_{\mathrm{eq}}^{0.5} F_{\mathrm{drms}}\}$, cps

* Note: These formulas apply for 6 db/octave pre-emphasis only. In addition for the second-order loop, the damping was taken to be $\xi = 1/\sqrt{2}$ and $\omega_n/K \to 0$.

Figs. 5–12 plot the results of the formulations in Table I for the equivalent CCIR loading [4], 6 db/octave pre-emphasis, CCIR channel arrangements, and of course the threshold assumption in which maximal use of spacecraft power is achieved.

One should note the large bandwidth occupancy required when one attempts to achieve full use of transmitter power. Presently these large bandwidths may be necessary; however, future satellite power development will allow the maximum bandwidths indicated in the figures to be reduced by increasing transmitter power and reducing deviations for the same net system performance. Of course in this situation the performance margin will always be less than the threshold margin.

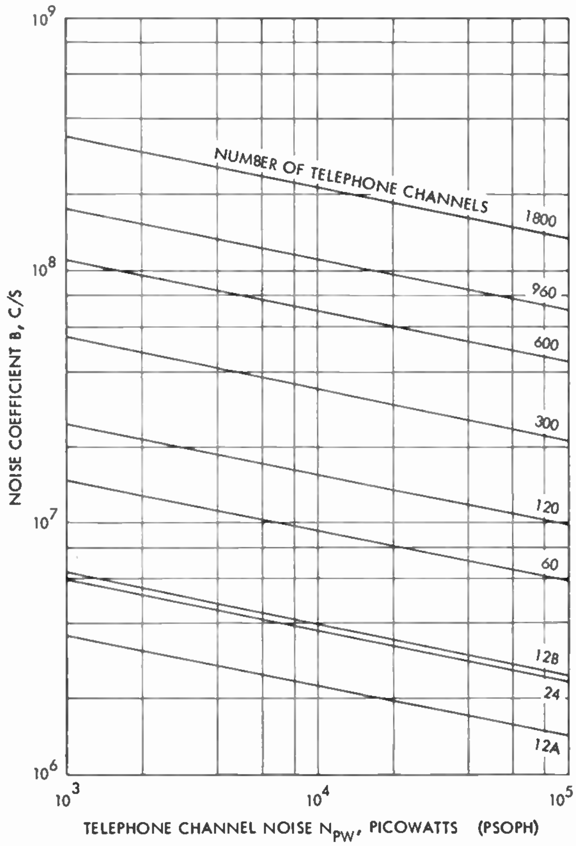

Fig. 5—Noise coefficient, second-order loop.

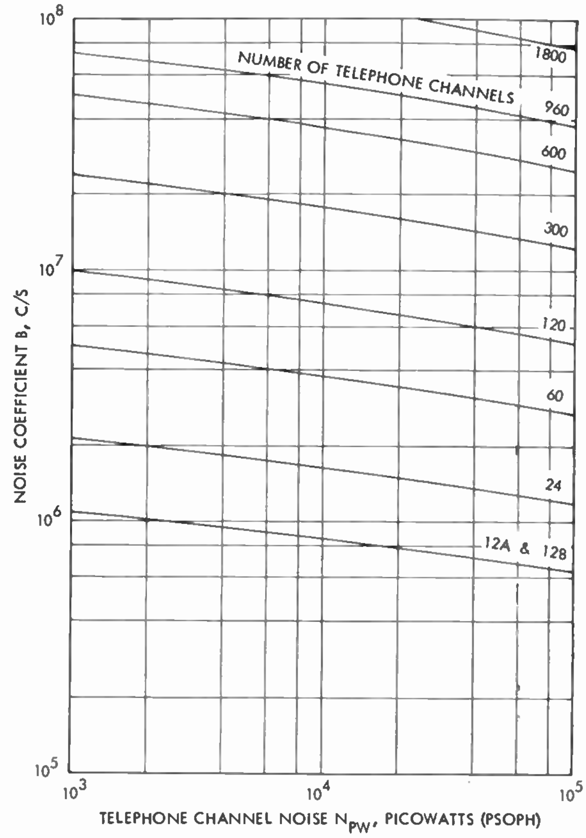

Fig. 6—Noise coefficient, optimal loop.

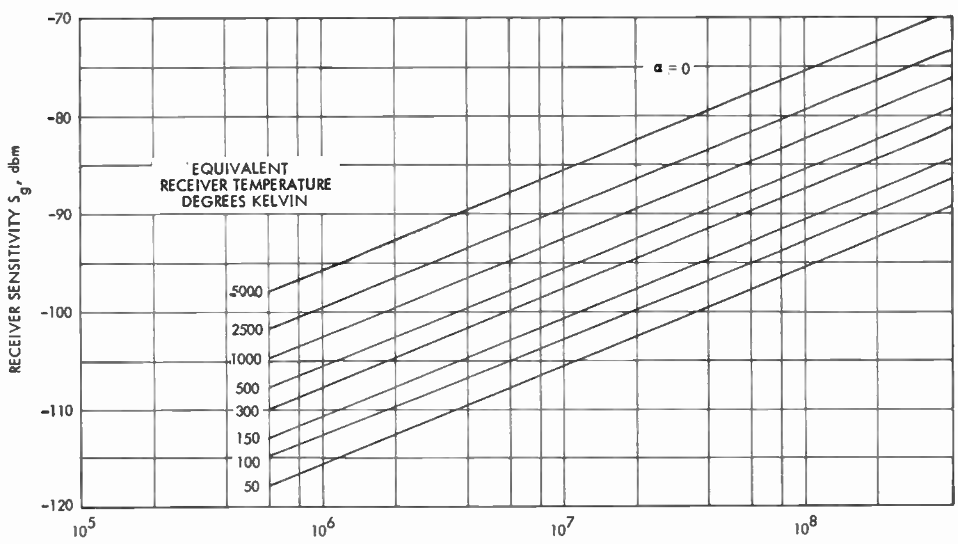

Fig. 7—Receiver sensitivity, dbm.

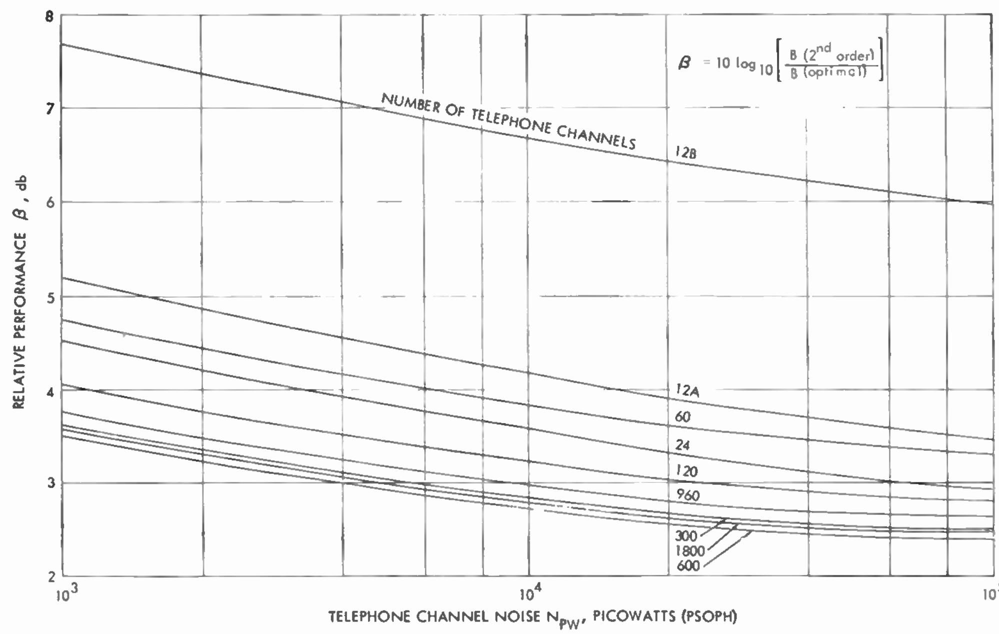

Fig. 8—Sensitivity comparison second-order vs optimal loop.

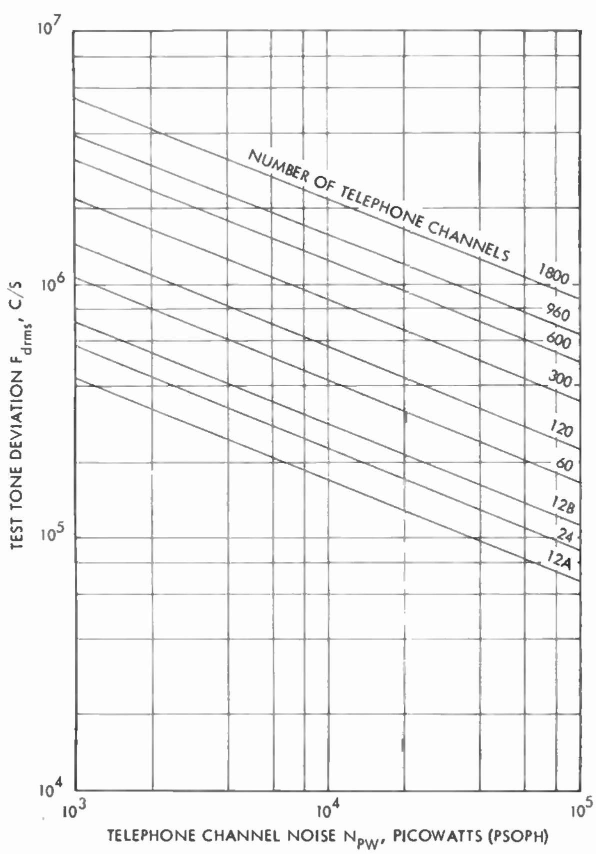

Fig. 9—One milliwatt 800 cps test tone deviation, second-order loop.

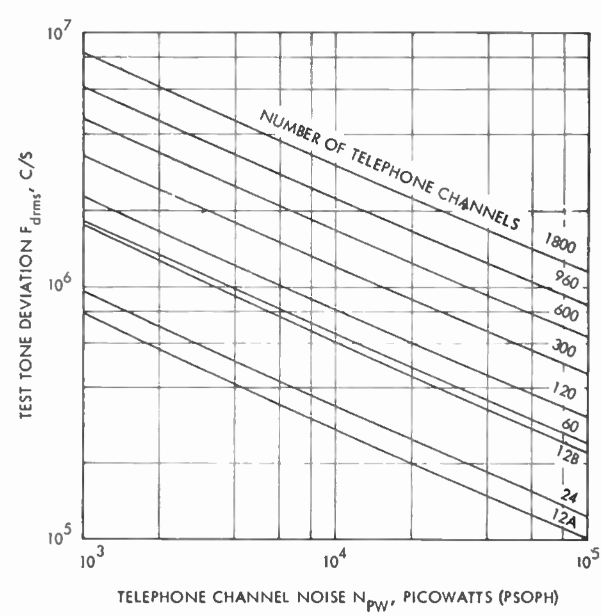

Fig. 10—One milliwatt 800 cps test tone deviation, optimal loop.

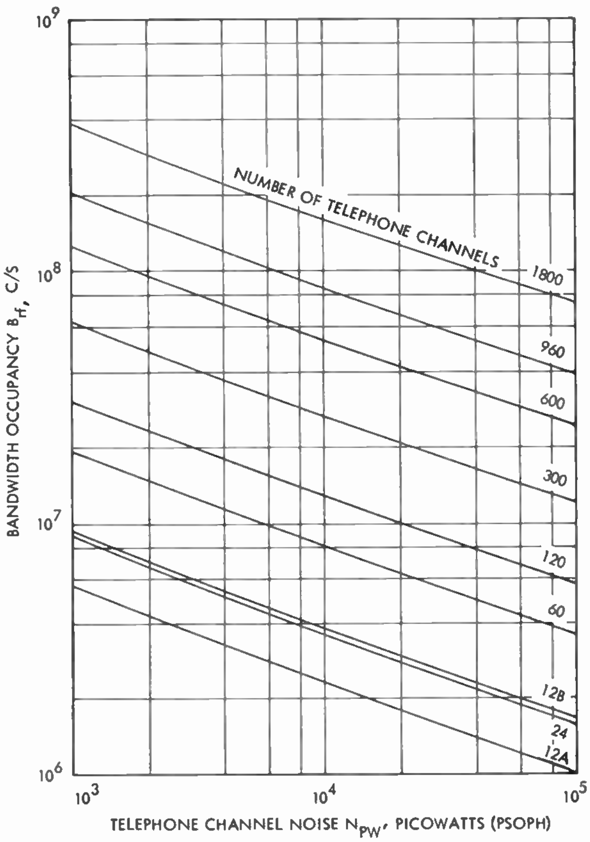

Fig. 11—RF bandwidth occupancy, second-order loop.

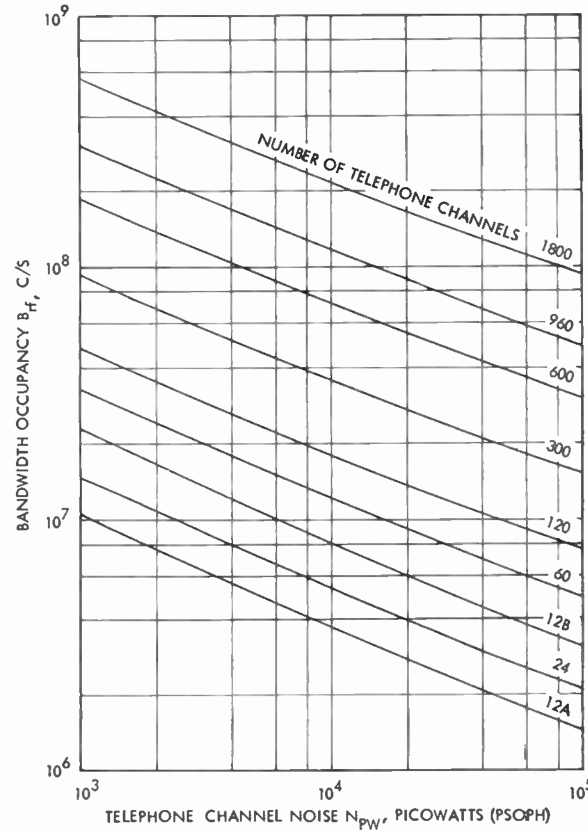

Fig. 12—RF bandwidth occupancy, optimal loop.

Acknowledgment

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A Piezoelectric-Piezomagnetic Gyrator*

M. ONOE†, MEMBER, IRE, AND M. SAWABE‡

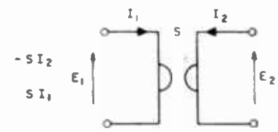
Summary—A linear passive unilateral element, which has a high forward/backward transmission ratio at all frequencies, is realized by combining *two* resistances and an electromechanical gyrator according to Gamo's theory. The gyrator consists of three mechanically coupled ceramic elements, two of which may be piezomagnetic and the third piezoelectric, or two of which may be piezoelectric and the third piezomagnetic. In either arrangement, a unilateral coupling between mechanical and electrical systems is provided by using one piezoelectric and one piezomagnetic element. The third element is merely for driving the mechanical system. Such a design makes manufacturing considerably easier. Characteristics of the gyrator have been discussed based on an equivalent circuit. A constant input resistance can be obtained at a terminal pair of the isolator by purely electrical means even after the fabrication of the gyrator. This simplifies the matching at this terminal. The theoretical minimum insertion loss is 3 db under the matched termination conditions. A few models have been made at 150 kc using both the sandwiched and the cascaded structures. A highly achromatic suppression of backward transmission (45 db) has been obtained in agreement with theoretical predictions. Minimum forward insertion loss consists of the theoretical minimum of 3 db and an excess loss of a few db due principally to mechanical losses.

I. Introduction

A LINEAR, passive electric circuit which violates the reciprocity theorem has been a subject of much study in recent years. Tellegen introduced the ideal gyrator as the fifth fundamental circuit element, besides the existing four elements, namely, resistance, capacitance, inductance and the ideal transformer.[1,2] An ideal gyrator has circuit equations shown in Fig. 1. Any linear, passive, nonreciprocal network can be represented by a suitable combination of ideal gyrators and reciprocal circuits.

Two features of a gyrator are especially important for practical applications. First, a gyrator is an immittance converter as shown in Fig. 2, which transforms any circuit into its dual. Second, a gyrator can be an isolator, which allows the transmission of signal in one direction only, by a suitable combination with a resistance as shown in Fig. 3.

In practice, a gyrator has been realized by using the Hall effect in a semiconductor,[3] the Faraday effect in a ferrite[4] or the nonreciprocal nature in a combination of two dual types of electromechanical transducers. These practical gyrators consist of an ideal gyrator and extra circuits, which usually limit their range of application. This is especially true for an electromechanical gyrator because the band-pass characteristics of the transducers make the gyrator a relatively narrow band device. Hence the most promising application of electromechanical gyrators seems to be in isolators, which will be the subject of the present paper. McMillan first noted that if an electric transducer is mechanically coupled to a magnetic transducer, the resultant system is an antireciprocal four-terminal circuit, which can be an isolator by a suitable combination with a circuit



$$E_1 = -SI_2$$
$$E_2 = SI_1$$

Fig. 1—Ideal gyrator.

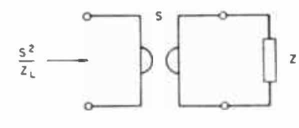
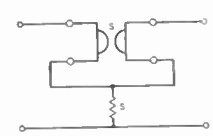
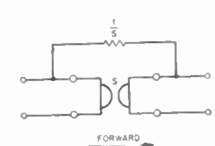


Fig. 2—Immittance converter.



(a)



(b)

Fig. 3—Ideal isolator.

* Received April 24, 1962; revised manuscript received May 23, 1962.

† Bell Telephone Laboratories, Whippany, N. J., on leave of absence from Institute of Industrial Science, University of Tokyo, Azabu, Tokyo, Japan.

‡ Institute of Industrial Science, University of Tokyo, Azabu, Tokyo, Japan.

[1] B. D. H. Tellegen, "The gyrator, a new electric network element," *Philips Res. Repts.*, vol. 3, pp. 81–101; April, 1948.

[2] The number of independent ideal circuit elements is somewhat debatable since the ideal gyrator can be used with an inductance to replace a capacitance or vice versa. Similarly, ideal transformers can be extracted from a cascaded pair of ideal gyrators. Hence, a complete system of independent circuit elements need consist only of resistance, inductance or capacitance, and the ideal gyrator. The authors wish to thank the reviewer for this comment.

[3] W. P. Mason, W. H. Hewitt and R. F. Wick, "Hall effect modulators and gyrators employing magnetic field independent orientations in germanium," *J. Appl. Phys.*, vol. 24, pp. 166–175; February, 1953.

[4] C. L. Hogan, "The ferromagnetic effect at microwave frequencies and its applications; the microwave gyrator," *Bell Sys. Tech. J.*, vol. 31, pp. 1–31; January, 1952.

component (usually a resistor) as shown in Fig. 4(a).[5] Fig. 4(b) shows a dual of McMillan's isolator. Black and Scott realized such a system by a combined piezoelectric-piezomagnetic vibrator made of a bar of X-cut quartz bonded to a bar of stainless steel.[6] More recently, piezoelectric ceramics and piezomagnetic ferrites with high electromechanical coupling have been used to obtain greatly improved characteristics by Germano and Curran,[7] and Curran, *et al.*,[8] and by the authors.[9]

The isolator described by McMillan has one severe drawback in comparison with the ideal isolator. Unilateral transmission can be obtained only for a single frequency or a group of frequencies, because the transfer immittances of combined vibrators vary widely with frequency. Gamo established, however, a general method for obtaining electromechanical systems, unilateral at all frequencies.[10] These systems are named "achromatic" isolators and are shown in Fig. 5. Black boxes are the same reciprocal four-terminal circuits having the transmission matrix shown in the figure. Cascade connection of the box with an ideal gyrator as shown in the upper branch yields an antireciprocal circuit, which may be realized by a piezoelectric-piezomagnetic combined vibrator similar to that previously discussed. The box in the lower branch may be realized by either a purely piezoelectric or a purely piezomagnetic vibrator. It should be noted that these isolators need two resistors instead of one as used in the ideal isolator. This increases the loss in the forward transmission but provides the advantages of achromatic, unilateral transmission characteristics. Realization of Gamo's isolator in practice, however, is extremely difficult, because it is necessary to match the frequency characteristics of the two vibrators in the upper and lower branches.

In this paper a novel type of electromechanical, achromatic isolator is presented. The isolator consists of three mechanically coupled transducers, two of which may be piezoelectric and the third piezomagnetic, or two of which may be piezomagnetic and the third piezoelectric. In either arrangement a unilateral coupling between mechanical and electrical systems is provided using one piezoelectric and one piezomagnetic transducer. The third transducer in merely for driving the mechanical system. Such a design makes manufacturing considerably easier, because there is only one mechanical system and because adjustment by purely electrical means is possible after fabrication of the mechanical parts.

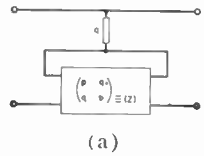
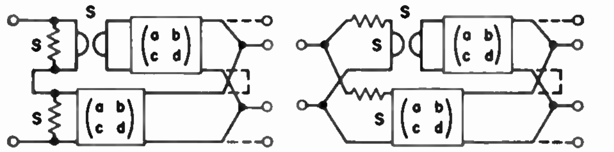


Fig. 4—Isolator after McMillan and its dual.



Dotted lines show the alternative series connection for the output.

Fig. 5—Achromatic isolators after Gamo.

II. The Electromechanical Gyrator

Fig. 6 shows a few configurations of electromechanical gyrators consisting of three transducers. The hatched portion is made of piezoelectric ceramic material and bonded to the rest of system made of piezomagnetic ferrite. The terminals 1-1′ are for a piezomagnetic transducer and the terminals 2-2′ are for a piezoelectric transducer. Combination of both transducers provides an achromatic unilateral transmission between electrical and mechanical systems. The terminals 3-3′ are for the third transducer, which may be piezoelectric or piezomagnetic.

In practice the piezoelectric third transducer is preferred. This is because there will be less stray coupling between the third transducer and the other transducers and because a higher electromechanical coupling coefficient is obtainable in piezoelectric ceramics than in ferrites. Hence all the examples except (a) in Fig. 6 have piezoelectric third transducers. Configurations (a), (b) and (c) may be called cascaded structures, and (d) and (e) sandwiched structures. Configurations (c) and (e) use a piezoelectric transducer with divided electrodes and, therefore, need only one bonding operation.

[5] E. M. McMillan, "Violation of the reciprocity theorem in linear passive electromechanical systems," *J. Acoust. Soc. Am.*, vol. 25, pp. 344–347; October, 1946.

[6] L. J. Black and H. J. Scott, "Measurements on nonreciprocity in an electromechanical system," *J. Acoust. Soc. Am.*, vol. 25, pp. 1137–1140; November, 1953.

[7] C. P. Germano and D. R. Curran, "Low frequency gyrators," Electronic Components Conf., San Francisco, Calif., Paper No. 24; May, 1961.

[8] D. R. Curran, J. H. Silverman and J. Schoeffler, "Passive electromechanical gyrators and isolators," private communication.

[9] M. Onoe and M. Sawabe, "A piezoelectric and piezomagnetic gyrator," *Proc. Spring Meeting Acoust. Soc.*, Japan, Paper No. 91; May, 1961.

[10] H. Gamo, "Four terminal networks violating reciprocal theorem and one-way systems," *Proc. 26th Joint Conf. of Elec. Engrg. Soc.*, Paper No. 9-1; 1952.

—— "Electromechanical one-way systems," *J. Acoust. Soc. (Japan)*, vol. 10, pp. 65–76; June, 1954.

—— "On passive one-way systems," *Trans. Internat'l Symp. on Circuit and Information Theory*, pp. 283–298; 1959.

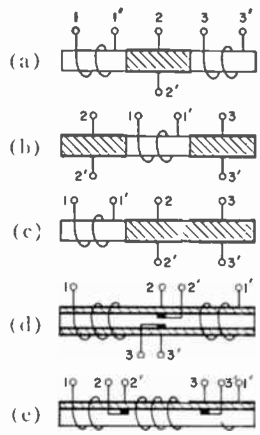


Fig. 6—Electromechanical gyrators consisting of three transducers.

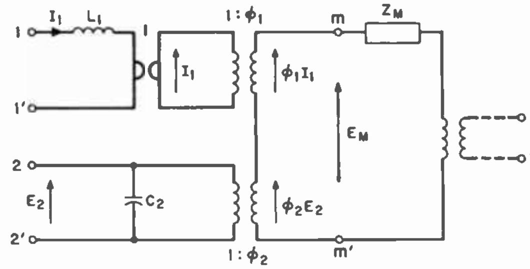


Fig. 7—Equivalent circuit of the gyrator.

III. Equivalent Circuit

Although Gamo's theorem can be extended to cover unilateral transmission between electrical and mechanical systems, the following analysis based on an equivalent circuit is much simpler. Fig. 7 shows the equivalent circuit based on the voltage force analogy near a mechanical resonance of composite structure. L_1 is the clamped inductance of the piezomagnetic transducer, while C_2 is the clamped capacitance of the piezoelectric transducer. Z_M represents a mechanical resonance of the composite structure and is a series resonant circuit as shown in the figure. It should be noted, however, that a more elaborate structure can be inserted between the first two transducers and the third transducer, so that much more complicated frequency characteristics like those of a mechanical filter may be obtained.

The condition for suppressing the transmission from the first two transducers to the third transducer can be described as follows:

$$\phi_1 I_1 + \phi_2 E_2 = 0 \tag{1}$$

where ϕ_1 and ϕ_2 are the transformer ratios of the piezomagnetic and piezoelectric transducers, respectively. It is assumed here that ϕ_1, instead of the gyrator constant, has a dimension of resistance. I_1 is a driving current for the piezomagnetic transducer and E_2 is a driving voltage for the piezoelectric transducer; both are to be derived from the same signal source. Such distributions of current and voltage are possible in the two circuits shown in Fig. 8, where

$$\frac{L_1}{C_2} = R_1 R_2 \tag{2}$$

$$-\frac{\phi_1}{\phi_2} = R_1. \tag{3}$$

In this figure the mechanical portion is omitted because there is no current in that portion under the condition

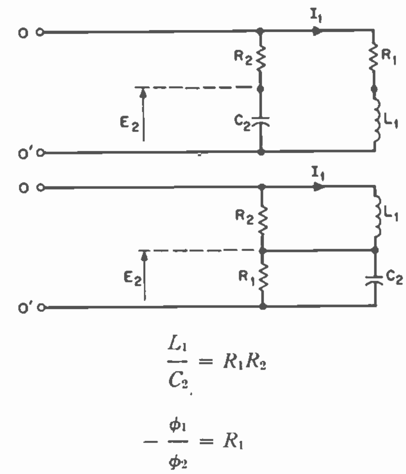


Fig. 8—Realization of achromatic suppression of transmission by means of associated resistances.

(1). It can be seen that the conditions (2) and (3) are independent of frequency and, therefore, an achromatic suppression of transmission is achieved. The need for two resistors is consistent with Gamo's theorem.

The values of L_1, C_2 can be measured at a frequency well removed from mechanical resonances. The ratio ϕ_1/ϕ_2 is conveniently obtained by measuring, first, the output voltage of the third transducer for a given input current I_1 into the piezomagnetic transducer with a short circuit of the terminals 2-2′, and second, the output voltage for a given input voltage E_2 into the piezoelectric transducer with an open circuit of the terminals 1-1′. The other parameters can be determined by a conventional method for measuring electromechanical vibrators as described in the Appendix.

IV. Condition for Constant Input Resistance

It can be seen that if the two resistances determined by (2) and (3) happen to be equal, the input impedance looking into the terminals 0-0′ becomes a constant resistance R, which is also equal to these two resistances,

$$R = R_1 = R_2. \tag{4}$$

This condition is interesting in practice because it simplifies the matching at the terminals 0-0′. The condition can be realized by a proper design of the physical dimensions of a vibrator. The realization is also possible by purely electrical means even after the fabrication of the vibrator.

Roughly speaking, ϕ_1, and by virtue of (3) also R_1, are proportional to the number of turns of the coil N, while L_1 is proportional to N^2. This yields R_2 proportional to N according to (2). Hence changing the number of turns of the coil in order to obtain a constant input resistance has little effect. On the other hand, the magnetic biasing field affects ϕ_1 considerably but L_1 only slightly. Hence the condition (4) is easily obtained by adjusting the biasing field, especially when the original R_1 is larger than R_2. When the original R_2 is larger than R_1, the condition (4) is achieved by merely adding some capacitance parallel to C_2.

V. Forward Transmission Characteristics

The tranmission characteristics in the forward direction, namely from the terminals 3-3′ to the terminals 0-0′, can be conveniently described based on the equivalent circuit shown in Fig. 7. The portion left of the terminals m-m' is simplified under conditions (2) and (3) as shown in Fig. 9(a) and (b). Further simplification is possible as shown in Fig. 10 if condition (4) is satisfied. Since this condition is preferable in practical applications and can always be achieved as discussed in the previous section, the present discussion is limited to this special case. The right half section of the equivalent circuits shown in Fig. 10 is a conventional filter circuit, characteristics of which have been well discussed in literature. If the mechanical loss is negligible, a perfect matching at the resonance frequency is possible by the use of a tuning inductance parallel to C_3. This yields the available power of the source, P_0,

$$P_0 = \frac{E^2}{4(2\phi_2{}^2 R)}. \qquad (5)$$

while the available power of the left half section, P_L, is,

$$P_L = \frac{E_M{}^2}{4\phi_2{}^2 R}. \qquad (6)$$

where $E_M = E/2$ at the resonance. Hence, at the resonance, the ratio P_0/P_L is two, and this yields the minimum insertion loss of 3 db. Any mechanical loss or any mismatching increases the insertion loss.

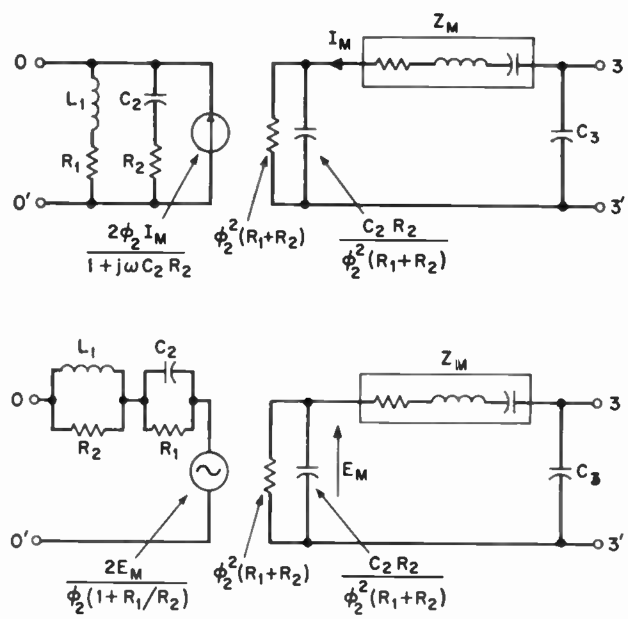


Fig. 9—Equivalent circuits for achromatic isolators.

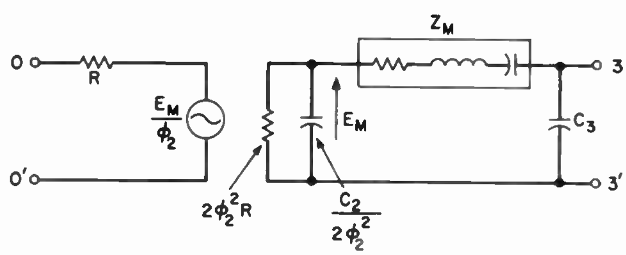


Fig. 10—Equivalent circuit in the case $R = R_1 = R_2$.

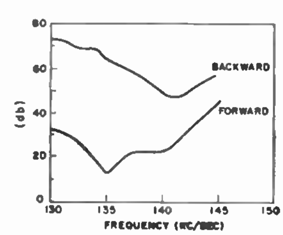


Fig. 11—Characteristics of an achromatic isolator.

VI. Experiments

An achromatic isolator operating at 135 kc was made of barium titanate and nickel-ferrite and reported in a previous paper.[9] The cascaded structure shown in Fig. 6(b) was used. The frequency characteristics are shown in Fig. 11. The 3-db bandwidth is 1.2 kc and unilateral discrimination of more than 50 db is obtained. An irregularity at 141 kc is due to an unwanted mode of vibration and may be removed by an improved mechanical design. The high forward insertion loss of 12 db is due to a poor electromechanical coupling of material and a mechanical loss.

In order to improve the forward insertion loss, Clevite PZT-5 ceramic ($k_{33} = 0.68$) and Kearfott N-51 ferrite ($k = 0.40$) have been used. Table I shows equivalent parameters for three models. Gyrator A uses the sandwiched structure shown in Fig. 6(d). Resistance values experimentally obtained for the best suppression of backward transmission are compared with the calculated values from (2) and (3). The discrepancy seems chiefly due to an error in determining the ratio ϕ_1/ϕ_2 (about 10 per cent). The condition for constant input resistance discussed in Section IV was achieved

TABLE I

Equivalent Circuit Parameters of Gyrators and Their Characteristics

Gyrator Type	A Fig. 6(d)	B Fig. 6(e)	C Fig. 6(b)	Unit
L_1	103	120	140	μh
C_2	720	330	440	pf
C_3	720	330	440	pf
f_0	152.33	164.32	113.42	kc
Q_M	246	299	157	—
L_M	165	522	340	mh
C_M	6.66	1.81	5.77	pf
R_M	640	1800	1.55	Ω
R_M'	630	1800	1.53	Ω
ϕ_2	1.01	1.00	1.01	—
$-\phi_1/\phi_2$	266	626	250	Ω
Calculated resistances from (2) and (3)				
R_1	266	626	250	Ω
R_2	537	580	1270	Ω
Resistances experimentally obtained for the best backward suppression				
R_1	281	565	—	Ω
R_2	460	578	—	Ω
Resistances experimentally obtained for the condition of constant input resistance				
R_1	308	—	236	Ω
R_2	304	—	236	Ω
Minimum insertion loss				
forward	5.0	6.0	6.0	db
backward	46	43	49.5	db
3-db bandwidth	2.9	1.5	2.0	kc

by adding a capacitance of about 375 pf parallel to C_2 and adjusting slightly the bias magnet. This yields experimental values shown in Table I. Fig. 12 shows the open circuit impedances of the terminals 0-0′ and 3-3′ measured by a Wayne-Kerr bridge, type B601. It can be seen that the former is almost constant and equal to R_1 and R_2 as the theory predicts. Insertion loss characteristics under matched termination conditions are shown in Fig. 13. A fairly flat suppression of backward transmission over the measured range of frequency was obtained as expected. The forward loss was reduced to 5 db, of which 3 db is the theoretical minimum and 2 db is an excess loss due to mechanical losses.

Gyrator B uses the sandwiched structure with divided electrodes shown in Fig. 6(e). Dividing the electrodes reduces the coupling of the piezoelectric transducer (2-2′) relative to that of the piezomagnetic transducer. This increases the ratio ϕ_1/ϕ_2 and achieves the condition for constant input resistance without any additional adjustment as seen in Table I. Fig. 14 shows the impedance characteristics and Fig. 15 (next page) the insertion loss characteristics under matched termination conditions. An irregularity in the backward insertion loss seems due to the off-nodal support of this particular model and can probably be removed by a slight change of the electrode configuration which allows the usual nodal support. Figs. 16 and 17 are the impedance and the insertion loss characteristics, respectively, of Gyrator C which uses the cascaded structure shown in Fig. 6(b). A capacitance of about 2270 pf is added parallel to C_2 in order to achieve the condition for constant input resistance.

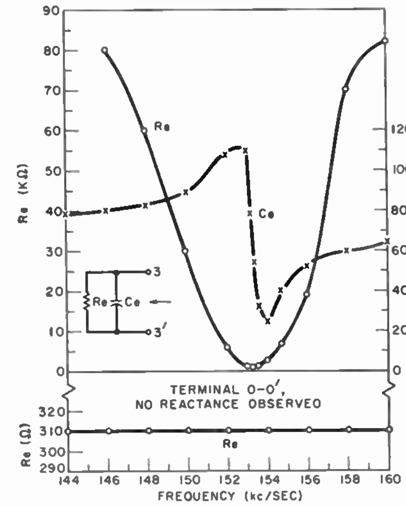


Fig. 12—Input impedances of isolator A.

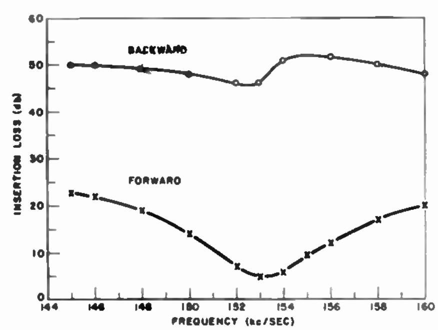


Fig. 13—Insertion loss of isolator A.

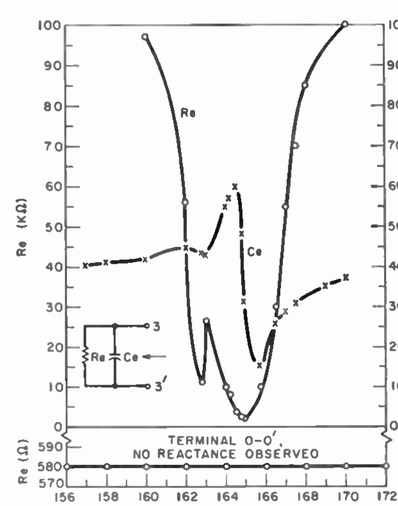


Fig. 14—Input impedances of isolator B.

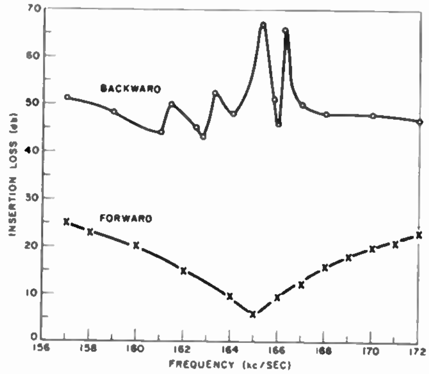


Fig. 15—Insertion loss of isolator B.

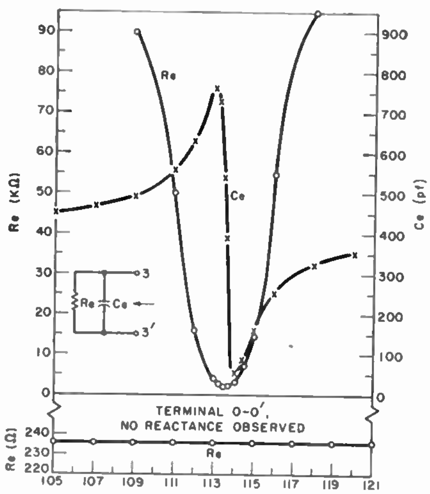


Fig. 16—Input impedances of isolator C.

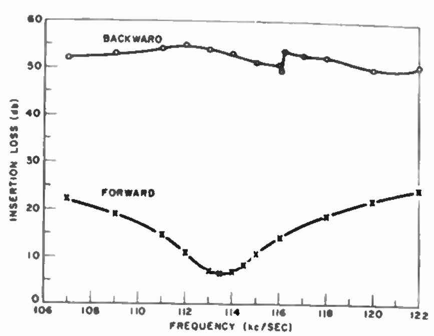


Fig. 17—Insertion loss of isolator C.

VII. Conclusion

A linear passive isolator, which has an achromatic suppression of backward transmission, has been realized by combining two resistances and an electromechanical gyrator according to Gamo's theory. The gyrator consists of three mechanically coupled, ceramic elements, two of which may be piezomagnetic and the third piezoelectric or two of which may be piezoelectric and the third peizomagnetic. In either arrangement a unilateral coupling between the mechanical and the electrical system is provided by using one piezoelectric and one piezomagnetic element. The third element is merely for driving the mechanical system.

Characteristics of the gyrator have been discussed based on an equivalent circuit. A constant input resistance can be obtained at a terminal pair of the isolator by purely electrical means even after the fabrication of the gyrator. This simplifies the matching at this terminal. The theoretical minimum insertion loss is 3 db under the matched termination conditions. A few models have been made using both the sandwiched and the cascaded structures. A highly achromatic suppression of backward transmission has been obtained in agreement with theoretical predictions. Minimum forward insertion loss consists of the theoretical minimum of 3 db and an excess loss of a few db due principally to mechanical losses.

Appendix I

Determination of Equivalent Circuit Parameters

The parameters of the equivalent circuit shown in Fig. 7 can be determined in the following manner. ϕ_3 can be set equal to one without loss of general validity.

1) Clamped parameters L_1, C_2 and C_3 are measured at a frequency well above the mechanical resonance.

2) Admittance measurement at the terminals 3-3′ with the terminals 2-2′ shorted and the terminals 1-1′ open yields the resonant frequency f_0 and the resonance resistance R_M.

3) Admittance measurement at the terminals 2-2′ with the terminals 3-3′ shorted and the terminals 1-1′ open yields the same resonant frequency and another resonance resistance R_M'. The transformer ratio ϕ_2 is given by the following:

$$\phi_2 = (R_M/R_M')^{1/2}. \tag{7}$$

4) The above two admittance measurements determine the quadrant frequencies f_1 and f_2 which correspond to conductances equal to half of the resonance conductance. Then

$$Q_M = \frac{f_0}{f_1 - f_2} \tag{8}$$

$$L_M = \frac{Q_M R_M}{2\pi f_0} \tag{9}$$

$$C_M = \frac{1}{(2\pi f_0)^2 L_M}. \tag{10}$$

It may be noted that the clamped capacitances C_2 and C_3 can also be obtained from the algebraic mean of susceptances at the quadrant frequencies.

5) The ratio ϕ_1/ϕ_2 is determined as described in the text. Since ϕ_2 is already known, ϕ_1 can be determined. It should be noted that impedance measurements at the terminals 1-1′ are possible, but the separation of parameters is much more complicated because of the loss of the coil.

Appendix II

Derivation of Equivalent Circuits for Achromatic Isolators

The portion left of the terminals m-m' in Fig. 7, associated with resistances in the manner shown in Fig. 8(a), yields a four-terminal network with terminals o-o' and m-m'. Let voltages across and currents flowing into these terminals be E, I, E_M and I_M, respectively, then the following equations hold:

$$E_M = \phi_1 I_1 + \phi_2 E_2, \tag{11}$$

$$E + \phi_1 I_M = (R_1 + j\omega L_1) I_1, \tag{12}$$

$$E - E_2 = R_2(I - I_1), \tag{13}$$

$$\phi_2 I_M = j\omega C_2 E_2 - I + I_1, \tag{14}$$

$$R_2 \phi_2 I_M = (1 + j\omega C_2 R_2) E_2 - E. \tag{15}$$

These equations together with (2) and (3) yield the following:

$$E_M = \frac{\phi_2^2(R_1+R_2)}{1+j\omega C_2 R_2} I_M, \tag{16}$$

$$I + \frac{2\phi_2}{1+j\omega C_2 R_2} I_M = \left(\frac{1}{R_1+j\omega L_1} + \frac{j\omega C_2}{1+j\omega C_2 R_2}\right) E, \tag{17}$$

which correspond to the left portion of equivalent circuits shown in Fig. 9(a). If the condition of constant input resistance holds, as given by (4), then (16) and (17) can be further reduced to the following:

$$E_M = \frac{2\phi_2^2 R}{1 + j\omega C_2 R} I_M, \tag{18}$$

$$I = \frac{E}{R} - \frac{E_M}{\phi_2}, \tag{19}$$

which correspond to the left portion of the equivalent circuit shown in Fig. 10

In the case of Fig. 8(b) the following similar set of equations apply instead of (11) to (15):

$$E_M = \phi_1 I_1 + \phi_2 E_2, \tag{20}$$

$$I = \frac{E_1}{R_2} + \frac{E_1 + \phi_1 I_M}{j\omega L_1} \tag{21}$$

$$= \left(\frac{1}{R_1} + j\omega C_2\right) E_2 - \phi_2 I_M, \tag{22}$$

$$E = E_1 + E_2, \tag{23}$$

where E_1 is the voltage across the resistance R_2. These equations together with (2) and (3) yield the following:

$$E_M = \frac{\phi_2^2(R_1 + R_2)}{1 + j\omega C_2 R_2} I_M, \tag{24}$$

$$E = \left(\frac{j\omega L_2 R_2}{R_2 + j\omega L_1} + \frac{R_1}{1 + j\omega C_2 R_1}\right) I + \frac{2R_2}{\phi_2(R_1 + R_2)} E_M, \tag{25}$$

which correspond to the left portion of the equivalent circuit shown in Fig. 9(b). If the condition of constant input resistance holds, these equations also reduce to (18) and (19).

Acknowledgment

The authors are grateful for the advice of Prof. N. Takagi. The models shown in Figs. 12 to 17 were made through the courtesy of Bell Telephone Laboratories. Thanks are due to W. J. Nowotarski for fabrication and J. J. Gallo for measurement. Drs. A. H. Meitzler and J. E. May, Jr., kindly read the English text.

Correction

E. I. Jury, author of "A Simplified Stability Criterion for Linear Discrete Systems," which appeared on pages 1493–1500 of the June, 1962, issue of Proceedings, has requested that the following corrections be made in the paper.

1) The first term of the right side of (26) should read

$$A_n A_{n-2}.$$

2) The numbering in the sentence following (52a) should be changed as follows:

$$(53) \to (50)$$
$$(52) \to (51)$$
$$(56) \to (53)$$
$$0 + 51 \to (51)$$

3) In part (b) on page 1499, *n odd* should be changed to *n even*, and *n even* should be *n odd*.

4) On page 1499, the two sentences preceding the section of the *Number of Roots of a Real Polynomial Inside the Unit Circle*, should be changed as follows:

> Furthermore, if the system is initially stable and some of its parameters change so as to approach instability, the constraints $A_{n-1} - B_{n-1}$ or $F(1)$, $F(-1)$ are the most critical. Hence, to obtain the maximum parameter variations for stability limit only the equations $A_{n-1} - B_{n-1} = 0$ and $F(1) = 0$, $F(-1) = 0$ need be solved. A detailed discussion of this fact will be presented at a later date.

How to Obtain the IRE Standards on Electron Tubes: Methods of Testing, 1962*

THIS MONTH THE IRE is publishing, as a separate volume, the largest Standard in its history: the "IRE Standards on Electron Tubes: Methods of Testing, 1962." Because of its large size, 160 pages, it was not feasible to follow the normal practice of including it in the PROCEEDINGS. However, *all IRE members and* PROCEEDINGS *subscribers on record as of September 1st may obtain one free copy*[1] *of this standard upon request*, by writing to IRE Headquarters, 1 East 79 Street, New York 21, N. Y. Ask for the IRE Standard No. 62 IRE 7. S1.

This Standard, which deals with the methods of measurement of the important characteristics of electron tubes, should be of great interest to electrical engineers, radio engineers, and physicists. It is unique since it contains complete information on the testing of all kinds of electron tubes including descriptions of test equipments and precautions to be observed to insure validity of test results. A more detailed description of the Standard is given below.

The Standard is the result of seven years activity of the IRE Electron Tubes Committee, during which 167 dedicated and experienced research and engineering workers in the field of electron tubes devoted approximately 20,000 man-hours of careful deliberation. This large expenditure of time and effort does not include outside investigation of source materials by the various workers, as well as time spent contacting other agencies interested in electron tubes. If this time were included, the number of man-hours devoted to this document could exceed 50,000.

Most of the Methods of Testing contained in the Standard were developed after thorough consultation with persons engaged in this field who were employed either by private industries, universities, or government agencies, in Canada, England, France, Japan, Netherlands, Poland, Switzerland, West Germany, and the United States. Among a number of organizations engaged in standardization work on electron tubes the IRE Electron Tubes Committee is pleased to acknowledge the cooperation and interest of the following organizations:

The American Standards Association
American Institute of Electrical Engineers
British Radio Valve Manufacturers' Association
Electronic Industries Association
International Electrotechnical Commission
Joint Electron Device Engineering Council
Nachrichtentechnische Gesellschaft im Verband Deutscher Elektrotechniker.

The Committee also gratefully acknowledges the work of the members of the IRE Standards Committee who spent many hours of deliberation before approving this document.

The IRE Electron Tubes Committee would appreciate very much hearing from users of this document relative to criticisms and suggestions for future proposed methods of testing electron tubes. Please send all communications to George A. Espersen, Chairman, IRE Electron Tubes Committee, The Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y.

CONTENTS OF STANDARD

The Standard consists of 10 Parts, as indicated below. Parts 1, 2, and 3, are completely revised versions of "IRE Standards on Electron Tubes: Methods of Testing, 1950" (50 IRE 7.S1).[2] Part 9 includes minor revisions of "IRE Standards on Methods of Measuring Noise in Linear Twoports, 1959" (59 IRE 20.S1).[3] The remaining six Parts of the Standard contain totally new, previously unpublished methods of testing.

The principal measurements covered by each Part are outlined below. In addition, many of the Parts provide references and bibliographies and give new definitions of terms.

PART 1: CONVENTIONAL RECEIVING TUBES

Filament or Heater Characteristics
Emission Tests
Characteristics of an Electron Tube
Residual Gas and Insulation Tests
Inverse Electrode Currents
Vacuum-Tube Admittances
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PART 2: CATHODE-RAY TUBES

Cutoff Voltage
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Focusing-Electrode Voltage of Electrostatic-Focus Types
Focusing-Coil Current of Magnetic-Focus Types

* Library of Congress Catalog Card Number: 62-20746.
[1] Additional copies may be obtained at the nonmember price of $2.50 per copy.

[2] PROC. IRE, vol. 38, August and September, 1950.
[3] PROC. IRE, vol. 48, January, 1960.

Correspondence

A Proposed First-Order Relativity Test Using Lasers*

The new-found ability to mix together the outputs of two coherent optical sources, such as gas lasers, and obtain stable beats, raises the possibility of a new test for ether drift in which the effect to be measured is proportional to the first power of v/c, where v is the expected ether drift.

The classical Michelson-Morley experiments, using one optical source, measure an effect proportional to v^2/c^2 because of the necessity of returning the light beam to its starting point. The difference in transit times, go and return, is of the form

$$L/(c-v) + L/(c+v) = 2L/c \cong (2L/c)(v^2/c^2)$$

where L is the one-way path length. If one could arrange to measure directly the difference in transit times for go and return, this would be in the form

$$L/(c-v) - L/(c+v) \cong (2L/c)(v/c),$$

greater than the Michelson-Morley effect by the factor c/v.

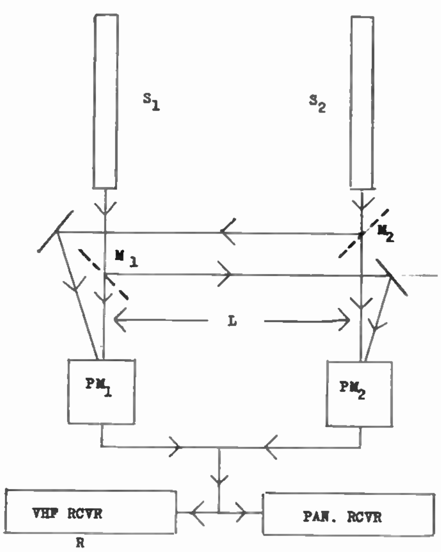

Fig. 1

Referring to Fig. 1, S_1 and S_2 are two CW lasers. Half-silvered mirrors M_1 and M_2 split the two beams, sending one half directly into the photomultipliers PM_1 and PM_2, and the other half over the optical path L. At the ends of L the beam from S_1 is reflected into PM_2, and the beam from S_2 into PM_1. The outputs of PM_1 and PM_2, which contain VHF components due to the mixing of the two optical laser frequencies, are added and their sum is measured by the VHF receiver R. A panoramic receiver is added at this point to facilitate spectrum examination. The two lasers are mounted vertically with the optical path horizontal, and the whole assembly is designed to rotate in the horizontal plane.

For simplicity let us assume that the two lasers are adjusted for one mode of oscillation only, and that they are offset in frequency by 100 mc. There will then be 100-mc components in the outputs of the mixers, and their sum will be selected and measured by the VHF receiver. Let us suppose that in a position of zero ether drift in the direction of L, the path length of one laser is adjusted to make the input to the receiver a maximum, *i.e.*, to assure that the two mixer outputs are in phase at the receiver. Now let us rotate the apparatus through 90° so that L is in the direction of the ether drift v. If this drift is from left to right in Fig. 1, the phase of the waves from S_1 at PM_2 is advanced by the angle $\phi = 2\pi(L/\lambda)(v/c)$, where λ is the wavelength, and the phase of the waves from S_2 at PM_1 is retarded by the same angle. Since the phase changes in the optical signals are transferred without loss to the phases of the mixer outputs, there will be a relative change in phase of 2ϕ between the 100-mc signals at the input to the receiver. If the intensities are equal and $\phi = \pi/2$, there will be complete cancellation; intermediate values will be a measure of the phase shift.

Note that the success of this method is due to the ability to measure the relative phase advance and retardation of the optical frequency components at the ends of the optical path by transferring these phase shifts to a greatly lower frequency in the VHF region, so that they may be brought together for comparison with a negligible compensation of the effect.

For a helium-neon laser, with $\lambda = 1.153 \times 10^{-4}$ cm, L may be as short as 30 cm for a phase shift of 180° with $v/c = 10^{-6}$, corresponding to an ether drift of 0.3 km/sec, the surface velocity of the earth in temperate latitudes due to its rotation. This short length is desirable since it simplifies problems of mechanical rigidity and also makes magnetic shielding possible if changes in magnetic field during rotation turn out to be bothersome.

In the interests of simplicity and lack of fussiness, confocal resonators should undoubtedly be used with the lasers. The presence of many modes in these resonators will produce a complicated spectrum of intermodal beats from the mixers. The panoramic receiver may then be used to select interlaser beats by noting components that are present in the mixer outputs only when both lasers are operating.

It is not anticipated that frequency drift in the lasers will be a problem, provided that the change in the beat frequency will not be so large or rapid as to be unmanageable in the receiver. A slight error will occur in the phase comparison if the frequency of one laser changes during rotation; for a 1-mc shift in one laser the error in the phase angle will be only 0.002 radians. Actually, the apparatus should be small enough to permit fairly rapid rotation.

As has been pointed out by this author,[1] the Michelson-Gale experiment[2] has its simplest nonrelativistic explanation in terms of an ether that does not rotate with the earth. Since this experiment has been explained by the theory of general relativity, and since the precision of the Michelson-Morley experiments was not great enough to detect a rotational ether drift, it is of considerable importance to try a similar method that does have the capability.

C. W. Carnahan
120 Fawn Lane
Menlo Park, Calif.

* Received June 13, 1962.

[1] C. W. Carnahan, "Light and gravitation," Proc. IRE (to be published).

[2] A. A. Michelson and H. G. Gale, *Astrophys. J.*, vol. 61, p. 140, 1925; also, G. Joos, "Theoretical Physics," Hafner Publishing Company, Inc., New York, N. Y., 2nd ed., p. 472, 1950.

A Light Source Modulated at Microwave Frequencies*

When a gallium arsenide *p-n* junction is biased in the forward direction, radiative band-to-band recombination is observed.[1] Since minority-carrier lifetimes of the order of 10^{-10} sec are readily obtained in GaAs, one may expect that the recombination radiation can be modulated at Gc rates. This communication reports a verification that efficient generation of light modulated at microwave frequencies is possible.

The current through a GaAs diode increases very rapidly when it is forward biased with an increasing voltage nearly equal to the energy gap (about 1.5 volts). Under this bias condition, the current consists of tunnel-assisted radiative band-to-band recombination in the space-charge region of the *p-n* junction.[2] This radiation occurs in a narrow spectral band in the near infrared (0.84 μ at 77°K). The intensity of the light output first increases very rapidly (more than linearly) with current and then linearly. In the linear range the process is extremely efficient. A quantum efficiency of 0.50 to 1.00 photons/electron has been obtained. However, with the geometry used in our experiment only about 1 per cent of the radiation comes out of the specimen. The over-all power efficiency of the light source is also somewhat reduced by a small ohmic loss due to the internal resistance of the diode.

The following measurements were made with a diode fabricated by alloying a tin dot

* Received June 2, 1962.

[1] J. I. Pankove and M. Massoulié, "Injection luminescence from GaAs," *Bull. Am. Phys. Soc.*, vol. 7, p. 88; January, 1962.

[2] J. I. Pankove, "Tunneling assisted photon emission in Gallium Arsenide *p-n* junctions," to be published.

to p-type GaAs having a hole concentration of 2.5×10^{18} cm^{-3}. The diode was mounted in series with a 50-ohm resistor at the end of a 50-ohm coaxial cable connected to a signal generator. The diode end of the cable was inserted in a Dewar filled with liquid nitrogen (Fig. 1). The radiation was collected through the two windows of the Dewar by a lens and focused onto a photomultiplier (RCA 7102) having an S-1 spectral response. The output of the photomultiplier was displayed on an oscilloscope. Fig. 2 shows the detection of 200-Mc modulation as displayed on a sampling oscilloscope. A dc bias was inserted in series with the generator to operate the diode in the light-emitting mode. The noise is believed to originate in the photomultiplier.

In its nonlinear range, the radiation from the diode is also modulated at harmonics of the driving frequency. This is illustrated in Fig. 3 where the upper curve (d) is a 6-Mc driving signal, and the lower curve (c), the photomultiplier output. (a) is the zero level for the photomultiplier output. The diode being insufficiently biased to give a linear light output, as the signal swings about the dc level (b), the light output is not symmetrical during the brightening and dimming half-cycles. The distortion of the driving voltage is due to the changing load impedance as the diode conductance increases.

The frequency limitation of our measurements is due to the transit time dispersion of electrons in the photomultiplier. Hence, an operating frequency of 200 Mc is not the upper limit for the diode. The RC limitation of this diode is of the order of 10 Gc.

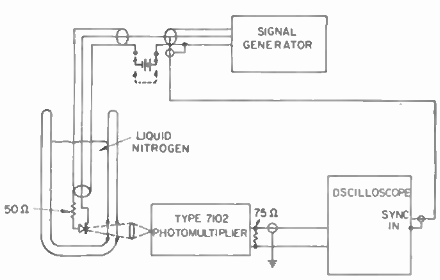

Fig. 1—Schematic diagram of test setup.

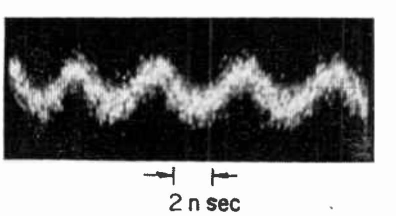

Fig. 2—Detection of an optical signal modulated at 200 Mc as displayed by a pulse sampling oscilloscope.

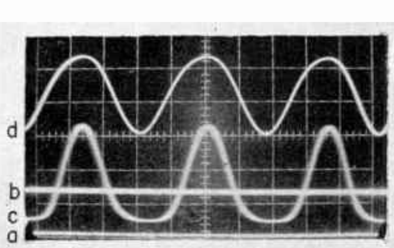

Fig. 3—Output of photomultiplier. (a) No light when no current flows through diode (b) when 60-ma dc forward bias flows through diode (c) when driving signal (d) is superposed on the dc current.

As was stated above, only about 1 per cent of the radiation leaves the specimen through the surface opposite the p-n junction. This light comes out in a 2π-steradian solid angle. An improvement of one to two orders of magnitude in light collection can be obtained by shaping the specimen into a Weierstrass sphere.[3]

We wish to thank G. B. Herzog for valuable suggestions.

J. I. Pankove
J. E. Berkeyheiser
RCA Laboratories
Princeton, N. J.

[3] P. Aigrain and C. Benoit-a-la-Guillaume, "Emission Infrarouge du Germanium," *J. de Phys. et Radium*, vol. 17, pp. 709–711; August–September, 1956.

Three-Dimensional Parametric Interactions of Waves and Quasi-Particles*

The parametric interaction in a cavity can be regarded as point interactions. The investigations of traveling-wave parametric interactions by Tien,[1] among others, extended the understanding to one-dimensional interactions. The purpose of this note is to show that the concept of these parametric interactions can be further generalized to two and three dimensions, and interpreted as scattering of coherent waves or quantized fields of quasi-particles. These concepts apply not only to electromagnetic waves, but also to interactions involving elastic waves, spin waves, plasma waves, etc. As quantized fields, parametric interactions can be interpreted as the annihilation or creation of photons, phonons, magnons, plasmons, etc.

In order to demonstrate the possibility of achieving parametric interaction of traveling waves in three-dimensional media, we choose a moving coordinate system which is moving at an arbitrary velocity $\mathbf{v}$. Let the frequencies of the original traveling waves be ω_p, ω_i, and ω_s for the pump, idler, and signal, respectively. The corresponding Doppler frequencies in the moving coordinates become ω_p', ω_i' and ω_s'. Then

$$\begin{aligned} \omega_p' &= \omega_p - \boldsymbol{\beta}_p \cdot \mathbf{v} \\ \omega_i' &= \omega_i - \boldsymbol{\beta}_i \cdot \mathbf{v} \\ \omega_s' &= \omega_s - \boldsymbol{\beta}_s \cdot \mathbf{v}, \end{aligned} \tag{1}$$

where $\boldsymbol{\beta}_p$, $\boldsymbol{\beta}_i$, and $\boldsymbol{\beta}_s$ are the phase constants of the three traveling waves. $\boldsymbol{\beta}$ and $\mathbf{v}$ are vector quantities. Parametric interaction becomes possible if the pump frequency is equal to the sum of the idling and signal frequencies, and if the relationship holds for any arbitrary velocity of the moving coordinate system. That means we have

$$\omega_p = \omega_i + \omega_s, \tag{2}$$

and require that

$$\omega_p' = \omega_i' + \omega_s'. \tag{3}$$

From the above three equations we get

$$\boldsymbol{\beta}_p = \boldsymbol{\beta}_i + \boldsymbol{\beta}_s. \tag{4}$$

Eq. (4) is the condition on the phase constants. Conversely if (2) and (4) hold, (3) becomes valid and parametric interaction is possible. Thus, (2) and (4) are the selection rules to be satisfied for traveling-wave parametric interactions.

For periodic structures such as crystals,[2] (4) is equivalent to

$$\boldsymbol{\beta}_p = \boldsymbol{\beta}_i + \boldsymbol{\beta}_s \mp 2\pi \mathbf{g}, \tag{4a}$$

where $\mathbf{g}$ is a lattice vector in the reciprocal lattice. When $\mathbf{g}$ is not zero, the interaction corresponds to the so-called "umklapp" process in solids.[3]

In the one-dimensional case, the phase constants can be regarded as scalars with either positive or negative signs. Then (2) and (4) are reduced to Tien's equations.[1] In a nonlinear medium it is possible to achieve either forward or backward traveling-wave parametric amplifications. The significance of the umklapp process has been demonstrated in backward traveling-wave parametric amplifiers.[4]

Eqs. (2) and (4) can be expressed as

$$\hbar\omega_p = \hbar\omega_i + \hbar\omega_s, \tag{5}$$

$$\hbar\boldsymbol{\beta}_p = \hbar\boldsymbol{\beta}_i + \hbar\boldsymbol{\beta}_s, \tag{6}$$

and

$$\hbar\boldsymbol{\beta}_p = \hbar\boldsymbol{\beta}_i + \hbar\boldsymbol{\beta}_s + h\mathbf{g}, \tag{6a}$$

where $\hbar$ is Planck's constant, h, divided by 2π.

Eqs. (5), (6) and (6a) can be regarded as the particle aspect of traveling-wave interactions. Eq. (5) indicates the conservation of energy, (6) the conservation of momentum, and (6a) the conservation of crystal momentum. It is expected that (5), (6), or (6a) will be satisfied in any scattering processes. Thus, parametric interactions due to scattering of the coherent quantized fields of quasi-particles can conceivably be achieved.

The above discussion can be readily extended to frequency mixing, harmonic generation and parametric interactions involving multiple frequencies.[5] The selection rules, corresponding to the conservation laws, can be generalized as

$$\sum_i E_i = \sum_s E_s \tag{7}$$

* Received February 19, 1962. This work was supported in part by U. S. Signal Corps Contract No. DA-36-039-SC-87209.

[1] P. K. Tien, "Parametric amplification and frequency mixing in propagating circuits," *J. Appl. Phys.*, vol. 29, pp. 1347–1357; September, 1958.

[2] L. Brillouin, "Wave Propagation in Periodic Structures," Dover Publications, Inc., New York, N. Y., p. 137; 1953.

[3] R. E. Peierls, "Quantum Theory of Solids," Oxford University Press, Oxford, England, p. 41; 1955.

[4] H. Hsu, "Backward traveling-wave parametric amplifier," in "Microwave Tubes," J. Wosnik, Ed., Academic Press, New York, N. Y., pp. 342–345; 1961.
H. Hsu and S. Wanuga, "The wide tuning range of backward traveling-wave parametric amplifiers," Proc. IRE (*Correspondence*), vol. 49, pp. 1339–1340; August, 1961.

[5] See, for example, H. Hsu, "Multiple frequency parametric devices," *Rept. of NSIA-ARDC Conf. on Molecular Electronics*, November 13–14, 1958, Washington, D. C., pp. 81–85, 1958; *Digest of Solid-State Circuits Conf.*, February 12–13, 1959, pp. 12–13, 1959.

and

$$\sum_i \beta_i = \sum_s \beta_s \tag{8}$$

or

$$\sum_i \beta_i = \sum_s \beta_s + 2\pi g, \tag{8a}$$

where E_i and E_s are the energies of the incident and scattered quasi-particles or traveling waves, β_i and β_s are the corresponding phase constants. Eq. (8) applies to continuous media, and (8a) to periodic media. Eqs. (7) and (8a) can be identified as the Bragg Law in the special case of direct scattering involving only one incident wave and one scattered wave.

It should be pointed out that the above selection rules can be calculated quantum mechanically from the interaction Hamiltonian in various collision processes. But the concept of three-dimensional parametric interactions was not evident in the formal treatment because the propagation of coherent quantized fields of quasi-particles was not believed possible earlier. With the recent development of the optical maser and the successful propagation of coherent phonons, the concept of three-dimensional parametric interaction may become important in the study of solid-state physics and quantum-field theories. Furthermore, the recent successful generation of optical harmonics utilizing the nonlinearity in the electric susceptibility of piezoelectric crystals,[6] is in fact a demonstration of the three-dimensional parametric interaction of coherent photons. Similar parametric interactions should be possible for phonons, magnons, etc. There appears to be unlimited variety in the potential development of new devices.

HSIUNG HSU
Electronics Lab.
General Electric Co.
Syracuse, N. Y.

[6] P. A. Franken, A. E. Hill, C. W. Peters and G. Weinriech, "Generation of optical harmonics," *Phys. Rev. Lett.*, vol. 7, pp. 118–119; August 15, 1961.

Comments on "Relativistic Beam-Wave Interaction"*

The theoretical results in Rowe's paper[1] relate to the interaction which takes place between an axial electric field and an electron beam. In the system he considers, it appears that if the circuit potential gradient in the z direction is finite then the forcing field falls to zero when its phase velocity is equal to the velocity of light. His equation (7) is:

$$E_{cz} = -(1 - k_0^2)\frac{\partial V_c}{\partial z}.$$

The application of this equation leads to the appearance of a factor

$$\left[1 - \frac{k_e^2}{(1 + Cb)^2}\right]$$

in the circuit force parts of (21) and the small-signal determinantal equation (28).

In many of the circuits used in high-power traveling-wave amplifiers, however, the axial forcing field is finite when the phase velocity of the fundamental is equal to the velocity of light. The same is true of the structures used in linear accelerators which are designed to give large axial fields at this phase velocity.[2]

One may consider the finite field as being due to the presence of space harmonics. Together with the fundamental these satisfy the condition $E_{\tan} = 0$ at all the conducting boundaries. To do this it is not necessary for them individually to be equal to zero. For example, if we consider a tubular structure with a section as shown in Fig. 1, then whilst $E_z = 0$ over the metal surface from A to B, a finite field may exist between B and C. To satisfy this condition E_z at this radius must be composed of a fundamental, and space harmonics of finite amplitude. These will have different radial propagation constants inside this radius and hence the field pattern along the axis will be of a different form. When the fundamental has a phase velocity equal to the velocity of light its radial propagation constant is zero and hence its amplitude is unchanged.

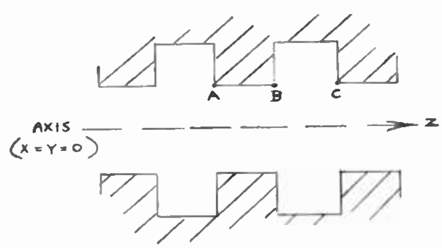

Fig. 1.

As an example of a circuit which would satisfy Rowe's theory we may consider a helix inside a conducting tube. To obtain a phase velocity equal to the velocity of light the helix must be stretched out until it becomes a straight wire. The system then becomes a coaxial line which propagates a TEM wave and E_z is zero at all points inside the circuit.

The cause of the anomaly appears to be in the fact that if V_c is a potential chosen so that E_{cz} is given by equation (7), it should not be identified with the voltage on an equivalent transmission line as is done by Pierce.[3] Since V_c tends to infinity as k_0 tends to unity, it is not really a very convenient variable. Pierce defines a different potential such that $E_{cz} = \Gamma V$, but this is not essential to the theory which can be carried through in terms of E_{cz} directly. The relativity corrections which have to be applied to the space-charge field can, for the small-signal theory, be most conveniently accounted for by working in terms of ω_q, which is the appropriately corrected effective plasma frequency.[4] When these modifications are made the small-signal, small C, determinantal equation becomes

$$\delta^2 \doteq (1 - k_e^2)^{3/2} \cdot \left\{\frac{(1 + jC\delta)^2(1 + C(b - jd))}{(-b + jd + j\delta)(1 + \frac{1}{2}C(b - jd + j\delta))}\right\} - \left\{\frac{\omega_q}{\omega \cdot C}\right\}^2$$

where

$$C^3 = \frac{E^2}{\beta^2 P} \cdot \frac{I_0}{4V_0}.$$

The value of $E^2/\beta^2 P$ used being that for the fundamental component of the electric field.

This can further be simplified by writing

$$\delta' = p_e\delta, \quad b' = p_e b, \quad d' = p_e d, \quad C' = C/p_e.$$

It then becomes

$$\delta'^2 = \left\{\frac{(1 + jC'\delta')^2(1 + C'(b' - jd'))}{(-b' + jd' + j\delta')(1 + \frac{1}{2}C'(b' - jd' + j\delta'))}\right\} - \left\{\frac{\omega_q}{\omega C'}\right\}^2.$$

Published data on the roots of this equation can be used,[5] the space-charge parameter being approximately

$$4Q'C' = (\omega_q/\omega C')^2.$$

A. J. MONK
H. J. CURNOW
S.E.R.L. Microwave Electronics Div.
Harlow
Essex, England

* Received April 23, 1962.

[1] J. E. Rowe, "Relativistic beam-wave interaction," PROC. IRE, vol. 50, pp. 170–177; February, 1962.

[2] R. B. R-Shersby-Harvie, "Travelling wave linear accelerators," *Proc. Phys. Soc.*, vol. 61, pp. 255–270; September, 1948.

[3] J. R. Pierce, "Traveling Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y., ch. 2; 1950.

[4] M. Chodorow, E. L. Ginzton, I. R. Neilson, and S. Sonkin, "Design and performance of a high-power pulsed klystron," PROC. IRE, vol. 41, pp. 1584–1602; November, 1953.

[5] C. K. Birdsall and G. R. Brewer, "Normalized Propagation Constants for a Traveling Wave Tube for Finite Values of C," Hughes Aircraft Co. Culver City, Calif., Tech. Memos. Nos. 331 and 396; October 1953 and June, 1955.

Author's Reply[6]

The relativistic device analysis developed by the author in "Relativistic Beam-Wave Interactions"[1] is a one-dimensional analysis in which only the fundamental RF charge density in the beam is considered in calculating the circuit RF voltage amplitude and phase. Harmonics of the fundamental charge density are however considered in computing the space-charge field in the force equation. Boundary conditions, of course, do not enter into a one-dimensional problem.

Fundamental to the analysis is the assumption that only one space-harmonic component of the circuit field has appreciable interaction with the beam. In a two-dimensional problem wherein one takes all space harmonics of the total field into account including the complete energy storage circuit, each individual space-harmonic field

[6] Received May 9, 1962.

component does not satisfy Maxwell's equations whereas the total field does. In this situation some space-harmonic components will be zero at $k_0=1$ whereas others may not be. The exact solution of the two-dimensional problem plus boundary conditions will indicate this.

For arbitrary periodic structures the axial component of electric field will go to zero at the upper cutoff frequency where the phase velocity is infinite and in general there will be a nonzero axial electric field at $v_p=c$.

In the present analysis the optimum interaction condition occurs when the beam is synchronized with the wave-phase velocity, and since the electron velocity is necessarily less than the velocity of light then both k_e and k_0 are less than unity and there will be a finite (though small) axial electric field. Synchronism between the beam and the wave results in the RF wave appearing as a static wave and it in turn exerts long continued forces on the electrons. As pointed out by Monk and Curnow the helical waveguide satisfies these conditions in that the axial electric field becomes vanishingly small at $v_p \rightarrow c$ and in the limit a TEM wave exists. The presence of other space harmonics may be accounted for in the circuit equations by including space-harmonic terms as additional driving terms. A specific RF structure type would then have to be considered in order to analytically specify the form of the driving terms.

The Lorentz transformation of the potential four-vector indicated in (1) assumes that three components of the vector potential are zero in the moving reference frame (purely electrostatic) which results in (7) and also the final working equations. An alternate transformation to (1) may be made in which the three components of the vector potential are assumed zero (purely electrostatic field) in the laboratory system leads to $E_z=-\partial V_c/\partial z$ which parallels the suggestion made by Monk and Curnow. Both situations are, of course, only approximate since the magnetic energy storage has been neglected. Its importance depends partly on the transverse structure dimensions relative to a free-space wavelength. If the alternate transformation is used the only change is that the factor $[1-k_e^2/(1+Cb)^2]$ is eliminated from (21) and (28). Representative efficiency calculations made using the alternate definition of E_c are shown in Fig. 2, and are compared with the earlier calculations.

In any case it is apparent that as $k_e \rightarrow 1$ the parameter C becomes small and large interaction lengths would be required. It is thus important to carry out beam bunching at low velocities and extract energy in a relativistic section.

The small-signal, *small-C*, determinantal equation given by Monk and Curnow is not restricted to small C and should contain another factor, $(1+jC\delta)^2$ in the second term, *i.e.*, $4QC(1+jC\delta)^2$ represents the space-charge field. Replacing $4QC^3$ by $(\omega_q/\omega)^2$ assumes that $\omega_q/\omega \ll 1$ which is not always valid in large C devices. In general

$$4QC^3 = \left\{ \frac{\dfrac{\omega_q}{\omega}}{1+\dfrac{\omega_q}{\omega}} \right\}^2.$$

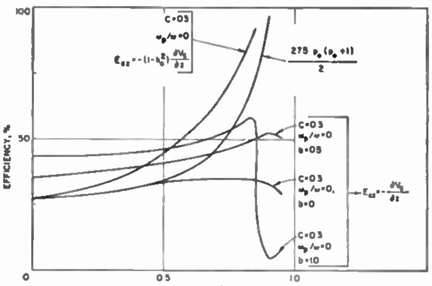

Fig. 2—Efficiency vs relativity factor.

A general relativistic analysis must include the effects of transverse fields since the Lorentz contraction applies differently to axial and transverse fields and in addition the space-charge field components are important in the two directions. The author is presently developing a general two-dimensional nonlinear analysis in which space-harmonic field components are considered.

J. E. Rowe
Electron Physics Laboratory
The University of Michigan
Ann Arbor, Mich.

Scatterer Echo Area Enhancement*

Various communications systems have been proposed in which the received signal is one scattered from an object irradiated by a distant source.[1,2] It is desired that the power available at the receiver be as large a fraction of the transmitted power as is economically feasible. One method of achieving this is by making the echo area of the scatterer large in the direction of the receiver. In the case of satellites this may be achieved by rotationally stabilized high gain scatterers or isotropic unstabilized scatterers. If it is not desired to use rotationally stabilized satellites, the weight of extremely large isotropic scatterers such as spheres ultimately becomes prohibitive. It would be desirable to further increase the apparent echo area of scatterers without a prohibitive weight increase while retaining most of the advantages of the rotationally unstabilized isotropic scatterer.

A method of enhancing the echo area of a satellite scatterer has been proposed in which a Van Atta array would have amplifiers inserted in the lines connecting the array elements. This approach has the merit of providing reliability through the redundancy of small components as well as requiring only a relatively coarse rotational stabilization of the vehicle. The realization of such an array in which the weight is not prohibitive is a formidable task in miniaturization techniques development. If a simple lightweight technique could be developed for enhancing the echo area of a scatterer, array theory could be used to develop an entire class of pseudopassive scatterers for application to problems of communications and field measurement.

Although the problem is one of realizing an enhanced bistatic scatterer (*i.e.*, a three-object system consisting of source, scatterer and receiver), it is very similar to that of the monostatic scatterer (*i.e.*, the two-object case in which scattering is directed back to the source). In the monostatic case echo area is defined as that area for which the field incident on the scatterer contains sufficient power to produce, by omnidirectional radiation, the same field as is actually backscattered to the source. Hereafter the words "echo area" will be understood to pertain to that defined above for the monostatic case.

The echo area of a scattering object having two closely spaced terminals to which is connected a load impedance, Z_L, has been formulated in terms of the two-port open circuit impedance parameters.[4] The subscript 2 pertains to the variables and parameters of the scatterer and the subscript 1 pertains to those of the radiating object. These objects are assumed to be imbedded in linear, isotropic matter.

The general relationships are specialized to the plane wave case by allowing the source to recede to an infinite distance from the scatterer. The echo area, O, of the scatterer then becomes[5]

$$\sigma = \frac{\lambda^2}{\pi}\left| (Z_{11}-Z_1) - \frac{Z_{12}^2}{Z_{22}+Z_L} \right|^2. \quad (1)$$

The open circuit impedance parameters comply with the standard two-port definitions. Z_{11} and Z_1 must be carefully distinguished. Z_{11} is the driving point impedance at the source with the scatterer terminals open circuited (*i.e.*, $Z_L=\infty$). Z_1 is the driving point impedance at the source with the scatterer removed (*i.e.*, the one-port condition). Z_{11} does not in general equal Z_1, otherwise open circuiting of the scatterer would require a zero echo area even though the two electrically isolated parts of the scatterer are capable of scattering the field.

If (1) is further specialized for the description of short dipole scatterers the term $Z_{11}-Z_1$ becomes appreciably small relative to the remaining term.[6] If attention is confined to this type of elementary scatterer and the echo area of the scatterer is normalized with respect to its echo area with its terminals shorted (*i.e.*, $Z_L=0$), the normalized echo area σ_N becomes

$$G_N \approx \left| \frac{Z_{22}}{Z_{22}+Z_L} \right|^2. \quad (2)$$

A very small dipole resonated with an inductive load may have a quality factor as high as 485 and behaves like a series RLC

* Received June 13, 1962.

[1] J. R. Pierce and R. Hompfner, "Transoceanic communications by means of satellites," PROC. IRE, vol. 47, pp. 372–380; March, 1959.

[2] J. L. Ryerson, "Passive satellite communications," PROC. IRE, vol. 48, pp. 613–619; April, 1960.

[3] R. C. Hansen "Communications satellites using arrays," PROC. IRE, vol. 49, pp. 1066–1074; June, 1961.

[4] R. F. Harrington "Small Resonant Scatterers and their use for Field Measurements," Syracuse Univ. Res. Inst., Syracuse, N. Y., Rept. No. EE492-6201TB; January, 1962.

[5] *Ibid.*, p. 7.

[6] *Ibid.*, p. 11.

circuit near the resonant frequency.[1] Although great cross section enhancement occurs, the narrow bandwidth of the scatterer makes it more suitable as a field measuring device than as a communications relay. A dimensionally resonant dipole would be more suitable as a communications signal scatterer as its equivalent RLC quality factor near resonance approximates 10. At 10 kMc this would represent a 10^3 Mc bandwidth between half-power points.

In the case of the dimensionally resonant, thin half-wave dipole Z_{22} is real and about 73 ohms. In the purely passive case equation (2) indicates that σ_N would be maximized to unity when $Z_L = 0$. Assuming that it were possible to realize a negative real load R_L that approached R_{22}, σ_N could rise to indefinitely large values. For example, if it desired to make σ_N equal 100, R_L of -65.7 ohms would be required. Assuming the equivalent lumped reactance remained constant near resonance the quality factor would have increased from 10 to 100. Under these conditions the half-power bandwidth at 10 kMc would be 100 Mc, a useful communications bandwidth.

Amplification at microwave frequencies using the negative resistance characteristics of tunnel diodes is well known. A slot transmission amplifier using this effect has been demonstrated.[7] Echo area enhancement of dimensionally resonant dipoles loaded by reflection-type tunnel-diode amplifiers has been demonstrated[8] although much work remains in achieving simple, lightweight, regulated biasing power supplies which are insensitive to frequency changes in the desired bandwidth.

In addition to the network problems involved in designing the combined diode and associated bias supply to appear as a pure negative resistance over the microwave bandwidth of interest is the problem of obtaining a long-life, lightweight power supply. The electron-emitting isotopes will provide power for long periods[9] and have promise for this application.

Joseph L. Ryerson
Rome Air Dev. Ctr.
Griffiss AFB, N. Y.

[7] M. E. Pedinoff, "A tunnel-diode slot transmission amplifier," Proc. IRE (*Correspondence*), vol. 49, pp. 1315–1316; August 1961.
[8] "Microwave Device," Directorate of Engrg., Rome Air Dev. Ctr., Griffiss AFB, N. Y., Tech. Prog. Rept., November 17, 1961, to March 16, 1962.
[9] P. Rappaport, "Electron voltaic effect in P-N junctions induced by β particle bombardment," *Phys. Rev.*, vol. 96, p. 246.

The Screening Effect of the Ionosphere*

The reflection of CW signals from the vicinity of an artificial earth satellite continues to be of interest. Studies concerning this phenomena are usually carried out at frequencies just above the critical frequency of the ionosphere where the largest effects are expected. However, it is in this exact frequency range that ionospheric screening is important. This phenomenon, although well known, is often overlooked in examining results of any CW reflection studies at diverse geographical locations. For example, the inclusion of ionospheric screening in statistical studies of CW reflection is essential in order to obtain good correlation.

* Received February 5, 1962.

In particular, WWV reflections, in which the transmitter is near Washington, D. C. and the receivers are in Columbus, Ohio, are only possible from induced ionization above the F layer maximum if the transmitter frequency is high enough to penetrate the ionosphere without being essentially "reflected" back to earth. Thus, to penetrate the ionosphere the transmitted frequency, f, must satisfy

$$f > f_c \sec \zeta$$

where

f_c is the critical frequency of the ionosphere, and

ζ is the angle of incidence, as shown in Fig. 1.

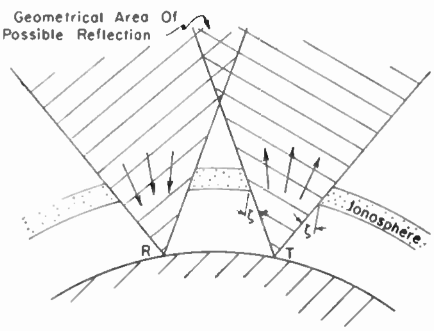

Fig. 1—Screening effect of the ionosphere relative to satellites above the F layer maximum.

For a spherical earth[1]

$$\zeta = \tan^{-1} \frac{\sin \psi/2}{1 + h/R - \cos \psi/2}$$

where

R is the radius of the earth,

ψ is the angle at the center of the earth subtended by the earth radii through the two points of interest, and

h is critical height of reflection in the ionosphere.

If f and f_c are fixed, Fig. 1 shows approximately how penetration of the ionosphere can occur from the standpoint of the transmitter location and how reflected signals at frequency f can repenetrate to the observing station.

These areas of penetration are "ionospheric holes"; their overlap allows the drawing of contours which enclose the maximum geometrical area of possible direct reflection from a satellite, at a fixed ratio of f to f_c, and a fixed satellite height (assumed to be above the F layer maximum). These contours are shown in Figs. 2–4 for three differ-

[1] "Ionospheric Radio Propagation," National Bureau of Standards, U. S. Dept. of Commerce, Washington, D. C., Circular 462, p. 68; June 25, 1948.

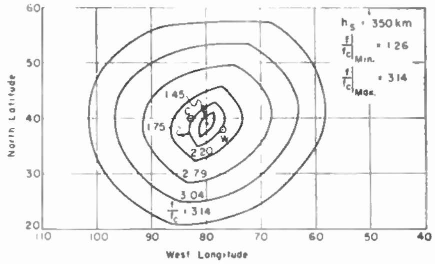

Fig. 2—Geometrical areas of possible reflection at a satellite height of 350 km. (Assumed above F_2 max.)

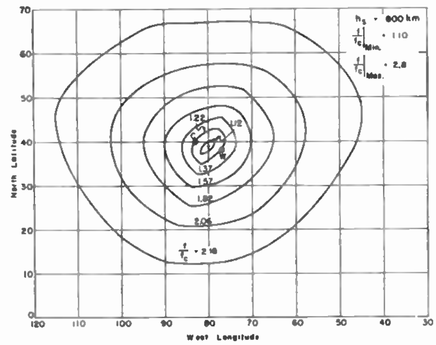

Fig. 3—Geometrical areas of possible reflection at a satellite height of 800 km.

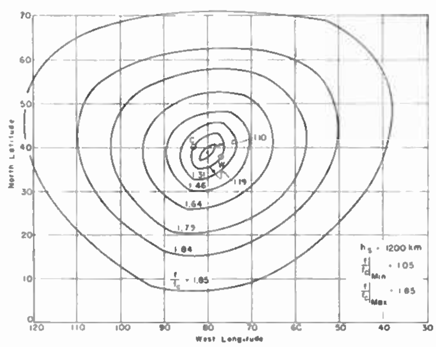

Fig. 4—Geometrical areas of possible reflection at a satellite height of 1200 km.

ent satellite heights over Washington, D. C. and Columbus, Ohio.

At each satellite height, there is a critical ratio $f/f_c|_{min}$ which is the lowest ratio for which any reflection area is possible. Also, at each height a ratio $f/f_c|_{max}$ gives the maximum geometrical area of possible reflection; higher ratios will yeild no larger area. It is emphasized that Figs. 1–4 are geometrical pictures and no allowance is made for refraction of the rays. It should be noted that the two stations which are the basis for Figs. 2–4 are close together (≈ 300 miles); however, when the transmitting station and the receiving station are separated by a larger distance $f/f_c|_{min}$ becomes larger, and, at a particular ratio, the area of possible reflection is smaller. But, most important, in Figs. 2–4, it is seen that the higher the frequency, relative to the critical frequency, the larger the area of expected reflection.

Byron C. Potts
Dept of Elec. Engrg.
The Ohio State Univ.
Columbus, Ohio

A New Criterion for Evaluating the Number of Complex Roots of an Algebraic Equation*

In many physical problems it is very interesting to be able to evaluate the number of real roots of an algebraic equation with real coefficients. The problem may be solved directly or by means of the evaluation of the number of complex root pairs of the same equation. In order to reach this last result we introduce here a new criterion, different from those presented by Budan-Fourier, Sturm, and Segre.

The gist of the present method consists in constructing a polynomial $R(x)$ having as many positive real part zeros as are the complex roots of the equation; then the number of positive real part zeros of $R(x)$ may be evaluated by applying to this polynomial either the Routh-criterion or the Hurwitz-criterion.

Let the equation be

$$P_n(x) = \sum_0^n a_i x^i = 0.$$

A first approach to the solution of the problem might be to choose as polynomial $R(x)$ the polynomial $Q_{2n}(x)$, which is the product of the complex coefficients polynomials $P_n'(x)$ and $P_n''(x)$ whose zeros are obtained by multiplying the zeros of $P_n(x)$ by j and $-j$. In fact $Q_{2n}(x)$ [Fig. 1(a) and (b)] has as many positive real part zeros as there are complex zeros of P_n, but it also has some imaginary zeros (corresponding to the real zeros of P_n), and so the criteria of Routh and of Hurwitz might not be applied in a direct way. Instead of Q_{2n} we have then to construct a polynomial $R_{2n}(x)$ whose zeros are shifted by a little negative real quantity δ from those of Q_{2n} [Fig. 1(c)].

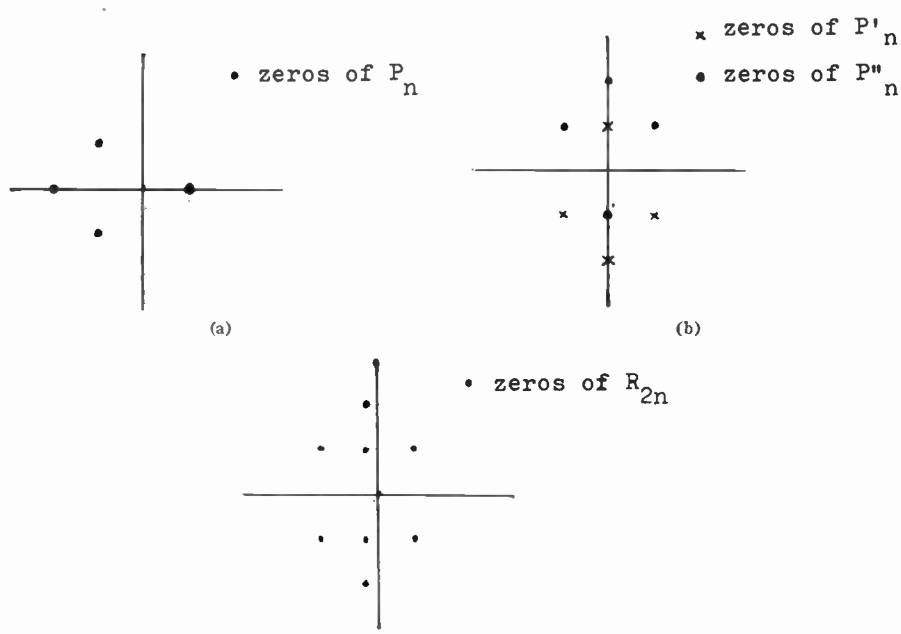

Fig. 1.

It is easy to realize that in the polynomial Q_{2n} the coefficients of the odd powers are equal to zero, and the coefficients b_{2i} of the even powers are

$$b_{2i} = \sum_{0}^{\min(i,n-i)}{}_k \; h a_{i-k} a_{i+k}$$

$$h = \begin{matrix} +1 & \text{if } k = 0 \\ 2(-1)^k & \text{if } k \neq 0 \end{matrix}$$

It is also easy to see that the coefficients c_i of the polynomial R_{2n} are given by the formulas

$$c_{2n} = b_{2n}$$

$$c_i = \left[b_i + \sum_1^{2n-i}{}_k (-1)^k \binom{i+k}{k} b_{i+k} \delta^k \right], \qquad (i \neq 2n)$$

For the purpose of the application of the criteria of Routh and of Hurwitz, we can choose a value of δ so small that its powers can be neglected. In such a way the criterion of Routh or the criterion of Hurwitz may be applied to the polynomial $R_{2n}'(x)$, very near to R_{2n} but whose coefficients c_i' are simpler functions of the coefficients a_i of the polynomial P_n.

In fact the coefficients c_i' are given by the formulas

$$\begin{aligned}
c_{2n}' &= b_{2n} = a_n^2 \\
c_{2n-1}' &= -2nb_{2n}\delta = -2na_n^2\delta \\
&\vdots \\
c_{2i}' &= b_{2i} = \sum h a_{i-k} a_{i+k} \\
c_{2i-1}' &= -2ib_{2i}\delta = -2\delta i \sum h a_{i-k} a_{i-k} \\
&\vdots \\
c_1' &= -2b_2\delta = -2\delta(a_1^2 - a_0 a_2) \\
c_0' &= b_0 = a_0^2.
\end{aligned}$$

If we will apply the Routh criterion, we may note that all the elements of the second row are multiplied by -2δ and that the elements of the third row are of the form

$$\frac{n-i+1}{n} b_{2(i-1)}.$$

Moreover the first element of the first row will always have a sign opposite to that of the first element of the second row.

Because of this, there is no need to consider the first row of Routh's set, and we may construct it starting from the elements of ths second row, divided by -2δ and from the elements of the third row multiplied by n. In such a way the criterion may be enunciated as follows.

The number of complex zeros of the polynomial with real coefficients $P_n(x) = a_i x^i$ is equal to the number of sign variations on the first column of the set:

$$\begin{array}{ccccccccc}
nA_n = \alpha_{n.1} & (n-1)A_{n-1} = \alpha_{n-1.1} & \cdots & (i+1)A_{i+1} = \alpha_{i+1.1} & iA_i = \alpha_{i.1} & \cdots & 2A_2 = \alpha_{2.1} & A_1 = \alpha_{1.1} \\
A_{n-1} = \alpha_{n.2} & 2A_{n-2} = \alpha_{n-1.2} & \cdots & (n-i)A_i = \alpha_{i+1.2} & (n-i+1)A_{i-1} = \alpha_{i.2} & \cdots & (n-1)A_1 = \alpha_{2.2} & nA_1 = \alpha_{1.2} \\
\alpha_{n.3} & \alpha_{n-1.3} & \cdots & \alpha_{i+1.3} & \alpha_{i.3} & \cdots & \alpha_{2.3} & 0 \\
\alpha_{n.4} & \alpha_{n-1.4} & \cdots & \alpha_{i+1.4} & \alpha_{i.4} & \cdots & \alpha_{2.4} & 0 \\
\vdots & \vdots & & \vdots & \vdots & & \vdots & \vdots \\
\alpha_{n.2n-1} & 0 & \cdots & 0 & 0 & \cdots & 0 & 0 \\
\alpha_{n.2n} & 0 & \cdots & 0 & 0 & \cdots & 0 & 0
\end{array}$$

where

$$A_i = \sum_0^{\min(i,n-i)}{}_k \; h a_{i-k} a_{i+k}$$

$$h = \begin{matrix} +1 & \text{if } k = 0 \\ 2(-1)^k & \text{if } k \neq 0 \end{matrix}$$

and

$$\alpha_{i.j} = \frac{-\begin{vmatrix} \alpha_{n.j-2} & \alpha_{i+1.j-2} \\ \alpha_{n.j-1} & \alpha_{i+1.j-2} \end{vmatrix}}{\alpha_{n.j-1}}.$$

To the above polynomial R_{2n}' we may also apply the Hurwitz criterion in the standard determinantal form; in this case too it is possible to introduce some simplifications by dividing the elements of the columns of the matrices to be considered by the same coefficient.

(More detailed information about the present criterion is given in a paper which is being published in the "Bollettino della Associazione Matematica Italiana.")

A. Lepschy
Istituto di Elettrotecnica
Università di Trieste, Italy

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Flow Graph Determination of the Over-All Scattering Matrix of Joined Multiports*

It is advantageous to be able to determine the over-all scattering matrix of a system of joined multiports from the known scattering matrixes of its components. Such a determination will allow us to synthesize a given multiport from simpler interconnected multiports. Conversely, we may extract the various components within a multiport.

The method will be illustrated with the simple case of a cascade of two, two ports as in Fig. 1(a). The results for this case have been solved[1] and presented elsewhere,[2] by other methods, allowing a comparison of results.

Draw S' and S'' in flow graph form as in Fig. 1(b), joining b_2' to a_1'' and b_1'' to a_2' as the output of one network will be the input of the other. The joining paths have a gain of unity indicating that the junction itself is matched. If this were not the case, we would have to add a third two port in between to account for the mismatch. Note that the inner ports are now within the composite network and the only external ports are those of a_1', b_1' and a_2'', b_2'' keeping the structure a two port. Calling a_1', b_1' the number one port, and a_2'', b_2'' the number two port, the S_{11} of the composite scattering matrix will be the total path gain from a_1' to b_1', and S_{21} will be the total path gain from a_1' to b_2'' etc. The structure in Fig. 1(b) has one loop which corresponds physically to the multiple reflection between the two networks due to their input mismatches (S_{22}' and S_{11}'') on each side of the junction. The path gains can be determined by inspection, using Mason's non-touching loop rule[3] with which we obtain

$$
\begin{aligned}
S_{11} &= S_{11}' + S_{21}'S_{11}''S_{12}'(1 - S_{22}'S_{11}'')^{-1}\\
S_{12} &= S_{12}''S_{12}'(1 - S_{22}'S_{11}'')^{-1}\\
S_{21} &= S_{21}'S_{21}''(1 - S_{22}'S_{11}'')^{-1}\\
S_{22} &= S_{22}'' + S_{12}''S_{22}'S_{21}''(1 - S_{22}'S_{11}'')^{-1}. \quad (1)
\end{aligned}
$$

The equations in (1) are identical to those obtained more laboriously elsewhere.[1,2]

With this method we shall now show how the turnstile junction[4] is electrically equivalent to two π hybrids joined at their parallel arms. With the proper choice of reference planes a π hybrid will have the scattering matrix:

$$
S_\pi = \frac{1}{\sqrt{2}}\begin{vmatrix} 0 & 0 & 1 & 1\\ 0 & 0 & -1 & 1\\ 1 & -1 & 0 & 0\\ 1 & 1 & 0 & 0 \end{vmatrix}. \qquad (2)
$$

The π hybrid is a matched four port (with one symmetry plane) in which the emergent energy from two of the ports is either in phase or 180° out of phase.

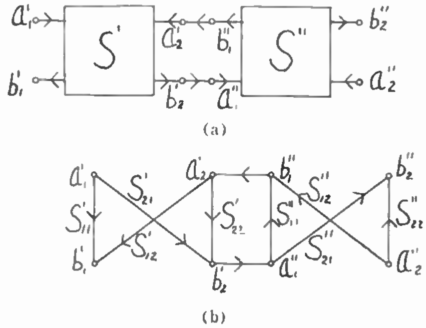

Fig. 1—(a) Cascade of two, two ports having scattering matrixes S' and S''. (b) Flow graph of the network of (a).

Fig 2(a) shows two four ports joined at one arm. The flow graph for this system with the four ports corresponding to (2) is shown in Fig. 2(b). Each vertical half of the flow graph is a representation of matrix (2). The isolation between ports is effected by the proper segregation of a's and b's. Ports 3 and 4 correspond to the series and parallel arms respectively. As all ports are matched, the junction will generate no loop and this flow graph may be evaluated very simply by path gain products.[5] Renaming ports 1′, 2′, 3′, 1″, 2″, and 3″, to 1, 2, 3, 4, 5, and 6 respectively, and defining the scattering coefficients of the composite network in the usual manner we obtain:

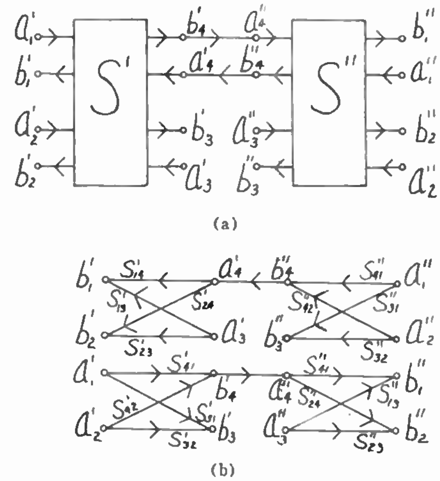

Fig. 2—(a) Two four ports joined at one arm. (b) Flow graph of two matched four ports joined at one arm.

(To change the port numbers of a scattering matrix first interchange the corresponding rows and then the corresponding columns, or vice versa.) S of (5) is identical to that of the matched turnstile junction.[4] Thus we have derived the properties of one type of matched six port and done so with less effort than would have been possible by previous methods. In addition we have shown how this six port may be synthesized from two π hybrids joined at their parallel arms.

$$
S = \begin{vmatrix}
0 & 0 & S_{13}' & S_{41}''S_{41}' & S_{42}''S_{41}' & 0\\
0 & 0 & S_{32}' & S_{41}''S_{42}' & S_{32}''S_{32}' & 0\\
S_{13}' & S_{32}' & 0 & 0 & 0 & 0\\
S_{41}''S_{41}' & S_{41}''S_{42}' & 0 & 0 & 0 & S_{13}'\\
S_{42}''S_{41}' & S_{32}''S_{32}' & 0 & 0 & 0 & S_{23}''\\
0 & 0 & 0 & S_{13} & S_{23}'' & 0
\end{vmatrix}. \qquad (3)
$$

Letting S' and $S'' = S_\pi$ of (2) we obtain:

$$
S = \begin{vmatrix}
0 & 0 & 1/\sqrt{2} & 1/2 & 1/2 & 0\\
0 & 0 & -1/\sqrt{2} & 1/2 & 1/2 & 0\\
1/\sqrt{2} & -1/\sqrt{2} & 0 & 0 & 0 & 0\\
1/2 & 1/2 & 0 & 0 & 0 & 1/\sqrt{2}\\
1/2 & 1/2 & 0 & 0 & 0 & -1/\sqrt{2}\\
0 & 0 & 0 & 1/\sqrt{2} & -1/\sqrt{2} & 0
\end{vmatrix}. \qquad (4)
$$

Renumbering ports 1, 2, 3, 4, 5, and 6 to 1, 4, 2, 5, 3, and 6, respectively, we obtain:

$$
S = \begin{vmatrix}
0 & 1/2 & 0 & 1/2 & 1/\sqrt{2} & 0\\
1/2 & 0 & 1/2 & 0 & 0 & 1/\sqrt{2}\\
0 & 1/2 & 0 & 1/2 & -1/\sqrt{2} & 0\\
1/2 & 0 & 1/2 & 0 & 0 & -1/\sqrt{2}\\
1/\sqrt{2} & 0 & -1/\sqrt{2} & 0 & 0 & 0\\
0 & 1/\sqrt{2} & 0 & -1/\sqrt{2} & 0 & 0
\end{vmatrix}. \qquad (5)
$$

Michael R. Leibowitz
1620 Ocean Ave.
Brooklyn, N. Y.

* Received February 14, 1962.

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On the Significance of Instantaneous and Short-Term Correlation Functions for a Class of Stochastic Processes*

Introduction

With attempts to develop instantaneous adaptive-learning systems a real need has arisen to define clearly the concept of instantaneous correlation functions and their physical realization. Several authors[1] have considered this problem partially. The purpose of this note is to define the notion of instantaneous correlation as a limiting value of short-term correlation and to develop appropriate expressions for such correlation functions in terms of an orthogonal series.

Short-Term Correlation Functions

In the actual measurement process, the correlation function is measured over some interval of time, let us say $2T$. Since the correlation function thus obtained is a function of this short interval,[2] $2T$, the time t and correlation interval τ, the resulting function, $\phi_{2T}(t, \tau)$ is called the short-term correlation function and is defined by

$$\phi_{2T}(t,\tau) = \frac{1}{2T}\int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi. \quad (1)$$

For a stationary time process, as $T\to\infty$, (1) becomes the usual definition of a correlation function, namely

$$\phi(\tau) = \lim_{T\to\infty} \frac{1}{2T}\int_{-T}^{T} x(\xi)x(\xi+\tau)d\xi. \quad (2)$$

This is evident by making use of the fact that for a stationary process

$$\lim_{T\to\infty} \phi_{2T}(t,\tau) = \lim_{T\to\infty} \phi_{2T}(0,\tau) \equiv \phi(\tau). \quad (3)$$

Instantaneous Correlation as a Limit of a Short-Term Correlation Function

It was shown above that as the interval T is lengthened, the correlation function of a stationary random process approaches the usual definition. What happens if T is shortened instead? In this case the correlation function is markedly dependent upon the interval length T as well as the origin in time t and the correlation variable τ. For very short intervals, T, the correlation is defined on an average only for that interval as is clearly evident. By analogy to the velocity of a particle of mass or the frequency of a wave motion, as the time of measurement is decreased the average velocity or frequency approaches what is defined as the instantaneous velocity or frequency. In the same way we define as the instantaneous correlation by allowing $T\to 0$.

Definition

The instantaneous correlation function $\phi(t, \tau)$ of a random process $x(t)$ is defined as the

$$\lim_{T\to 0} \phi_{2T}(t,\tau) = \lim_{T\to 0} \frac{1}{2T}\int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi = \phi(t,\tau). \quad (4)$$

To show that (4) is consistent with the usual notion of instantaneous quantities such as, say, velocity or frequency, the limit in (4) is now evaluated.

Let us define the function $\varphi(\xi, \tau)$:

$$\varphi(\xi\ \tau) \equiv \int x(\xi)x(\xi+\tau)d\xi \quad (5)$$

and form the difference function

$$\varphi(t+T,\tau) - \varphi(t-T,\tau) = \int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi. \quad (6)$$

Dividing both sides of (6) by $2T$ gives

$$\frac{\varphi(t+T,\tau)-\varphi(t-T,\tau)}{2T} = \frac{1}{2T}\int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi. \quad (7)$$

The difference quotient on the left side tends to a derivative as $T\to 0$ under certain conditions[3] of the function φ, *i.e.*,

$$\lim_{T\to 0}\frac{\varphi(t+T,\tau)-\varphi(t-T,\tau)}{2T} = \frac{\partial\varphi(t,\tau)}{\partial t}. \quad (8)$$

Hence

$$\frac{\partial\phi(t,\tau)}{\partial t} = \lim_{T\to 0}\frac{1}{2T}\int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi. \quad (9)$$

It is clear then that the instantaneous correlation function of the random process $x(t)$ is the derivative of the function $\varphi(t, \tau)$. This is roughly analogous to the derivative idea arising in the notion of instantaneous velocity and frequency.

To evaluate (9), note that

$$\varphi(t,\tau) = \int_0^t x(\xi)x(\xi+\tau)d\xi + \varphi(0,\tau) \quad (10)$$

where φ $(0, \tau)$ is the initial function of integration. Taking partial derivatives of both sides of (10) with respect to the time variable t gives

$$\frac{\partial\varphi(t,\tau)}{\partial t} = \frac{\partial}{\partial t}\int_0^t x(\xi)x(\xi+\tau)d\xi + \frac{\partial\varphi(0,\tau)}{\partial t}. \quad (11)$$

The last term in the right member of (11) is evidently zero. The first term of the right member of (11) is evaluated according to the rules[4] of differentiating an integral with respect to its upper limit t:

$$\frac{\partial}{\partial t}\int_0^t x(\xi)x(\xi+\tau)d\xi = x(t)x(t+\tau). \quad (12)$$

Thus it follows that

$$\frac{\partial\varphi(t,\tau)}{\partial t} = x(t)x(t+\tau) = \phi(t,\tau) \quad (13)$$

or

$$\lim_{T\to 0}\frac{1}{2T}\int_{t-T}^{t+T} x(\xi)x(\xi+\tau)d\xi = x(t)x(t+\tau). \quad (14)$$

From this it follows that the instantaneous correlation function is merely the product of the random function $x(t)$ and the function of its continuous translation $x(t+\tau)$.

Synthesis of Instantaneous Correlation Functions in Terms of a Complete Set of Orthogonal Functions

Wolf,[5] and Wolf and Dietz,[6,7] give a general theory for probing systems and correlation systems. It is shown that

$$A_n = \int_0^\infty h_n(\tau)\phi(\tau)d\tau, \quad (15)$$

where $h_n(\tau)$ are the impulse responses of a set of orthogonal filters, A_n = Wiener statistics obtained by averaging the product of the responses of the orthogonal filter to noise, and the noise. If $\{h_n(\tau)\}$ form a complete orthogonal set, a solution for $\phi(\tau)$ the correlation function is easily obtained,[5] *i.e.*,

$$\phi(\tau) = \sum_{n=1}^{\infty} A_n h_n(\tau). \quad (16)$$

From the theory given above, it follows that the instantaneous correlation function is obtained utilizing the mechanization given by Wolf[5] with $T\to 0$ where T is the averaging time and $g(t)$, the random variable, is assumed to have been on since $\tau = \to\infty$. Thus

$$A_n(T,t) = \int_{-\infty}^{\infty} \phi_T(t,\tau)h_n(\tau)d\tau \quad (17)$$

or

$$\phi_T(t,\tau) = \sum_{n=1}^{\infty} A_n(T,t)h_n(\tau). \quad (18)$$

Passing to the limit as $T\to 0$ gives

$$\phi(t,\tau) = \sum_{n=1}^{\infty} A_n(t)h_n(\tau) \quad (19)$$

where

$$A_n(t) = g(t)v_n(t) = \int_{-\infty}^{\infty} g(t)g(t+\tau)h_n(\tau)d\tau \quad (20)$$

is now a random variable.

The display of $\phi(t, \tau)$ is via a three-dimensional plot. The variable t measures the change in statistics of the correlation function as the origin of measurement changes. The variable τ measures the correlation spread. The third axis gives the correlation amplitude, $\phi(t, \tau)$.

Alfred A. Wolf
Emertron, Inc.
Emerson Research Labs.
Silver Spring, Md.

* Received February 19, 1962; revised manuscript received, March 5, 1962.

[1] D. R. Rothschild, "A note on instantaneous spectrum," Proc. IRE (*Correspondence*), vol. 49, p. 649; March, 1961.

[2] Over the interval T, the short-term correlation function $\phi_T(t, \tau)$ is defined as

$$\phi_T(t,\tau) = \frac{1}{T}\int_{t-T}^{t} x(\xi)x(\xi+\tau)d\xi. \quad (1a)$$

[3] F. Riesz and B. Sz. Nagy, "Functional Analysis," Ungar Press, ch. 1; 1957.

[4] I. S. Sokolnikoff, "Advanced Calculus," McGraw-Hill Book Co., Inc., New York, N. Y., p. 121; 1939.

[5] A. A. Wolf, "Some recent advances in the analysis and synthesis of nonlinear systems," *Trans. AIEE*, vol. 80, pp. 289–300; November, 1961.

[6] A. A. Wolf and J. H. Dietz, "A statistical theory for probing a class of linear and nonlinear systems for faults, parameter identification, and transmission characteristics," to be published.

[7] A. A. Wolf and J. H. Dietz, "Adaptive correlation techniques for signal classification, recognition, and filtering," to be published.

WWV And WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible since December 1, 1957, with respect to an improved United States Frequency Standard (USFS).[1] The corrections reported here were arrived at by means of improved measurement methods based on transmissions from the NBS stations WWVB (60 kc) and WWVL (20 kc). The values given in the table are 5-day running averages of the daily 24-hour values for the period beginning at 1800 UT of each day listed.

WWV FREQUENCY
WITH RESPECT TO U. S. FREQUENCY STANDARD

1962 June	Parts in 10^{10}†
1	−130.2
2	−130.2
3	−130.3
4	−130.3
5	−130.3
6	−130.3
7	−130.3
8	−130.2
9	−130.2
10	−130.2
11	−130.1
12	−130.1
13	−130.1
14	−130.0
15	−129.9
16	−129.8
17	−129.9
18	−129.9
19	−129.9
20	−129.9
21	−129.9
22	−129.8
23	−129.7
24	−129.8
25	−129.8
26	−129.8
27	−129.7
28	−129.6
29	−129.5
30	−129.3

† A minus sign indicates that the broadcast frequency was below nominal. The uncertainty associated with these values is $\pm 5 \times 10^{-11}$.

The time signals of WWV and WWVH are also kept in agreement with each other. Since these signals are locked to the frequency of the transmissions, a continuous departure from UT2 may occur. Corrections are determined and published by the U. S. Naval Observatory. The time signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960.

Subsequent changes were as follows:

Frequency Offset, with Reference to the USFS

January 1, 1960, −150 parts in 10^{10}
January 1, 1962, −130 parts in 10^{10}

Time Adjustments, with Reference to the Time Scale UT2

December 16, 1959, retardation, 20 milliseconds
January 1, 1961, retardation, 5 milliseconds
August 1, 1961, advancement, 50 milliseconds

Adjustments were made at 0000 UT on the foregoing dates; an advancement means that the signals were adjusted to occur at an earlier time than before.

NATIONAL BUREAU OF STANDARDS
Boulder, Colo.

* Received July 19, 1962.

[1] See "National standards of time and frequency in the United States," PROC. IRE (*Correspondence*), vol. 48, pp. 105–106; January, 1960.

The Coherent Memory Spectrum Analyzer with Loop Gain K<1*

The coherent memory spectrum analyzer is a relatively simple and flexible system for real-time spectral analysis. The basic system (Fig. 1) has been treated by several writers.[1,2,3] These writers indicate that, in unity loop gain, the square of the envelope of the system output is equivalent to scanning the spectrum of the system input once every T seconds by a filter with power transfer given by

$$\left\{\frac{\text{Sin}\,[\pi(N+1)(fT-t/T)]}{\text{Sin}\,[\pi(fT-t/T)]}\right\}^2$$

where N is the number of times a given signal component is allowed to circulate in the loop before being removed by the low-pass filter. The shape or resolution of this filter depends on N, while the center frequency depends on the time (*i.e.*, the center frequency is t/T^2 plus any multiple of $1/T$). Because the response of the system is periodic in frequency, input spectrum must be limited to a bandwidth of less than $1/T$ for unambiguous analysis.

This writer has found that the general shape of the power transfer function can be improved by removing the low-pass filter and reducing the loop gain to $K<1$. To investigate this case, let the system input be

$$i(t) = Ae^{2\pi jft}. \tag{1}$$

Let the time be represented as

$$t = nT + \tau, \qquad 0 \le \tau < T$$
$$n = 0, 1, 2, \cdots. \tag{2}$$

Then

$$i(t) = i_n(\tau) = Ae^{2\pi jf(nT+\tau)}. \tag{3}$$

After n loop circulations with loop gain K, the signal in the loop is

$$0_n(\tau) = \sum_{m=0}^{n} AK^m e^{2\pi j[f(n-m)T+f\tau+m\tau/T]}$$
$$= Ae^{2\pi jf(nT+\tau)}\frac{1-\{Ke^{2\pi j[\tau/T-fT]}\}^{n+1}}{1-Ke^{2\pi j[\tau/T-fT]}} \tag{4}$$

where the well-known relation

$$\sum_{n=0}^{N} x^n = \frac{1-x^{N+1}}{1-x} \tag{5}$$

has been used. The square of the instantaneous envelope of the signal in the loop is $|0_n(\tau)|^2$. Since the loop gain K is less than unity, the instantaneous envelope of the steady-state system output is easily obtained

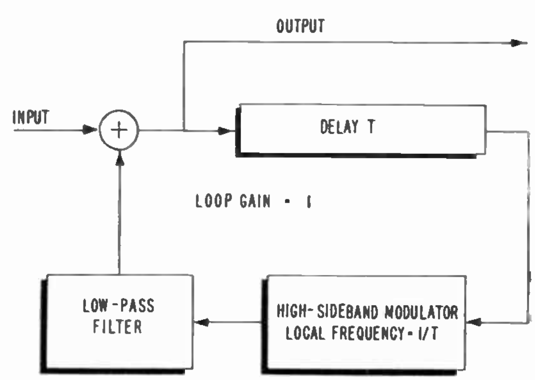

Fig. 1—Basic coherent memory spectrum analyzer.

$$\lim_{n\to\infty} |0_n(\tau)|^2$$
$$= \frac{A^2}{1+K^2-2K\cos 2\pi(\tau/T-fT)}. \tag{6}$$

Eq. (7) is the power response of the system to a single frequency input of frequency f and amplitude A. The output is periodic in $t=T$ and $f=1/T$, proportional to A^2, and peaks when

$$\tau/T - fT = 0, \quad \pm 1, \quad \pm 2, \cdots. \tag{7}$$

Therefore, the square of the envelope of the system output is equivalent to scanning the power spectrum of the input every T seconds with a filter with power transfer function given by

$$\frac{1}{1+K^2-2K\cos 2\pi\,\tau/T - fT)}.$$

* Received May 17, 1962; revised manuscript received, June 5, 1962.

[1] S. Applebaum, U. S. Patent No. 2,997,650; August 22, 1961.

[2] H. J. Bickel, "Spectrum analysis with delay line filters," 1959 IRE WESCON CONVENTION RECORD, pt. 8, pp. 59–67.

[3] J. Capon, "On the properties of an active time-variable network: the coherent memory filter," *Proc. Symp. on Active Networks and Feedback Systems*, Polytechnic Inst. of Brooklyn, Brooklyn, N. Y.; April, 1960.

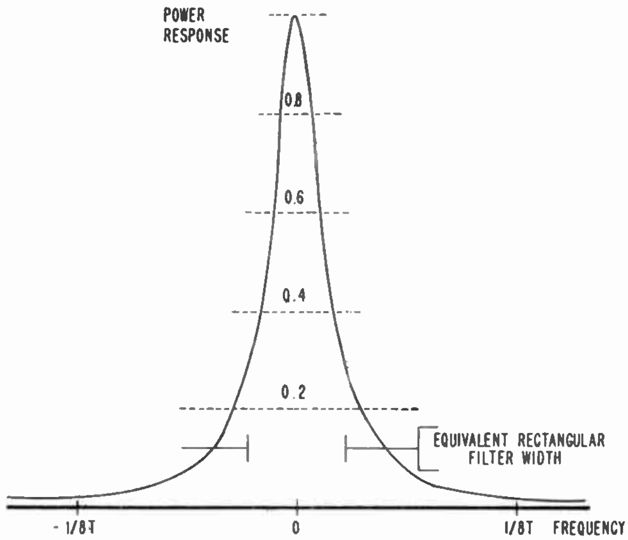

Fig. 2—Power transfer function for loop gain $K=0.9$.

The filter shape is determined by the loop gain K, while the center frequency is time-dependent and equal to t/T^2 plus any multiple of $1/T$. The filter shape is shown in Fig. 2 for the case $K=0.9$; notice the absence of sidelobes.

The equivalent rectangular bandwidth (*i.e.*, resolution) of a filter with power transfer $P(f)$ is defined as[4]

$$W_e = \frac{\left[\int_0^\infty P(f)df\right]^2}{\int_0^\infty P^2(f)df}. \tag{8}$$

Because the power transfer function of this system is periodic in frequency and because in practice the input spectrum is band-limited to the periodic interval (*i.e.*, $1/T$), (8) will be integrated over the periodic interval only. Making the substitution $2\pi(t/T-fT)=x$ yields

$$W_e = \frac{\left[\int_0^\pi [1+K^2-2K\cos x]^{-1}dx\right]^2}{2\pi T\int_0^\pi [1+K^2-2K\cos x]^{-2}dx}$$

$$= \frac{1}{2T}\left[\frac{1-K^2}{1+K^2}\right] \tag{9}$$

which results in the useful relation

$$\frac{\text{Resolution}}{\text{System Bandwidth}} = \frac{1}{2}\left[\frac{1-K^2}{1+K^2}\right]. \tag{10}$$

Adjusting the loop gain, K, theoretically yields any desired resolution, a flexibility that is one of the advantages of the analyzer—although there are practical limitations. As $K\to 1$, the system becomes sensitive to small perturbations in the loop gain, but there are no problems with oscillations because the feedback at any frequency is zero as a result of the frequency shifting by the SSB modulator. The linear dynamic range and bandpass of the memory loop will also set practical limitations on the resolution.

WINSLOW R. REMLEY
Communications Systems Ctr.
IBM Federal Systems Div.
Rockville, Md.

[4] R. B. Blackman and J. W. Tukey, "The Measurement of Power Spectra," Dover Publications, Inc., New York, N. Y., p. 19; 1958.

Shot Noise in Thin Film Transistors*

In this note the theory of shot noise in transistors is adapted to the thin film transistors consisting of metal-oxide-metal-oxide-metal or metal-oxide-metal-semiconductor structures.[1,2]

Let I_e be the emitter current, I_{cc} the leakage current between base and collector, and α_{dc} the dc current amplification factor; then the collector current $I_c=\alpha_{dc}I_e+I_{cc}$.

According to Zijlstra[3] the current in metal-oxide-metal diodes shows full shot noise at small currents and suppressed shot noise at larger currents. Consequently one would expect the same for the emitter current of the thin film transistor. The current distribution between base and collector is a partition problem and as a consequence the collector noise should consist of collected shot noise, partition noise and full shot noise of the current I_{cc}. Hence if the noise is represented by a current generator i_1 in parallel to the emitter, and a current generator i_2 in parallel to the collector (Fig. 1), and if F_e^2 is the noise suppression factor of the emitter noise,[4] one would expect

$$\overline{i_1^*i_1} = 2eF_e^2I_e\Delta f \tag{1}$$

$$\overline{i_2^*i_2} = 2eF_e^2I_e\Delta f\alpha_{dc}^2 + 2eI_e\alpha_{dc}(1-\alpha_{dc})\Delta f + 2eI_{cc}\Delta f \tag{2}$$

$$\overline{i_1^*i_2} = 2eF_e^2I_e\Delta f\alpha_{dc} \tag{3}$$

since the cross correlation between i_1 and i_2 is caused by the fact that the part $\alpha_{dc}I_e$ of the emitter current is collected.

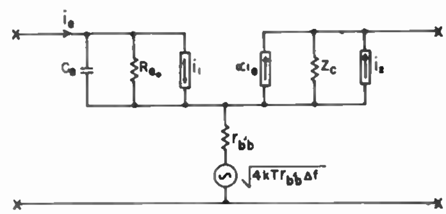

Fig. 1—Equivalent circuit of thin film transistor.

Since the emission and collection processes are *fast* processes, (1), (2) and (3) should hold for a wide range of frequencies. The high-frequency behavior of the diode is not caused by these processes, but by the emitter capacitance C_e.

The differences with the semiconductor transistor are as follows:

1) For the semiconductor transistor $F_e^2=1$.

2) I_e is not a simple exponential function of V_e, and hence the equations for $g_{e0}=\partial I_e/\partial V_e$ and $g_{ce0}=\partial I_c/\delta V_e$ no longer hold.

3) The frequency dependence of the thin film transistor must be attributed to the emitter capacitance C_e. Hence if $\alpha_0=\partial I_c/\partial I_e$, the high-frequency current amplification factor α is

$$\alpha = \frac{\alpha_0}{1+j\omega C_eR_{e0}} = \frac{\alpha_0}{1+jf/f_0} \tag{4}$$

where $R_{e0}=1/g_{e0}$, and $f_0=(2\pi C_eR_{e0})^{-1}$ is the α-cutoff frequency of the thin film transistor.

Sometimes the noise is represented by an emitter emf e and a collector current generator i (Fig. 2). In that case,

$$e = i_1Z_e;\quad i = i_2-\alpha i_1;\quad Z_e = \frac{R_{e0}}{1+jf/f_0}. \tag{5}$$

Consequently

$$\overline{e^*e} = \frac{2eF_e^2I_e\Delta fR_{e0}^2}{(1+f^2/f_0^2)} \tag{6}$$

$$\overline{i^*i} = 2eF_e^2I_e\Delta f\left[\alpha_{dc}^2 - \frac{2\alpha_{dc}\alpha_0}{(1+f^2/f_0^2)} + \frac{\alpha_0^2}{(1+f^2/f_0^2)}\right] + 2eI_e\alpha_{dc}(1-\alpha_{dc})\Delta f + 2eI_{cc}\Delta f \tag{7}$$

$$\overline{e^*i} = 2eF_e^2I_e\Delta fR_{e0}\left[\frac{\alpha_{dc}}{1-jf/f_0} - \frac{\alpha_0}{1+f^2/f_0^2}\right]. \tag{8}$$

* Received June 25, 1962. Supported by U. S. Signal Corps Contract.

[1] C. A. Mead, "Operation of tunnel-emission devices," *J. Appl. Phys.*, vol. 32, pp. 646–652; April, 1961.

[2] J. P. Spratt, R. F. Schwarz, and W. M. Kane, "Hot electrons in metal films, injection and collection," *Phys. Rev. Lett.*, vol. 6, pp. 341–342; April 1, 1961.

[3] R. J. J. Zijlstra, "Noise in currents through thin insulating layers," *Physica* (to be published).

[4] A. van der Ziel, "Noise," Prentice-Hall, Inc., New York, N. Y.; 1954.

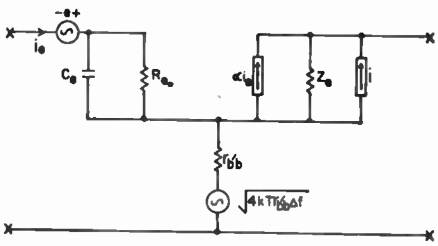

Fig. 2—Alternate equivalent circuit of thin film transistor.

When α_{dc} does not depend upon the emitter current I_e, $\alpha_0 = \alpha_{dc}$ and (7) and (8) can be simplified accordingly.

Figs. 1 and 2 incorporate also the thermal noise of the resistance $r_{b'b}$ of the base layer. The theory of the noise in thin film transistors is thus brought to the point where the standard procedures for calculating the noise resistance and the noise figure can be applied.[5]

A. VAN DER ZIEL
Dept. of Elec. Engrg.
University of Minnesota
Minneapolis, Minn.

[5] A. van der Ziel, "Fluctuation Phenomena in Semiconductors," T. Butterworth, Ltd., London, Eng.; 1959.
E. R. Chenette and A. van der Ziel, "Accurate noise measurements on transistors," IRE TRANS. ON ELECTRON DEVICES, Vol. ED-9, pp. 123–128; March, 1962. See especially the Appendix.

A Variable-Parameter Direct-Current Switching Filter*

Using contacts to switch a direct current involves two difficulties. The first is the contact arc, which generates radio-frequency interference and oxidizes or erodes the contact material. The second is the rapid rise and decay of the switched current, which may generate undesirable circuit transients as well as interference. Switching filters for arc suppression or transient control often contain reactive elements. With a linear filter, the switch must dissipate previously stored energy during either the contact make or the contact break, and it seems impossible to relieve the switch of this burden. The filter discussed here, developed for envelope shaping in a radiotelegraph transmitter,[1] eliminates the contact dissipation by using a rectifier to vary the network structure.

Consider Fig. 1, where the circuit to be switched is represented by the battery E and load resistance R_0 to the left of the terminals. The network has two circuit equivalents. The first [Fig. 2(a)] is appropriate following the instant when the switch is closed. The second [Fig. 2(b)] is appropriate following the instant when the switch is opened. Because of the rectifier, neither equivalent is an unconditionally valid representation of Fig. 1 (for all time, $t \geq 0$) unless the rectifier current is either positive for all $t \geq 0$ or zero for all $t \geq 0$.

* Received January 19, 1962.
[1] G. F. Montgomery, "Thoughts on keying filters," *QST*, vol. 45, pp. 64–65; November, 1961.

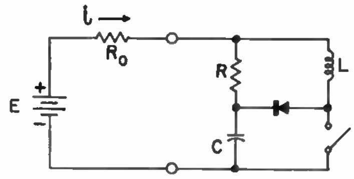

Fig. 1—Switching filter.

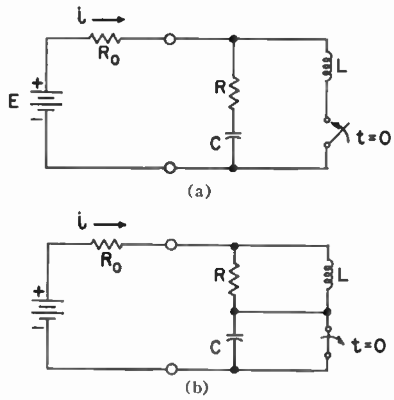

Fig. 2—Equivalent circuits.

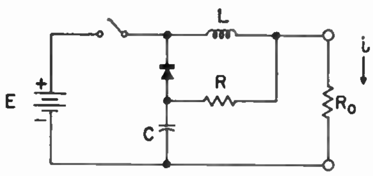

Fig. 3—Alternative filter circuit.

In Fig. 2(a), the contact current at closure is initially zero and immediately afterward rises slowly, being constrained by the inductance L. During this transient interval, the accumulated charge on capacitor C is drained through R and L in series. In Fig. 2(b), the contact voltage at break is initially zero and immediately afterward rises slowly, being constrained by the capacitance C. With the following definitions:

$$R_s = R + R_0$$
$$R_p = RR_0/(R + R_0)$$
$$2a = (1/R_sC) + (R_p/L)$$
$$\omega^2 = 1/LC, \tag{1}$$

the load current following contact closure at $t=0$ is

$$i = (E/R_0)(1 - e^{-at}\cos \mu t),$$
$$\mu^2 = (R_0/R_s)\omega^2 - a^2. \tag{2}$$

Following contact break at $t=0$, the load current is

$$i = (E/R_0)e^{-at}\cos \nu t, \quad \nu^2 = (R/R_s)\omega^2 - a^2. \tag{3}$$

The equivalent circuits are valid representations for all $t \geq 0$ if the currents (2) and (3) are either critically damped or overdamped. For $R = R_0$, $\mu^2 = \nu^2$; critical damping is obtained when $\mu^2 = \nu^2 = 0$. These conditions yield

$$a = \omega/\sqrt{2}$$
$$Q = \omega L/R_0 = 1/\omega CR_0 = \sqrt{2} \pm 1, \tag{4}$$

so that there are two L/C ratios that provide critical damping. A design for the filter proceeds as follows. Let $R = R_0$, and choose a load-current time constant $1/a$. Then,

$$L = (1 \pm \sqrt{1/2})(R_0/a)$$
$$C = (1 \mp \sqrt{1/2})(1/aR_0), \tag{5}$$

which result from eliminating ω in (4).

Fig. 3 is an alternate for the circuit of Fig. 1. These filters should be useful for reducing radio-frequency interference, for preserving relay and chopper contacts, and perhaps for reducing the dissipation in transistor and controlled-rectifier switches.

G. FRANKLIN MONTGOMERY
Natl. Bur. Standards
Washington, D. C.

On the Approximation Problem for Band-Pass Delay Lines*

The design of band-pass networks having constant delay in the pass band has been considered a difficult problem because the approximation must be done in the frequency domain directly, rather than by transformation of results already worked out for the low-pass case. This is so because the usual low-pass-to-band-pass transformation does not preserve constant delay.

Consider, however, a pole cluster having constant delay in the low-pass interval, and apply the "half transformation":

$$p = s + j. \tag{1}$$

This has the effect of translating the entire cluster in a positive direction along the $j\omega$-axis. The cluster now bears the same relationship to unit frequency that it formerly bore to zero frequency, and therefore has constant delay in the band-interval case. While we must still add the complex-conjugate pole cluster for realizability, it can be said that the approximation problem is solved so long as this addition does not seriously impair the desired constancy of delay.

Neglect of the contribution from the negative cluster has been called the "narrow-band approximation," but an examination discloses that this contribution is negligible for astonishingly large bandwidths. The design procedure is therefore notably easy:

1) Scale the low-pass pole cluster so that its interval of constant delay is one half the interval which is desired in the band-pass network.
2) Add $\pm j$ to all poles.
3) The result can now be realized as a resistively terminated ladder of series

* Received March 8, 1962.

coils and shunt capacitors. By adding appropriate zeros, it is also realizable as an all-pass network, or in one of the usual band-pass configurations.

Design Example: We are given[1] a low-pass 3-pole cluster having a mean delay of 4.576 seconds from $\omega=0$ to $\omega=1$. The delay ripples are ± 0.1 sec, and are equal:

$$P_1 = -0.7263$$

$$P_2 = -0.6148 \pm 0.9493j.$$

For a desired bandwidth of 0.5, we multiply these numbers by 0.25, and add $\pm j$. The transformed poles are

$$-0.1816 \pm j$$

$$-0.1537 \pm 1.2373j$$

$$-0.1537 \pm 0.7627j$$

having a nearly constant delay of 18.3 sec from $\omega=0.75$ to $\omega=1.25$ with nearly equal ripples of ± 0.4 sec.

Since the negative pole cluster contributes a delay of 0.328 sec at $\omega=0.75$, and 0.196 sec at $\omega=1.25$, the error is 0.132 sec, *i.e.*, this is the error in the so-called "narrow-band approximation." Since the error due to the design value of ripple is over six times as great as this, it can reasonably be claimed that the approximation is tolerably accurate.

This transformation has received previous mention,[1] but its merits were not described and may have been overlooked. It was used as a first approximation to be followed by a root-improvement procedure. The result was, of course, marvelously accurate; but a very large family of applications exist for which the inherent accuracy of the transformation is more than adequate.

As a final comment, it is worthy of mention (and of further study) that this transformation also produces amplitude pass bands having arithmetic symmetry within a very close tolerance.

Philip R. Geffe
Axel Electronics, Inc.
Jamaica, N. Y.

[1] E. Ulbrich and H. Piloty, "Über den Entwurf von Allpässen, Tiefpässen and Bandpässen mit einer im Tschebyscheffschen Sinne approximierte konstanten Gruppenlaufzeit," *A.E.U.*, vol. 14, pp. 451–467; October, 1960.

A New Frequency Demultiplier with a Tunnel Diode*

A tunnel diode is well known as a negative resistance device. It can, however, be operated as a rectifier device, since its voltage-current characteristic is asymmetric. With this characteristic, a new frequency demultiplier has been developed. The circuit is very simple and contains no dc power source. Only one tunnel diode used in this circuit operates not only as a negative resistance but also as a rectifier device.

* Received February 5, 1962; revised manuscript received, February 27, 1962.

The tunnel diode is connected in series with inductance L_0 and capacitance C_0 as shown in Fig. 1. When external forcing oscillation with frequency $\omega_0 \approx 1/\sqrt{L_0C_0}$ is applied to this circuit, the resonant current is rectified by the tunnel diode with its rectifying characteristic mentioned above and charges the capacitance C_0. With this charge, the tunnel diode is biased forward and its operating point moves toward the peak on its characteristic curve. At the Nth cycle of the external forcing oscillation (input signal), the operating point passes over the peak and jumps to the higher voltage level. That is, this circuit is effectively similar to the monostable circuit as shown in Fig. 2 and in this case the capacitance C_0 may be considered as the dc power source supplying forward-bias voltage to the tunnel diode. Thus, the operating point passes down along the characteristic curve and jumps back from the valley to the lower voltage level (starting point).

In the same manner, the capacitance is recharged and the operating point moves toward the peak again.

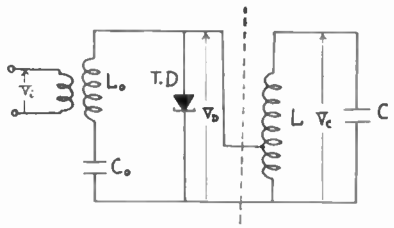

Fig. 1—The frequency demultiplier circuit

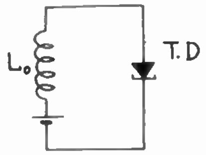

Fig. 2—The monostable circuit.

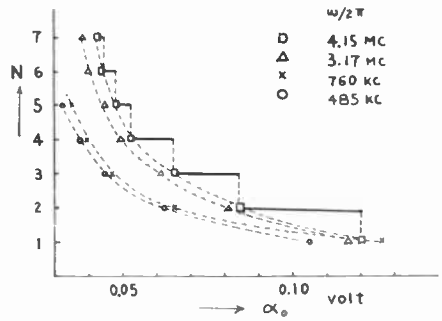

Fig. 3—The relations between the number N and the input signal amplitude.

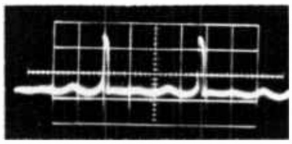
Fig. 4—The voltage waveform across the tunnel diode where $N=3$.

Fig. 3 shows the relation between the number N and the input signal amplitude α_0 for various frequencies. It is easily seen that when the input signal is large, N may be small in number and when small, large in number.

The voltage waveform across the tunnel diode is the pulse form as shown in Fig. 4. We can also obtain the sine wave by connecting LC tank circuit ($LC=N^2L_0C_0$) to the tunnel diode.

In animal auditory mechanisms, sound entering the ear makes the tympanum vibrate and this vibration is transmitted to the cochlea, in which the auditory signal is changed to the pulse train. This pulse frequency is dependent on the auditory signal intensity and frequency. The description of the complete behavior of the auditory mechanism is beyond the scope of this correspondence. However, the pulse train observed in the above circuit is experimentally found to be much analogous to that of the output signal from the cochlea. This circuit may be considered, therefore, as an electronic model of the pulse-frequency-modulator of an auditory receptor in the cochlea.

The author wishes to express his thanks to Assistant Professor Jin-ichi Nagumo for suggesting this investigation as well as for constant guidance in the course of the work.

Masamichi Shimura
Dept. of Applied Physics
Faculty of Engineering
University of Tokyo
Tokyo, Japan

Measurement of the Impurity Distribution in Diffused Layers in Germanium*

Two techniques have been developed to measure precisely the impurity distribution of diffused layers in thin, planar N-type germanium samples. A geometry is chosen so that the spatial relationships of the electric fields and voltages in the depletion layers can be mathematically related to the impurity density profile by means of Poisson's equation for one-dimensional geometry. The usual electrochemical transistor structure fulfills this requirement. Both the techniques described herein involve the measurement of punch-through voltages as a function of the position of the emitter in the diffused layer with respect to a fixed collector electrode located outside the diffused layer. The first (punch-through convergence) technique requires the measurement of both normal and inverse punch-through voltages as a function of the emitter position in the diffused layer. Such data is plotted in Fig. 1. The point at which the two curves converge is the location of the edge of the diffused layers. Fig. 1 also illustrates the precision of the method by show-

* Received February 6, 1962.

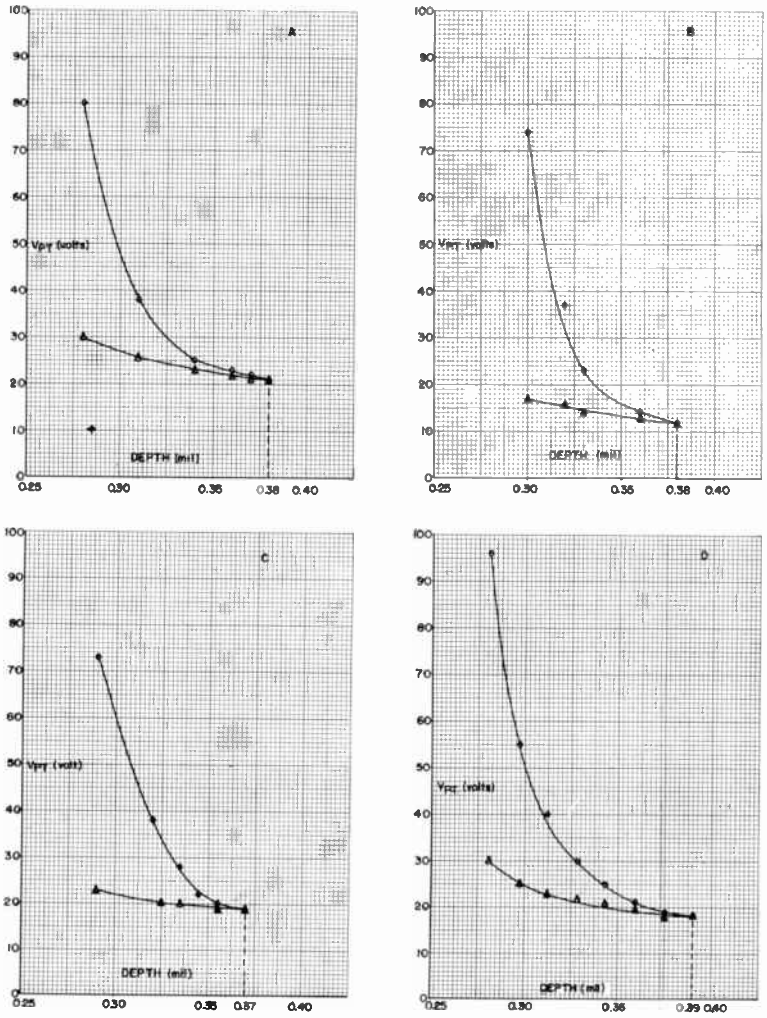

Fig. 1.

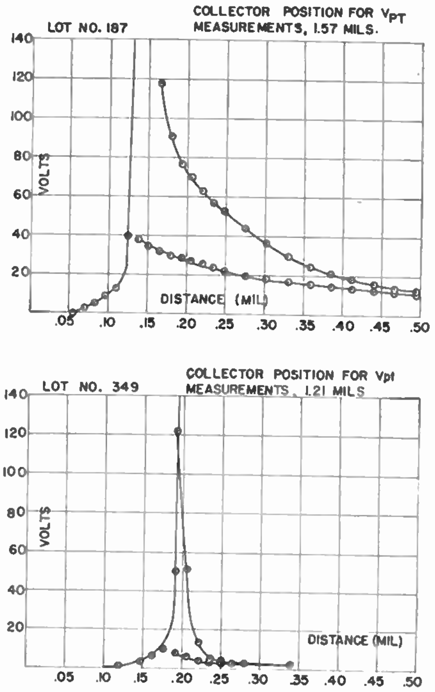

Fig. 2.

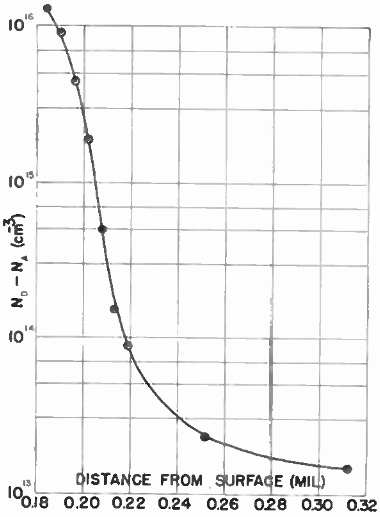

Fig. 3.

ing the data collected by two persons, on two completely separate sets of equipment, on four samples from the same diffused lot.

While it is not necessary to place the collector electrode at any particular location, the value of the data is increased if the collector is placed where it would be in the transistor. The upper punch-through voltage curve then relates the position of the edge of the collector-depletion layer to the applied collector voltage. By this means one can determine quantitatively the variation in base width and collector-depletion-layer width with the applied collector voltage.

Fig. 2 shows clearly the advantage this technique has over the diode-breakdown voltage technique. Diode-breakdown voltage measurements have been the most useful method for evaluation and control of diffused layers. For a number of reasons diode-breakdown voltage measurements are not very useful near the edge of the diffused layer. The punch-through voltage curves in Fig. 2 show the difference in depth of two diffused layers to be about 0.25 mil, while the breakdown voltage data would seem to show a difference of about 0.06 mil. Furthermore, the diode-breakdown voltage measurements would indicate that the bottom sample has a deeper diffused layer while the punch-through convergence technique shows that the top sample actually has the deeper diffused layer. The time required to collect data for curves such as shown in Fig. 1 is about 45 minutes.

The second technique to be discussed involves the calculation of the impurity distribution from measurements of the punch-through voltage from the fixed collector to the nonfixed emitter electrode as a function of the position of the latter electrode. Poisson's equation relates the impurity density to the gradient of the field:

$$N = \frac{\varepsilon}{q}\frac{dE}{dx}.$$

The value of dE/dx can be determined from successive measurements of punch-through voltage V and base width W, by the approximation:

$$\frac{dE}{dx} \approx \frac{\Delta V}{W \Delta W}.$$

Fig. 3 shows the result of such work on an N-type germanium blank into which a layer of arsenic has been diffused.

In the second technique described the depletion layer at the junction at which punch-through is detected has been neglected.

The etching, plating, and measuring equipments, and the techniques developed for electrochemical transistor technology made the approach described in this letter practical. The results of these approaches, in turn, have improved the control of the semiconductor material involved in electrochemical technology.

The authors wish to acknowledge the fact that Dr. G. L. Lang, presently at Carnegie Institute of Technology, made a major contribution in the early development of the theory of this method in 1958, and that L. Pomante, formerly of their laboratory, made some of the measurements.

E. S. Schlegel
D. P. Sanders
Lansdale Div.
Philco Corp.
Lansdale, Pa.

Phase Shift in Ferrites at High Power*

Although the high power absorption effects observed in ferrites are twofold in nature (thermal and nonlinear), the phase shift is expected to exhibit only thermal effects. That the phase shift characteristics are temperature sensitive is known from the strong dependence of magnetization on temperature and the measurements of Martin,[1] and Geiszler and Henschke.[2] The absence of nonlinear behavior can be demonstrated once the thermal characteristic effects are eliminated.

This communication describes a technique devised at this laboratory (as part of an experimental investigation of ferrite materials) for eliminating or minimizing heating effects in the measurement of the phase shift.

Measurements were carried out in regular and reduced S-band waveguides for longitudinally and transversally magnetized ferrite slabs. The results indicate that the phase change is practically independent of power level (up to the power measured).

A block diagram of the equipment used is given in Fig. 1. The high power source was a 4J39 magnetron with an operating frequency of 3525 Mc and an output peak power of 750 kw at a 0.5-μsec pulse width and a repetition rate of 1000 pps. The power was fed into a power splitter leading to two arms, one containing the ferrite test sample, the other a variable calibrated Riblet-type phase shifter and power divider. The phase shift of the standard phase shifter, consisting of a short-slot hybrid and variable short, is a linear function of the position of the short, which was measured with an Ames gage. The power divider was constructed of three short-slot hybrids and an adjustable short.

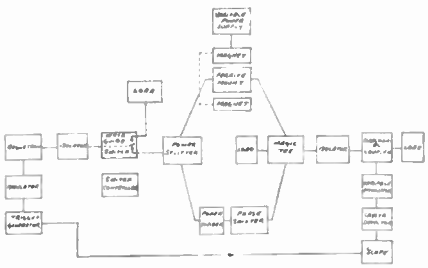

Fig. 1—Block diagram of equipment used for high-power shift measurements of ferrite materials.

The introduction of the power divider in one of the arms provided a means of equalizing the amplitudes of the power in the two arms of the comparison circuit. In the process of varying the power by the power divider, an additional phase shift is also introduced. Thus, the measured ferrite phase shift is the sum of the phase shifts of the Riblet phase shifter and power divider. The signals from both arms were then recombined in a magic-T and detected by a crystal. The detected signal was applied to an oscilloscope. When the detected signal is a minimum the two arms of the bridge are in phase balance.

The procedure for the phase shift measurement was to adjust the Riblet phase shifter and power divider for a minimum deflection on the oscilloscope, when the sample was present, without and with an applied magnetic field. The phase difference was then determined from a change in setting of the phase shifter and power divider. Due to the high sensitivity of the setup, the phase measurements were accurate to within two degrees.

Precautions against heating effects were taken through the introduction of a high-speed waveguide switch past the magnetron and by means of a stepwise determination of the minimum (see below). A control circuit was designed to adjust the switching time interval (the time in which the shutter opens completely and returns to its original position). During the switching cycle the magnetron power applied to the ferrite is swept in amplitude from 0- to 375-kw peak power and then back to 0 power. By changing the voltage and/or the resistance in the control circuit, the time interval is adjustable from 20 to 100 msec. Once a minimum was found, the power was turned off to allow the ferrite to cool. The power was then reapplied and the short adjusted slightly for a minimum. This process was repeated till no further adjustment was necessary.

It will be shown that, since only one switching cycle was necessary for the final determination of the phase shift, the rise in temperature was negligible. The energy incident on the ferrite during one cycle is

$$W = (2/\pi)NP_p u,$$

W = energy in joules
N = number of pulses = 20
P_p = peak power = 375×10^3 w
u = pulse width = 0.5×10^{-6} sec

The factor $2/\pi$ arises from the fact that the sweep in power from 0 to maximum and back to zero has a cosine dependence. Assuming that, under the worst condition, all the energy is absorbed, the rise in temperature is

$T = W/(cm)$
T = temperature in °C
c = approximate specific heat of ferrites = 708 joules/kg/°C
m = mass in kg.

Thus, Ferrite Motorola Y-188 with $m = 320$ gm and Trans-Tech 469 with $m = 80$ gm had, at most, an increase of 0.042°C and 0.011°C, respectively.

Low-power measurements were also performed for comparison. The phase changes of one representative sample (out of twelve) as a function of magnetic field is given in Fig. 2. The curves clearly show that hardly any phase distortion is introduced with the increase of peak power.

Martin[1] and Brown[3] obtained similar results. Martin noticed in ferrite slabs at X band that at a repetition rate of 1 pps (no heating effect), throughout the range of high

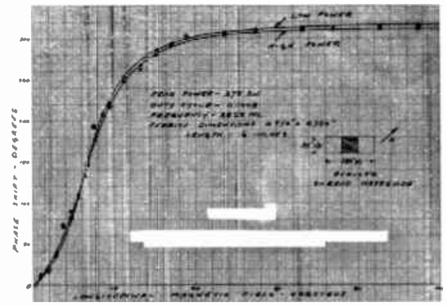

Fig. 2—Phase shift of Ferrite Motorola Y-188 at high and low power vs longitudinal magnetic field.

power used (up to 120 kw), the phase changed at most a few degrees over that observed at low power. In an X-band limiter, Brown found that the phase shift was constant over a 30-db range of input power.

K. Kalikstein
N. Cooper
J. Troy
Radar and Microwave Section
Material Laboratory,
New York Naval Shipyard
Brooklyn, N. Y.

* Received February 6, 1962; revised manuscript received, February 13, 1962.

[1] R. L. Martin, "High power effects in ferrite slabs at X-band," *J. Appl. Phys.*, vol. 30, pp. 159–160; April, 1959.

[2] T. D. Geiszler and R. A. Henschke, "Broad band reciprocal phase shifters," *J. Appl. Phys.*, vol. 31, pp. 174–175; May, 1960.

[3] J. Brown, "On phase distortion of ferro-magnetic limiter," Proc. IRE, vol. 49, pp. 362; January, 1961.

A New Class of Distributed RC Ladder Networks*

A class of solutions to the linear second order differential equation

$$v'' - (r'/r)v' - srcv = 0, \qquad (1)$$

in which s is a constant, r and c are functions of the independent variable x, and the prime denotes differentiation with respect to x, has been discussed by Kazansky[1] and Jacobs.[2] Sugai[3] also derives this class from a method of Hildebrand.[4] It is defined by making constant the logarithmic derivative of r/c expressed as a function of

$$z = \int_0^x \sqrt{rc}\, dx.$$

That is, let

$$G(z) = d(\ln \sqrt{r/c})/dz = 2m \qquad (2)$$

where m is a real constant.

Described here is a broader class of networks which encompasses that discussed above and which provides network behavior that cannot be achieved with the $G(z) = 2m$ class.

* Received February 2, 1962. This material constitutes a portion of the work in progress on a Doctoral Dissertation at the Polytechnic Institute of Brooklyn.

[1] B. G. Kazansky, "Outline of a theory of non-uniform transmission lines," *Proc. IEE*, pt. C, vol. 105, pp. 126–138; March, 1958.

[2] I. Jacobs, "A generalization of the exponential transmission line," Proc. IRE, vol. 47, pp. 97–98; January, 1959.

[3] I. Sugai, "A generalized Hildebrand's method for nonuniform transmission lines," Proc. IRE (*Correspondence*), vol. 49, p. 1944; December, 1961.

[4] F. B. Hildebrand, "Advanced Calculus for Engineers," Prentice Hall, Inc., New York, N. Y., p. 50; 1948.

The basis for this class is the Liouville Transformation[5] in which the independent and dependent variables are simultaneously transformed according to

$$z = \int_0^x \sqrt{rc}\, dx \qquad (3)$$

and

$$u = (c/r)^{1/4} v. \qquad (4)$$

The transformed differential equation is now in the Liouville Normal Form,

$$\ddot{u} - [s - F(z)]u = 0, \qquad (5)$$

where

$$F(z) = (1/rc)\{(1/4)[(c/r)'/(c/r)](1/2)[(rc)'/(rc)] - (1/4)[(c/r)''/(c/r)] + (3/16)[(c/r)'/(c/r)]^2\}. \qquad (6)$$

This relation can be written in many forms but it is particularly useful to express it in terms of the function $G(x)$. Note $G(x)$ is simply related to $G(z)$ by virtue of (3) and its derivative,

$$dz/dx = \sqrt{rc}.$$

Of course $F(z)$ can also be expressed as a function of x, thus,

$$F(x) = (1/4)\{2[(r/c)/(r/c)'](G^2)' - (G^2)\}. \quad (7)$$

If, now, the function $F(x)$ is equated to a constant then the Liouville Normal Form of the differential equation becomes readily solvable since it would have constant coefficients. This operation specifies a relation between r and c which defines this new class of networks. That is,

$$F(x) = -m^2$$

or

$$2[(r/c)/(r/c)'](G^2)' - (G^2) = -4m^2. \quad (8)$$

This last equation is a first order linear differential equation in (G^2) with variable coefficients. Its solution may be obtained by the method of variation of parameters as

$$G^2 = 4m^2 + K\sqrt{r/c} \qquad (9)$$

where K is a constant of integration. This equation is the criterion for this new class of solutions.

It is seen from (9) that when K is chosen as zero (9) yields the criterion (2) for the earlier class which has been referred to as a "Generalized Exponential Class."[2]

The class introduced here and defined by (9) might be called a "Generalized Hyperbolic Class" (G.H.C.) since the solutions of (5) with $F(z) = -m^2$ are hyperbolic in nature.

The G.H.C. criterion (9) has four cases according to

a) $m^2=0$, $K=0$
b) $m^2\neq 0$, $K=0$
c) $m^2=0$, $K\neq 0$
d) $m^2\neq 0$, $K\neq 0$.

Case a) is the simplest and corresponds to the "Proportional" Network[6] in which r/c is independent of x. The second case b), as was mentioned above, is the "Generalized Exponential Class." Cases c) and d) are the two new solutions introduced by the G.H.C. and the performance of networks of these types differs from that of a) and b). This performance in terms of network parameters and graphical determination of their zeros has been studied in detail. The relation between r and c for the two new G.H.C. cases has also been worked out and specific practical examples formulated. It is hoped that these details will be published at some later date.

Other possibilities besides letting $F(z) = -m^2$ exist which, by taking advantage of differential equations whose solutions are known, lead to still other classes of networks. For example the Liouville Normal Form (5) can be made into a Bessel equation by the choice

$$F(z) = -m^2 - [a^2 - (1/4)]/z^2 \quad (10)$$

where a = constant.

This leads to the following as a criterion for this "Generalized Bessel Class" of networks:

$$G^2 = 4m^2 + K\sqrt{r/c} - (4a^2 - 1)\sqrt{r/c} \cdot \int_0^x \frac{(\sqrt{r/c})'/(r/c)}{\left[\int_0^x \sqrt{rc}\, dx\right]^2}\, dx. \qquad (11)$$

It is seen from this criterion that the G.H.C. criterion (9) is a special case of (11) wherein $a = \frac{1}{2}$.

M. J. Hellstrom
Television-Radio Div.
Westinghouse Elec. Corp.
Metuchen, N. J.

[5] E. Kamke, "Differentialgleichungen, Lösungsmethoden und Lösungen," Edwards Bros., Inc., Ann Arbor, Mich., p. 261; 1945.

[6] M. J. Hellstrom, "Symmetrical RC distributed networks," Proc. IRE, vol. 50, pp. 97–98; January, 1962.

The Magnetic Monopole and the Principle of Parity*

In a recent letter, Leen asked for a logical argument for the nonexistence of an isolated magnetic pole.[1] How about the following reasoning using the well-known principle of parity: Let us assume a current i. Around this current we have a circular magnetic field H. We test this magnetic field by our (hypothetical) isolated magnetic pole $+p$ fixing this pole by, say, a string on the conductor carrying the current i (see Fig. 1). The magnetic pole rotates around i under the action of the force F induced by the magnetic field H—let us assume *counterclockwise*. Now we quote the principle of parity stating that a possible experiment seen in a mirror is a possible experiment too. Looking in a mirror, however, we see our pole rotating *clockwise* around i.

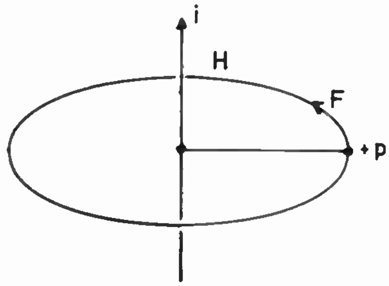

Fig. 1.

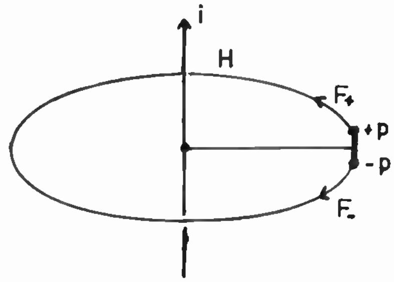

Fig. 2.

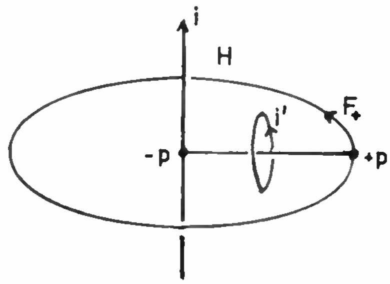

Fig. 3.

This contradiction solves only if we assume that an isolated magnetic pole $+p$ must not exist, but that a magnetic dipole $(+p, -p)$ does. Fig. 2 shows that for a dipole, the forces F_+ on $+p$ and F_- on $-p$ cancel; the dipole does not move.

But let us fix the dipole $(+p, -p)$ so that the pole $-p$ falls within the path of the current i (see Fig. 3). We again observe the counterclockwise rotation of $+p$ around i. To avoid the former contradiction against the principle of parity we must now consider the dipole (like Ampère) as produced by a circular current i'. If we look in the mirror now, this current reverses its direction, the magnetic dipole reverses too, and we see a likewise possible experiment.

Consequently the principle of parity implies the nonexistence of the magnetic monopole and the "axial" character of the magnetic dipole.

Peter Schnupp
Max Planck-Institut für Physik
und Astrophysik
Munich, Germany

* Received February 6, 1962.

[1] M. W. Leen, "Physical basis for electromagnetic theory," Proc. IRE (*Correspondence*), vol. 50, p. 90; January, 1962.

Author's Comment[2]

The comments of Schnupp are certainly pertinent to an examination of the minimum number of postulates required for the derivation of Maxwell's equations. However, the author does not agree that the principle of parity is useful for that purpose.

Consider a current flowing through a line element $d\mathbf{l}$, in a positive direction along the Z axis of a right-handed system of coordinates. Then the magnetic field at a point P_2, a distance r from the Z axis, has the direction and magnitude given by

$$\mathbf{B} = \left(\frac{\mu_0 i}{4\pi}\right)\frac{d\mathbf{l}\times\mathbf{r}}{r^3}.$$

The direction of $\mathbf{B}$ is determined by the conventional direction of a vector product in a right-handed coordinate system as shown in the solid lines in Fig. 4. In addition, if we place the magnetic dipole N-S at P_2 with its axis inclined to the direction of $\mathbf{B}$, it will be subject to a clockwise torque (looking from P_2 to P_1) tending to align it with $\mathbf{B}$.

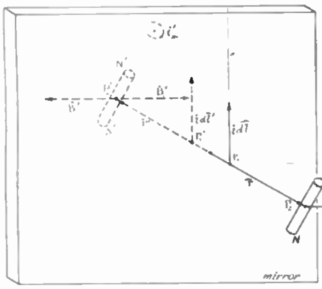

Fig. 4.

Now if we apply the principle of parity as quoted by Schnupp, and construct the mirror image as shown in dotted lines, it is seen that the current element $i\mathbf{dl}'$ and the distance $\mathbf{r}'$ form, with $\mathbf{B}_s'$ (the mirror image of $\mathbf{B}$) a left handed system, whereas the true field due to $i\mathbf{dl}'$ is

$$\mathbf{B}_L' = \left(\frac{\mu_0 i}{4\pi}\right)\frac{d\mathbf{l}'\times\mathbf{r}'}{r^3}.$$

Now if we postulate a (south) magnetic monopole at P_2, the force it experiences is in the direction of $\mathbf{B}$ while its image at P_2' would experience a force in the direction of $\mathbf{B}_L'$. This is essentially in agreement with Schnupp's argument up to this point. However, the apparent contradiction between the results of the real experiment at P_2' and the image at P_2' of the real experiment at P_2 can be resolved without concluding that monopoles do not exist.

Note that the direction of rotation of a small current loop, i_m, is reversed by imaging in the mirror and hence the dipole N-S is reversed by imaging to take the position N'-S'. Thus we can resolve the dilemma by assuming that reflection in a mirror reverses the polarity of a monopole (if one were to exist) as well as a dipole.

To summarize, the derivation of Maxwell's equations in the previously quoted articles[3,4] requires in addition to the assumption of the linearity of electromagnetic phenomena, the assumption of the nonexistence of magnetic monopoles. The principle of parity cannot be used as a substitute.

M. W. Leen
6 Maple Row
Bethel, Conn.

[2] Received February 20, 1962.

[3] P. Clavier, "Electromagnetic theory from a mathematical viewpoint," Proc. IRE (*Correspondence*), vol. 48, pp. 1494–1495; August, 1960.

[4] I. Sugai, "Vectors for waves and electrons," Proc. IRE (*Correspondence*), vol. 49, pp. 628–629; March, 1961.

The Effect of Mutual Inductance Upon Tunnel Diode Locked Pair Switching*

In previous discussions of the design and operation of the tunnel diode locked pair logic circuit, it has been assumed that mutual inductance between the loops of the two diodes could be neglected.[1–4] This note presents some results of a computer analysis which indicate that significant improvement in high-speed switching may be obtained by adding mutual coupling.

The circuit to be considered is shown in Fig. 1. For digital computer analysis the tunnel diodes have been represented by an analytic volt-ampere relation which closely approximates a germanium unit with 10 ma of peak current. A piecewise linear diode approximation was used on an analog computer. The coupled inductances of the two loops are represented by their "tee" equivalent.

If the load resistance R is large enough so the rise time into the load

$$\tau = M/R \tag{1}$$

may be neglected, the switching transient depends only upon the difference between total and mutual inductance or L-M. This effect is shown in Fig. 2. If L-M is made too large and highest speed operation is desired, the circuit does not switch but oscillates with both diodes in phase so no power is coupled to the output. The condition for this oscillation is

$$L > \frac{R_S R_N C}{1-k} \tag{2}$$

where $k = M/L$ is the coefficient of coupling. In Fig. 3 the maximum inductance for which switching occurs is plotted vs $1/1-k$. The supply waveforms are step functions which correspond to operation at the maximum frequency. The relation is linear and the slope implies $R_N = 35$ ohms which is not unreasonable for the "average" negative resistance of a 10 ma diode. Examination of the switching waveforms for several combinations of circuit parameters indicates that switching time is essentially unimpaired if

$$L - M < \tfrac{1}{2} R_S R_N C. \tag{3}$$

Circuits have been constructed with interlocked supply resistors as shown in Fig. 4. Both diode loops surround approximately the same area so the majority of the magnetic flux couples both loops. The inductance and coupling coefficient of this fixture have been measured as about 1.5 nanohenry

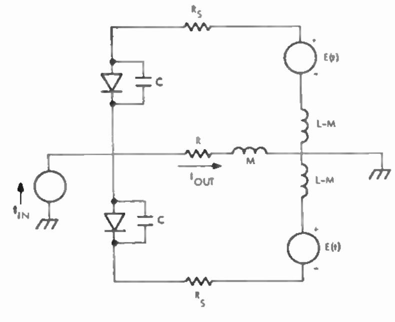

Fig. 1—Tunnel diode locked pair including mutual inductance.

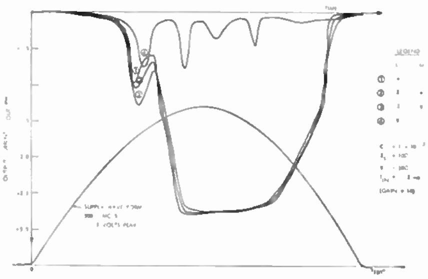

Fig. 2—Switching transient for various values of inductance L and mutual inductance M with a sinusoidal supply waveform. (Analog computer solution.)

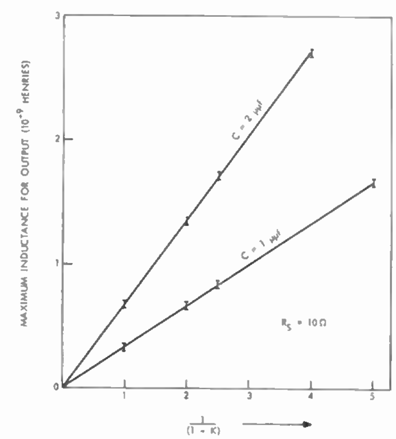

Fig. 3—Maximum inductance for output vs $1/1-k$. Circuit parameters are as in Fig. 2. Supply waveforms are 0.2 volt steps. (Digital computer solution.)

* Received June 18, 1962.

[1] E. Goto, *et al.*, "Esaki diode high speed logical circuits," IRE Transactions on Electronic Computers, vol. EC-9, pp. 25–29; March, 1960.

[2] W. F. Chow, "Tunnel diode digital circuitry," IRE Transactions on Electronic Computers, vol. EC-9, pp. 295–301; September, 1960.

[3] J. J. Gibson, *et al.*, "Tunnel diode balanced pair switching characteristics," *Digest of 1962 Internat'l. Solid-State Circuits Conf.*, Philadelphia, Pa., February, 1962, Lewis Winner, New York, N. Y., pp. 54–55; 1962.

[4] L. Esaki, "Characterization of tunnel diode performance in terms of device figure of merit and circuit time constant," *IBM J. Res. & Dev.*, vol. 6, pp. 170–178; April, 1962.

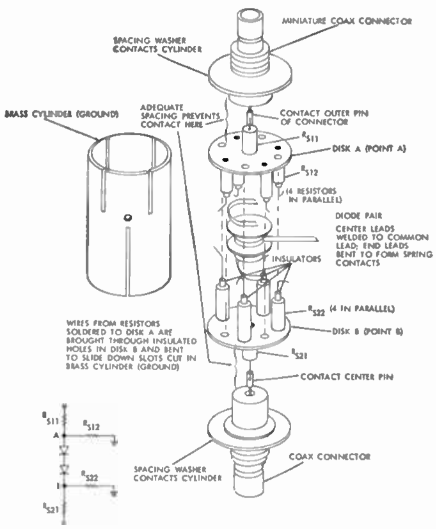

Fig. 4—Exploded view of diode mount. (Not to scale.)

and 0.7, respectively. Using 10 ma diodes with about 3 $\mu\mu$f of capacitance, a current gain of 50 has been measured at 200 mc where current gain is the ratio of maximum output current into the load resistor divided by one half of the input current necessary to switch from one output polarity to the other.

A. C. Scott
Dept. of Elec. Engrg.
University of Wisconsin
Madison, Wis.
W. J. Dunnet
R. W. Taylor
Electronics Systems
Sylvania Products, Inc.
Needham, Mass.

Designing a Phase-Locked Loop as a Doppler Tracker*

The transfer function and low-pass filter of a phase-locked loop may be precisely determined to give adequate transient response with minimum noise bandwidth when the loop is used as a tracking filter. This note gives a straightforward design procedure.

The well-known linear equivalent of a phase-locked loop is shown in Fig. 1, and the root locus is shown in Fig. 2.[1] The open-loop transfer function is composed of the two poles and the zero on the real axis, and we desire to place the closed-loop poles so that the absolute magnitude of the phase error $e(t)$ is less than approximately 0.6 rad when tracking a given Doppler signal. (This will result in a near optimum choice between noise bandwidth and threshold.)

* Received February 7, 1962. This research was supported by AF Space Systems Division Contract No. AF 04(647)-829.

[1] C. S. Weaver, "A new approach to the linear design and analysis of phase-locked loops," IRE Trans. on Space Electronics and Telemetry, vol. SET-5, pp. 166–178; December, 1959.

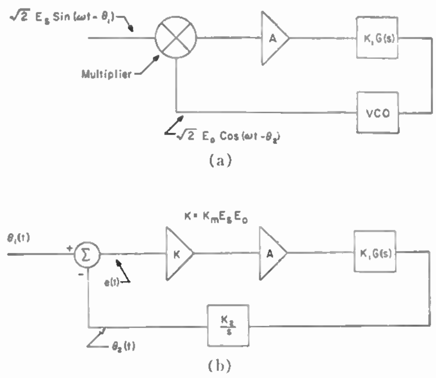

Fig. 1—(a) A phase-locked loop. (b) Linear equivalent.

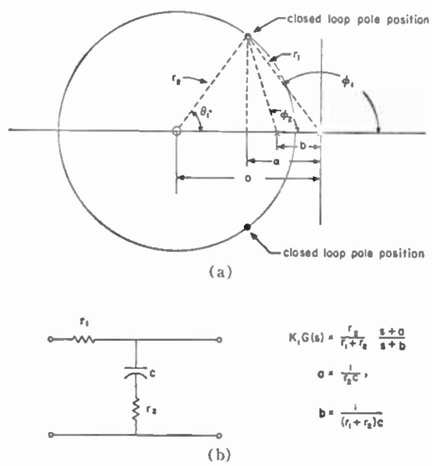

Fig. 2—(a) The root locus when $KG(s)$ is as shown in (b). (b) The loop filter.

From Fig. 2(a) we see that the transfer function from $\theta_1(t)$ to $e(t)$ is

$$\frac{E(S)}{\theta_1(S)} = \frac{1}{1 + KAK_1K_2\dfrac{S+a}{S+b}\dfrac{1}{S}} = \frac{S(S+b)}{S(S+b) + K_{OL}(S+a)} \tag{1}$$

where K_{OL}, the open-loop gain, is

$$K_{OL} = KAK_1K_2.$$

Over most of the range of a typical Doppler curve the slope is very nearly a straight line, or a ramp in frequency. Let this slope be D cps. Because phase is the integral of frequency,

$$\theta_1(S) = \frac{2\pi D}{S^2} \times \frac{1}{S} = \frac{2\pi D}{S^3}\cdot \tag{2}$$

and

$$E(S) = \frac{2\pi D}{S^3}\,\frac{S(S+b)}{S(S+b)+K_{OL}(S+a)} = \frac{2\pi D}{S^2}\,\frac{S+b}{S(S+b)+K_{OL}(S+a)}\cdot \tag{3}$$

From a table of Laplace transforms we see that $e(t)$ eventually settles down into a ramp plus a constant given by

$$e(t) = \frac{2\pi D}{r_1^2}\left[bt + 1 - \frac{2b\alpha}{r_1^2}\right]. \tag{4}$$

For a large tracking range (compared to the loop bandwidth) b should be small compared to r_1. Then $e(t)$ is approximately

$$e(t) \cong \frac{2\pi D}{r_1^2}[bt+1] < 0.6. \tag{5}$$

If ϕ_1 is 135°, there will be an overshoot of less than 10 per cent. Then

$$r_1 > \sqrt{\frac{2\pi D}{(0.6)}(bt+1)} \tag{6}$$

will guarantee that we are in the low-phase error region.

The root locus condition requires that

$$180° = \theta_1 - \phi_1 - \phi_2. \tag{7}$$

Then the design procedure is as follows: r_1 is found from inequality (6), and a is adjusted to satisfy (7). The open-loop gain is

$$K_{OL} = \frac{r_1r_3}{r_2}\cdot$$

If possible the VCO frequency with no phase error should be set at the center of the Doppler curve (zero frequency shift). Then t in inequality (6) would be no larger than the time, t_0, it takes the frequency ramp to go from the higher frequency limit of the Doppler curve to the center of the curve. If the VCO must be set at some other point, the largest time of inequality (6) will be $2t_0$.

Since the noise bandwidth is proportional to r_1, it is obvious that the narrowest bandwidths may be obtained by making b equal zero (an active integrator). This ideal integrator also eliminates the VCO center-frequency setting problem.

C. S. Weaver
Philco Corp.
Western Dev. Labs.
Palo Alto, Calif.

An Experimental Technique for Parametric Devices*

In experimental work at radio frequencies an alteration to a circuit may be accomplished by unsoldering one component and soldering in another. Unfortunately, in the microwave region of the spectrum, an equivalent change usually demands an elaborate machining operation. We present here a description of a method which we have used to facilitate rapid alterations to an experimental parametric amplifier, together with a brief note on the performance of the amplifier.

The amplifier is completely coaxial, and in its experimental form consists basically of a center conductor and an array of brass washers. These washers all have the same outer diameter (2.25 in), but vary in inner diameter and thickness. They are clamped in a rack by a slight longitudinal pressure,

* Received February 12, 1962.

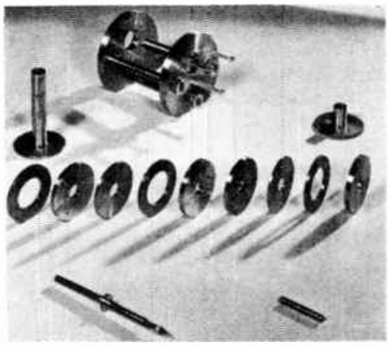

Fig. 1—Experimental *L*-band parametric amplifier, exploded view.

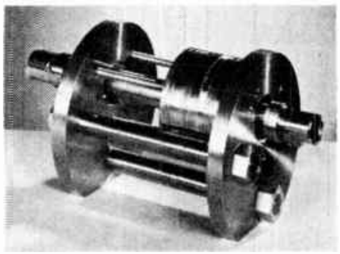

Fig. 2—The amplifier assembled.

and together comprise the outer conductor of a coaxial line. Some of the washers have a relatively large inner diameter, and, to form a radial rejection filter, one of these special washers is sandwiched between two of normal inner diameter. The rejection frequency, is, of course, determined by the inner diameter of the special washer. The washers and the clamping rack, together with the other parts of the amplifier, are shown spread out in Fig. 1.

The essential feature of the clamping rack is a pair of heavy drill rods which support the assembly of washers. These rigidly connected, parallel drill rods have a space between them which is considerably less than the diameter of the washers. Thus, they accurately position the washers, but they do not prevent their removal and rapid replacement once the center conductor has been shifted lengthwise. A stock of washers of various dimensions is available to permit modification of the amplifier's configuration.

The actual amplifier, which is seen assembled in Fig. 2, is a one-port, nondegenerate difference frequency reactance amplifier. The normal inner and outer diameters have been chosen to match General Radio standard connectors. The diode is mounted at the end of a coaxial line; dc connection between the diode and one of the washers is made by three radial wires spaced 120 degrees. These wires have appreciable inductance at the pump frequency. This washer is insulated for dc with 0.003-in teflon dielectric, thus permitting the application of dc bias. This insulated washer, through which the bias voltage is applied, is the fourth washer from the right in Fig. 1.

The diode is a Microwave Associates Type MA450E. The impedance presented to the diode by the source is controlled by a matching device in the signal line, while the idler resonant cavity is made up of a short section of the coaxial line which is bounded by two idler rejection filters.

The pump frequency is 9800 Mc. The pump energy is fed in (from the right in Fig. 2) and is capacitively coupled to the diode. A radial rejection filter in the signal line isolates the pump and signal circuits.

The signal frequency is 1315 Mc and the 3-db bandwidth at 20-db gain is 30 Mc. The over-all noise figure is 2.0 db, which includes a contribution of 0.3 db from the circulator and 0.1 db from the second stage. At a sacrifice of a further 0.25 db in noise figure, the gain-bandwidth product may be increased from 300 to 400 Mc.

A. C. Hudson
E. J. Stevens
Radio and Electrical Engineering Division
National Research Council
Ottawa, Can.

On the Use of Pulse Compression for the Enhancement of Radar Echoes from Diffuse Targets*

A diffuse radar target is one which is made up of a large number of discrete targets each of which is quite small. The specific example of this kind of target considered here is a weather cloud which is composed of myriads of water droplets. These droplets may range in diameter from on the the order of a micron to on the order of a centimeter. In this case the phasor $\check{E}_i$ representing the (radar) voltage return from the ith droplet of this ensemble is given by[1]

$$\check{E}_i = B_i a_i^3 e^{j(\omega t - 4\pi r_i/\lambda + \phi_i)}, \qquad (1)$$

where

- B_i is a factor which is a function of transmitted power, frequency antenna gain, range, propagation attenuation, and the complex dielectric constant of the drop;
- a_i is the radius of the ith drop;
- ω is the radar angular frequency;
- r_i is the range of the ith drop;
- λ is the free space wavelength of the radar energy; and
- ϕ_i is the phase shift incurred upon scattering by the drop.

The time waveform of a radar echo from a diffuse target will be composed of a summation of echoes from each drop, each of which will be of magnitude $B_i a_i^3$ and of duration T where T is the duration of the transmitted pulse. Therefore, at a time t_0 the composite phasor $\check{E}$ of the return is given by the summation[2]

$$\check{E} = \sum_{i=1}^{N_0} B a_i^3 e^{j(\omega t_0 - 4\pi r_i/\lambda + \phi_i)}, \qquad (2)$$

where N_0 is the total number of drops in the effectively illuminated volume V_0 associated with t_0. Since N_0 is in actuality quite large, it is sometimes desirable to describe the drop size distribution by a continuous function $\rho(a)$ which is assumed to be homogeneous over V_0. This function is defined by letting $\rho(a)da$ be equal to the number of drops per unit volume with radius falling between a and $a+da$.

The summation in (2) may be thought of as a sum of N_0 two-dimensional vectors in phasor space. Further if $N_0 \gg 1$ and $Tc \gg \lambda$, where c is the speed of light, then the phase angle of each term may be considered uncorrelated or random. The problem, therefore, is one of a random walk in two dimensions in which the size of each step is subject to a probability density given by

$$p(E_i) = \frac{\rho[a_i = f(E_i)]}{3B^{1/3}E_i^{2/3}\int_0^\infty \rho(a)da}, \qquad (3)$$

where E_i is the magnitude of $\check{E}_i$ and $f(E_i)$ is the appropriate function [from (1)] relating a_i to E_i. The solution of the two-dimensional random walk problem for large N in which the step size varies according to a probability distribution is given in the literature[3] and is

$$W(E)dE = \frac{2E}{N\langle E_i^2\rangle}\exp\left[\frac{-E^2}{N\langle E_i^2\rangle}\right]dE, \qquad (4)$$

where E is the magnitude of $\check{E}$, $W(E)dE$ is the probability that the magnitude of the (phasor) summation of (2) will fall in the range between E and $E+dE$, and $\langle E_i^2\rangle$ is the expected value of the square of the step size. From (1) and (3) $\langle E_i^2\rangle$ is found to be

$$\langle E_i^2\rangle = \int_0^\infty E_i^2 p(E_i)dE_i = \frac{B^2}{\int_0^\infty \rho(a)da}\int_0^\infty a^6\rho(a)da. \qquad (5)$$

It is evident from (4) that the statistics of the magnitude of the total voltage return are in the form of a Rayleigh distribution which depends only upon the product $N\langle E_i^2\rangle$. In order to determine the effect of a pulse compression process upon these parameters and the resultant effect upon the probability distribution consider an idealized rectangular representation of pulse compression. If the transmitted pulse is modulated in such a way that a pulse compression may be effected, then upon compression each individual pulse return in the time waveform will become shorter by some factor $\beta = \tau/T$ where τ is the duration of the compressed pulse. Conservation of energy dictates that $E_{ia} = E_{ib}/\beta^{1/2}$ which means

$$B_a = B_b/\beta^{1/2}, \qquad (6)$$

where the subscripts a and b stand for after and before compression. Eq. (6) reiterates the fact that for the case of a discrete target pulse compression (ideally) increases the effective signal magnitude by $\beta^{1/2}$. At the same time there will also be an improvement in range resolution by the factor β. This

* Received February 16, 1962.

[1] This equation assumes that the Rayleigh approximation to the back-scattering cross section of the drop is valid, *i.e.*, $a \ll \lambda$. In this case the back-scattering cross section is proportional to a^6.

[2] Note that the subscript has been dropped from the factor B since it may be assumed that B_i will not vary significantly over V_0.

[3] J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," M.I.T. Rad. Lab. Ser., McGraw-Hill Book Co., New York, N. Y., vol. 24, pp. 52–53; 1950.

range resolution improvement is due to a reduction, by the factor β, in the effectively illuminated volume. It follows from this fact that

$$N_a = \beta N_b \quad (7)$$

for regions greater than a distance $Tc/2$ (in the direction of propagation) from the edges of the target region. It should be observed that N_b is in general sufficiently large so that the inequality $N_a \gg 1$ will remain valid for any reasonable value of β.

Using (5)–(7), it can be seen that

$$N_a \langle E_i^2 \rangle_a = N_b \langle E_i^2 \rangle_b, \quad (8)$$

and as a consequence the statistics of E_r remain invariant upon pulse compression.

The physical interpretation of this conclusion is that although the compression process increases the magnitude of the return from each individual target element, the number of individual returns that come in simultaneously at any given time is proportionately reduced. And as has been shown above, these two effects exactly cancel in the summation process leaving the total time waveform (statistics) essentially unchanged. Thus, when pulse compression is considered it may be seen that there is a certain amount of parallelism as well as a notable difference between the cases of diffuse and discrete targets. When system performance has reached a peak power limitation and average power is increased by increasing the pulse duration, range resolu- is lost proportionately in both cases. However, the (peak) SNR is proportionately improved in the case of a diffuse target (due to an increase in N) while the (peak) SNR associated with the discrete target remains unchanged.[4] If this longer pulse is now modulated and compressed, range resolution is regained in both cases, but the (peak) SNR in the diffuse case remains unchanged while that associated with the discrete target is proportionately increased. Thus, it is seen that the end result is the same in both cases; the compression process itself only regains lost resolution for the diffuse target while it improves both resolution and (peak) SNR for discrete target case.

DAVID FRYBERGER
Armour Res. Foundation
Illinois Inst. Tech.
Chicago, Ill.

[4] The system noise bandwidth is assumed constant here since both cases are subject to the same constraints.

Industrial Patentry*

The occasion of issue of the 3,000,000th Patent calls to mind some troublesome aspects of industrial patentry.

Basically, the patent evolution was conceived to provide incentive to an individual, the inventor, and to provide a period of time during which he could reap the fruits of his invention while protected against imitation. These fruits are profit and prestige.

In our industrial process, the individual must sign a patent assignment agreement, usually as a condition of employment. Such agreements often cite company patent plans which vary in the liberalness of their terms according to the company. Since an agreement usually connotes a half-way meeting place where the parties give toward an equitable exchange, the inventor is here cautioned that if he looks for a great deal more than *the employment itself*, he generally looks in vain. There is no easy way, or obvious plan, to circumvent this loss of some of the original point to patentry in our present industrial system. But this does not mean that we should not be looking for better ways. Perhaps we should at least review the premises in the following numbered series of paragraphs.

1) The patent system is intended to provide an incentive toward original thought. The thought process is an attribute of *homo sapiens*. While confessing to little research on the subject, I feel that patentry, in its essentials, relates the concept of the individual, and the concept of incentive, in an intimate union. A patent can issue to more than one party but it is my opinion that the total constructive good of a patent is inversely proportional to the number of people involved.
2) The patent system is intended to provide protection against theft or depredation. It provides the inventor with an avenue of redress. It is not automatic. The inventor must initiate the action to restate his primacy and thus the protection is primarily a defensive advantage.
3) Industrial patentry adds a new dimension to the patent system and I am not sure that this is clearly recognized. In most of the present industrial agreements known to the author, the inventor exchanges his patent rights for employment. The agreement favors the employer. This is probably a justifiable starting place because the individual is usually more articulate in his objectives than the company. The individual can discern easier *what is best for him* than the company can discern *what is best for it*. It is the nature of all such agreements therefore to obligate the individual in certain specific matters without obligating the company very much, if at all.

In an assignment to an industrial organization, it is intended, I am sure, that the foregoing elements 1) and 2) be preserved. We shall of course assume good faith by all parties. Without it there is little point to this discussion. Even with good faith, however, it is apparent that the elements of incentive and protection are not entirely preserved.

Let us look at some delicate balances. If an incentive plan is designed around the yield from a patent (royalty or similar reckoning) the inventor may become diligent on a single matter which has little real future for the company, because it is the one extra benefit *he* could receive.

If the incentive is designed around the man's progress in the company, he might become interested in patentry which disclosed little of a new nature just in order to accumulate a number of patents.

If the plan is conceived primarily to protect the company, with little reduction-to-practice, the individual may "dry up." He may in all good faith become devoid of ideas simply because there seems to be little future in them. He might even go further and start to think that somehow the original germ of his thoughts arrived outside the framework of his obligation to his company to divulge them.

Regardless of the original good faith, if industrial implementation of the patent process results in extreme pursuits and serious imbalance in either direction, the whole process becomes a drain on our resources rather than a boon to them.

The fundamental question remains to be answered. By what method can we keep the individual's innovation motivation high and keep his entire energy devoted to his company, while providing his company with a good reason to have a patent plan and a good reason to implement it? While based on incentive, the question above should also relate to protection, in which both the individual and the company have a stake. The individual has a technical prestige (a forgotten part of which certainly involves an investment in education and experience) to safeguard and enhance, and the company has its primary business investment to protect.

The following personal observations may be helpful. If a price tag can be set on each patent prosecution, then I feel that much of the "junk" will be abandoned early by the inventor and his company. The common sense of the inventor, and the common sense of the company's "guiding light" in its patent effort should prevail.

I feel that the company "guiding light" in the patent area should not have any line technical interest which could alter his objectivity. He should, however, be thoroughly sympathetic to the patent concept. It should go without saying, too, that no element in the industrial patent effort should be rewarded on the basis of size or volume. This includes the head of the patent department.

I feel that upon proper presentation to a Board of Directors, annually, and after their deliberation, the individuals who have achieved patents or filings of note should be conspicuously honored in accordance with a definitive result of the Board's deliberation. This honor should be both professional and monetary. Admittedly, this idea could not survive without a continuous executive "push."

Perhaps, the basic nature of successful patentry is compounded of a judicious interplay between those two autocrats, prestige and profit. The corporate nature of industry may have to be reborn so that it can be expressed in a unified, and therefore unitary, approach to profit. Somehow we must get back to a single autocratic being dedicated to profit. This being can then find common cause with that other autocrat, prestige.

* Received February 23, 1962.

Prestige is more naturally an individualistic attribute. A dual relationship between these two autocrats, wherein prestige and profit can be interchanged, will restore the original virtue of the patent concept. It appears that when the relationship strays beyond this duality, to a three-poled problem and on to companies, committees, really to pluralities of any kind, the whole thing breaks down.

I am seeking the observations of others on this subject. Perhaps the viewpoint deserves further clarification or even outright refutation. Certainly there are more answers to choose from than the ones offered here.

FRANKLYN E. DAILEY, JR.
65 Oswego Street
Springfield, Mass.

On Network Realizability Conditions*

Recently Hazony and Nain[1] stated a set of realizability constraints on the Z matrix of a passive n-port. As pointed out by Slepian[2] these conditions are not sufficient for realization. Since the revised conditions stated by Hazony and Nain[3] are also not sufficient, it seems that further clarification is necessary. This is especially true since other somewhat different, but admittedly incomplete, conditions also have been given.[4]

Here we give the necessary and sufficient conditions for an impedance matrix Z to correspond to a finite, passive n-port. Because of space limitations we refer elsewhere for the proof and don't even attempt a definition of terms such as linear, finite, passive and n-port.

Consider a linear, time-invariant, passive n-port N described by an $n \times nZ$ matrix. Then by exciting with increasing exponentials we can see that Z is necessarily positive real,[5] *i.e.*,

1) $Z(p)$ is analytic in Re $p>0$
2) $Z^*(p)=Z(p^*)$ in Re $p>0$
3) $Z_H(p)$ is positive semi-definite in Re $p>0$.

Here $p=\sigma+j\omega$, a superscript asterisk denotes complex conjugation, and Z_H is the Hermitian part of Z.

If N is finite then Z must be rational and condition 2) implies real coefficients; we then call Z real-rational. In the real-rational case, that considered by Hazony and Nain,[3] 3) implies 1). In a form closer to that given given by Hazony and Nain[1] the positive real conditions in the real-rational case can be stated as follows:

Theorem: The necessary and sufficient conditions that an $n \times n$ matrix $Z(p)$ be the impedance matrix of a finite, passive n-port N are

1) Z is real-rational and
2) Z is analytic in Re $p>0$ and
3) Poles of Z on Re $p=0$ are simple (including infinity) and
4) The residue matrix of Z for each pole on Re $p=0$ (including infinity) is Hermitian with every principal minor non-negative and
5) All principal minors of Z_H are non-negative for each p on Re $p=0$ for which they are defined.

Conditions 4) and 5) are equivalent to the respective statements that the residue matrices (on Re $p=0$) and $Z_H(j\omega)$ are positive semi-definite. Each of these conditions can be given a physical interpretation. Thus 1) states that N can be built with a finite number of real valued elements (including gyrators and transformers). 2) and 3) state that N is stable but perhaps not asymptotically stable. 4) indicates that poles on $p=j\omega$ are due to lossless subnetworks. Finally 5) shows that the average power input in the sinusoidal steady state is nonnegative (recall that the steady state can't be defined for open circuit natural frequencies). Several synthesis methods prove the sufficiency; such are those of Oono and Yasuura,[6] Belevitch[7] and Newcomb.[8] The necessity proof relates the conditions of the theorem to the positive-real definition.[5] This follows the standard *pr* test[9] and is available in notes for lectures given at Stanford.

To some extent we can compare this with the statements in the previous notes. We can write

$$Z = R_{SY} + R_{SS} + jX_{SY} + jX_{SS}$$

where the R's and X's are real matrices and the subscripts SY and SS stand for symmetric and skew-symmetric, respectively. In Hazony and Nain[3] the condition det $(\text{Re } Z)=\det(R_{SY}+R_{SS})\geq 0$ for $p=j\omega$ is stated. Note that this doesn't agree with condition 5) of the above theorem which requires det $(R_{SY}+jX_{SS})\geq 0$. Of course the requirement of det $(\text{Re } Z(j\omega))\geq 0$ is a wrong condition, as seen by the example

$$Z(p) = \begin{bmatrix} -1 & 2 \\ -2 & 1 \end{bmatrix}$$

which has det $Z(p)=3>0$ but can't be realized by a passive N. In Hazony[4] the condition det $R_{SY}\geq 0$ for $p=j\omega$ is given. This is seen to be a necessary condition, but much more is required, since det $Z_H(j\omega)\geq 0$ must hold. The necessity of det $R_{SY}\geq 0$, $p=j\omega$, can be seen by connecting transformers to N in the manner attributed to Brune.[10] However, this interpretation fails when looking at $Z_H(j\omega)$ where the nonphysical complex transformer would have to be used.

ACKNOWLEDGMENT

The author would like to thank R. Espinosa who brought the relations of Hazony[4] to his attention, Professor C. A. Desoer who guided the work of which the above theorem is a portion, and Professor E. S. Kuh who suggested the existence of the theorem.

R. W. NEWCOMB
Department of Electrical Engineering
Stanford University
Stanford, Calif.

* Received February 26, 1962.

[1] D. Hazony and H. J. Nain, "A synthesis procedure for an n-port network," PROC. IRE, vol. 49, pp. 1431–1432; September, 1961.

[2] P. Slepian, "Comments on a synthesis procedure for an n-port network," PROC. IRE (*Correspondence*), vol. 50, p. 81; January, 1962.

[3] D. Hazony and H. J. Nain, "Author's comments," PROC. IRE (*Correspondence*), vol. 50, p. 81; January, 1962.

[4] D. Hazony, "Two extensions of the Darlington synthesis procedure," IRE TRANS. ON CIRCUIT THEORY, vol. CT-9, pp. 284–288; September, 1961, See p. 287.

[5] D. C. Youla, L. J. Castriota, and H. J. Carlin, "Bounded real scattering matrices and the foundations of linear passive network theory," IRE TRANS. ON CIRCUIT THEORY, vol. CT-6, pp. 102–124; March, 1959. See p. 122 (Def. 21).

[6] Y. Oono and K. Yasuura, "Synthesis of finite passive $2n$-terminal networks with prescribed scattering matrices," *Memoirs of the Faculty of Engineering*, Kyushu University, vol. 14, pp. 125–177; May 1954. See pp. 153–158 and 163–167.

[7] V. Belevitch, "On the Brune process for n-ports," IRE TRANS. ON CIRCUIT THEORY, vol. CT-7, pp. 280–296; September; 1960.

[8] R. Newcomb, "Synthesis of Non-Reciprocal and Reciprocal Finite Passive 2N-Poles," Ph.D. Thesis, University of California, Berkeley; 1960.

[9] D. F. Tuttle, Jr., "Network Synthesis," vol. 1, John Wiley and Sons, Inc., New York, N. Y., p. 182; 1958.

[10] E. A. Guillemin, "Synthesis of Passive Networks," John Wiley and Sons, Inc., New York, N. Y., pp. 7–9; 1957.

On the Origin of the Word "Radio"*

In regard to the Editorial in the September, 1961 issue of the PROCEEDINGS, I think that the reference mentioned to the earliest use of the prefix "radio" in the magazine *Tit-Bits* in May, 1898 has been taken from the Oxford English Dictionary, vol. 8, page 101. There is, however, a somewhat earlier use of that prefix—and in a somewhat more technical journal than *Tit-Bits*. The coining of the word "radioconductor," or rather its French equivalent "radioconducteur," no doubt goes back to Edouard Branly. The earliest use of the word "radioconducteur"—as far as I can see—appears in a footnote in a paper by Branly.[1] An English translation of this paper can be found in *Electrician*.[2] The translation of the footnote runs thus:

> My tubes of filings received the name "coherers" from Lodge and this name has been generally accepted. The expression is based on an incomplete examination of the phenomenon and on an inaccurate interpretation. I proposed the name "radioconductors" which recalls the essential property of discontinuous conductors of being excited by electric radiation. M. Ducretet uses my various radioconductors in the apparatus he has constructed to realise "Hertzian Telegraphy" without wires.

It seems to be most likely that Branly termed his device "radioconducteur" only

* Received February 9, 1962.

[1] E. Branly, "Sur la Conductibilité Électrique des Substances Conductrices Discontinues, à propos de la Télégraphie San Fils," *Compt. rend. Acad. Sci., Paris*, vol. 125, pp. 939–942; December 6, 1897. See p. 941.

[2] E. Branly, "On the electrical conductivity of discontinuous conducting substances," *Electrician*, vol. 40, p. 333; December 31, 1897.

after G. Marconi had used it in his first experiments on wireless telegraphic transmission in 1896. Branly, a physics professor at the Institut Catholique in Paris, had discovered the phenomenon in 1890.[3] He published a somewhat more detailed description in 1891[4] and a shortened version in English can be found in *Electrician*.[5]

Branly certainly did not try a transmission of intelligence over great distances. In fact, he only found that the phenomenon of a column of metal filings becoming conductive due to a spark discharge can be observed over a distance of about 20 meters (about 70 ft). It should, however, be borne in mind that Marconi used the Branly "radioconducteur" in his transatlantic transmission experiment.

There is a very interesting paper by Sir Oliver Lodge[6] in which he investigates the coherer principle under a wider aspect, though by no means belittling the achievements of Branly. There is an interesting passage in Lodge (p. 91):

> Mr. Marconi is to be congratulated on the results of his enterprise, the newspaper press of this and other countries have taken the matter up, popular magazine articles have been written about it, and so now the British Public has heard, apparently for the first time, that there are such things as electric waves, which can travel across space and through apparent obstacles to a considerable distance, and be detected in a startling fashion. Thus the public has been educated by a secret box more than it could have been by many volumes of Philosophical Transactions and Physical Society Proceedings; our old friends the Hertz waves and coherers have entered upon their stage of notoriety, and have become affairs of national and almost international importance.

In regard to the term "radiophone" there is an interesting paper in *Engineering*.[7] On the development of the term "radiophony" Preece writes (page 29):

> Many engineers have been investigating the subject, and it is rather amusing to notice the various titles adopted by them. Graham Bell, in his paper on the 27th August, 1880, adopted the title "Upon Production and Reproduction of Sound by Light." As recently as 21st April, 1881, in a paper read before the American Society of Science, he discusses the same question under the title, "Upon the Production of Sound by Radiant Energy." M. Mercadier, the head of the technical school of the telegraph administration in Paris, an extremely able experimenter, as well as a very clever physicist, has written several papers from week to week, published in *La Lumière Electrique* (and which have been read before the French Academy), which invariably have been entitled "Notes on Radiophony."

Preece continues:

> By radiophony, the term adopted by Mercadier, Bell and Tainter, and myself, I simply mean the production of sounds by radiant energy. . . . By "radiant energy" physicists now speak of the motion of the ether, that highly elastic medium, which fills all space. The heat of the sun, the light of the stars, the effects of which we call actinism, and all the physical effects that pass between astronomical bodies are transmitted by this medium, ether, and its movements or vibrations are called radiant energy. Sometimes this term is called "radiation," and it is a very frequent thing to see in papers at the present day the word "radiation" so employed.

During my investigations on the word "radio," I came across a short note in *Nature*[8] in which there is an astonishing picture (p. 18) (shown here as Figs. 1 and 2). The similarity of Fig. 2 to the receiving end of a modern microwave link is very surprising. More can be found on Bell's photophone in chapter 16 of the Bell biography by C. Mackenzie.[9]

H. Drubba
Geibelstr. 25
Hannover, Germany

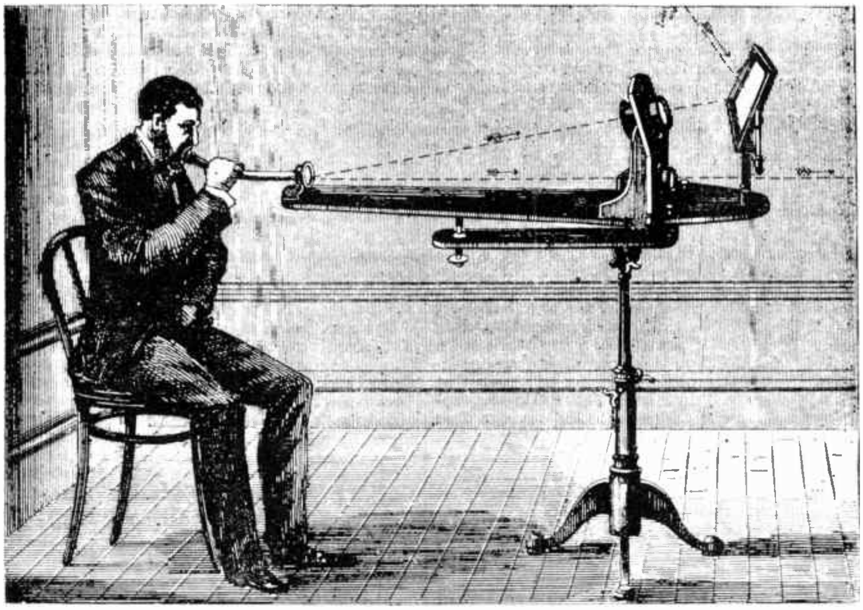

Fig. 1—Bell's articulating photophone. The transmitter. (From *Nature*, November 4, 1880.)

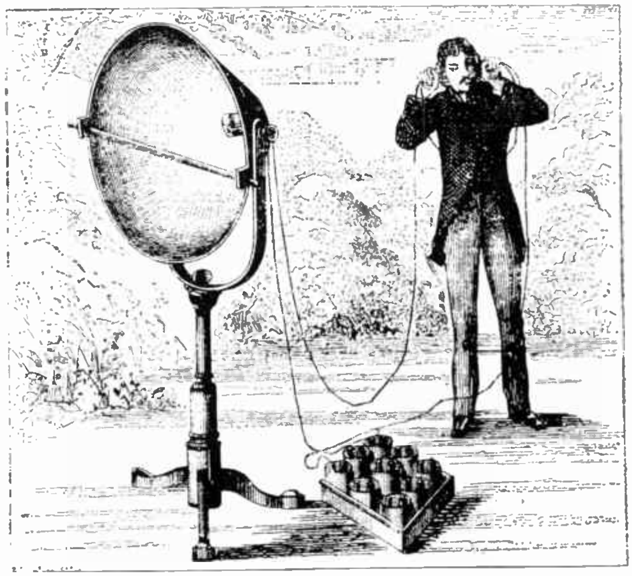

Fig. 2—Bell's articulating photophone. The selenium receiver. (From *Nature*, November 4, 1880.)

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On the Characteristic Function of a Stationary Random Process with Gaussian Envelope*

In two recent letters to the editor[1,2] Bunkin and Gudzenko have pointed out the following properties of a stationary random process $x(t)$:

1) If such a process is written in the form

$$x(t) = A(t)\cos[\omega_0 t - \theta(t)], \qquad (1)$$

the introduction of two random functions instead of one puts a restriction on the distribution (density[3]) of $\theta(t)$ which must necessarily be uniform. Furthermore the characteristic function $f(\xi)$ corresponding to $p(x)$, the distribution of the original process $x(t)$, is related to $W(A)$, the distribution of the envelope, by the simple integral transformation

$$f(\xi) = \int_0^\infty J_0(\xi A)W(A)dA. \qquad (2)$$

2) As a consequence of (1) among all the possible stationary processes only the ones with symmetrical distributions, *i.e.*, $p(x)=p(-x)$, can be written in the form 1 and an immediate relation may be established from (2) between the even moments of the distributions $p(x)$ and $W(A)$. Both properties can be derived assuming only independence of time of the distribution $W(A)$ and the conditional distributiln $W_\theta(\theta|A)$.

The purpose of this note is to use (2) to find the characteristic function of a stationary random process with Gaussian envelope.

Stochastic processes with approximately Gaussian envelopes have been reported lately in the literature, namely in the case of noisy oscillators[4,5] and in the so-called "shallow" fading.[6] Strictly speaking these processes are not stationary; Golay[7] has even shown intuitively the fundamental impossibility of defining a probability distribution for a noisy oscillator in view of the "random walk" type of behavior of the phase perturbations.

Still an engineering approach such as the one used by Brennan[6] in the problem of mild fading is particularly useful and can be extended even to non-monochromatic oscillators provided that only AM noise is considered. Substantially one thinks of intervals of time long enough to define a meaningful distribution but short enough to make negligible the dependence of time of the distribution itself.

The variance of the Gaussian envelope is assumed much smaller than the square of the mean value: this corresponds to most of the physical situations and makes the distribution vanishingly small at zero.

The required characteristic function is then

$$f(\xi) = \frac{1}{\sqrt{2\pi\sigma^2}}\int_0^\infty J_0(\xi A) \cdot \exp\left[-\frac{(A-A_0)^2}{2\sigma^2}\right]dA. \qquad (3)$$

Changing the variable of integration and extending the lower limit of the integral in view of the assumed small variance

$$f(\xi) \cong \frac{1}{\sqrt{2\pi\sigma^2}}\int_{-\infty}^{+\infty} J_0[(A_0+x)\xi] \cdot \exp\left(-\frac{x^2}{2\sigma^2}\right)dx \qquad (4)$$

and using the addition formula for the Bessel function of zero order

$$f(\xi) = \frac{1}{\sqrt{2\pi\sigma^2}}\int_{-\infty}^{+\infty}\left[J_0(A_0\xi)J_0(x\xi) + \sum_{n=1}^{\infty}2\cdot(-1)^n J_n(A_0\xi)J_n(x\xi)\right] \cdot \exp\left(-\frac{x^2}{2\sigma^2}\right)dx. \qquad (5)$$

The series can be integrated termwise and the resulting series is rapidly convergent in the range of interesting values, furthermore all the terms of odd order vanish. Retaining then only the leading term, the value of the integral is

$$f(\xi) = J_0(A_0\xi)\,{}_1F_1\left[\frac{1}{2};1;-\frac{\sigma^2\xi^2}{2}\right]. \qquad (6)$$

Taking advantage of the identity

$${}_1F_1[\tfrac{1}{2};1;-x] = \exp\left(-\frac{x}{2}\right)I_0\left(\frac{x}{2}\right), \qquad (7)$$

the final form of the characteristic function is

$$f(\xi) = J_0(A_0\xi)\exp\left(-\frac{\sigma^2\xi^2}{4}\right)I_0\left(\frac{\sigma^2\xi^2}{4}\right) \qquad (8)$$

As it was to be expected in view of the smallness of the ratio σ^2/A_0^2, the leading term of the characteristic function, as obtained by a series expansion of the modified Bessel function, is similar to the characteristic function of a process with a generalized Rayleigh distribution of envelope (familiar case of a sine wave and additive normal noise[8]).

The probability distribution $p(x)$ of the original process $x(t)$, is by definition

$$p(x) = \frac{1}{2\pi}\int_{-\infty}^{+\infty} f(\xi)e^{-jx\xi}d\xi. \qquad (9)$$

The leading term (aside from a scaled variance) is then the distribution found for instance by Rice[8] (case of high SNR) while correction terms can be evaluated by termwise integration.

RAFFAELE ESPOSITO
Selenia S.p.A.
Rome, Italy
Formerly with
Raytheon Research Division
Waltham, Mass.

* Received February 26, 1962.

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Comment on "A New Precision Low-Level Bolometer Bridge"*

In response to the IRE evaluation[1] of the bolometer bridge described by Reisener and Birx[2] I would like to suggest that bridge sensitivity is not the only criterion to be considered in making an evaluation as to what comprises the "best bridge detection and calibration method available today." In the existing microwave measurements art there are a variety of adaptations of the bolometer bridge concept, and the choice of what is best for a given application usually involves additional considerations such as operating convenience, cost, and the required accuracy of the measurement.

For example, in an application where operating convenience is the prime requirement, the "best" available method would probably include the use of one of the commercially available power meters which use audio frequency power for bridge balancing. In an application where accuracy is the most important consideration, one may forego the convenience of automatic bridge balancing and choose among the manually balanced dc bridges, some of which are also commercially available. Finally the "Self-Balancing D.C. Bolometer Bridge"[3] developed in this laboratory provides the accuracy expectancy of a high quality manual bridge and while retaining much of the operating convenience of the self-balancing audio bridges. This bridge has been commercially available for the past two years.

Admittedly, if one is interested in power measurements below the microwatt level, the greater sensitivity claimed for the method described by Reisener and Birx will presumably take precedence over all other considerations. At the milliwatt level, however, the cited techniques provide substantial improvements over the latter method in the areas of operating convenience and/or accuracy.

GLENN F. ENGEN
Microwave Power Standards
Radio Standards Laboratory
National Bureau of Standards
Boulder, Colo.

* Received February, 23, 1962.

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A Method for Obtaining Compatible Single-Sideband Modulation*

Recently Kahn published an article about a new type of amplitude modulation called compatible single-sideband modulation.[1] His methods to obtain CSSB modulation are rather complicated and therefore it seems to be useful to refer to our article,[2] in which a new, simpler method is described. The principle of this method consists of multiplication of a full-carrier SSB signal by itself and suppression of all the frequency components except those around the double-carrier frequency.

This very simple operation provides in the case of single tone modulation a full compatible SSB signal, consisting of only three frequency components, as is shown by the following calculations:

The full-carrier SSB signal

$$\left(\text{carrier frequency} = \frac{1}{2\pi}\,\frac{\omega}{2}\right)$$

$$\cos\frac{\omega}{2}t + a\cos\left(\frac{\omega}{2}+p\right)t$$

gives after multiplication by itself and filtering around the double carrier frequency a signal

$$\cos\omega t + 2a\cos(\omega+p)t + a^2\cos(\omega+2p)t$$

with an undistorted envelope

$$1 + a^2 + 2a\cos pt.$$

The same calculation for two cosine waves gives for the envelope

$$1 + a^2 + b^2 + 2a\cos pt + 2b\cos qt + 2ab\cos(p-q)t.$$

The last formula shows that the multiplication provides an ideal solution for CSSB modulation if it is possible to eliminate the intermodulation distortion in the envelope. It is remarkable that the increasing of the modulation depth by multiplication also reduces the intermodulation by a factor two. There are different means to eliminate the intermodulation: One is to separate the intermodulation terms by comparing the envelope, produced by a peak detector, with the original audio signal. Then the carrier ω is balanced modulated in amplitude with these terms and the output of the modulator is added in counter phase to the multiplied SSB signal. In the envelope, the intermodulation now almost disappears. Another method is to limit the CSSB signal and to modulate the so-obtained phase-modulated carrier with the undistorted audio signal. In this case it is very simple to determine the spectrum of the multiplied SSB signal after limiting by working the other way round: What spectrum gives

$$\cos\omega t + 2a\cos(\omega+p)t + a^2\cos(\omega+2p)t$$

after modulation with $a^2+2a\cos pt$. One finally finds for the spectrum

$$A_1 = a \quad A_0 = (1-a^2) \quad A_{-1} = -a(1-a^2)$$
$$A_{-2} = +a^2(1-a^2) \quad A_{-3} = -a^3(1-a^2), \text{ etc.}$$

(A_1 = upper sideband, A_0 = carrier, $A_{-1\ldots n}$ = lower sideband.) If the phase-modulated carrier is modulated not with a signal $a^2+2a\cos pt$, but with one of the shape $2b\cos pt$, the sideband components (B_{-1}, B_{-2}, etc.) do not totally disappear. The suppression depends on the choice of $K=b/a$ (see Figs. 1 and 2).

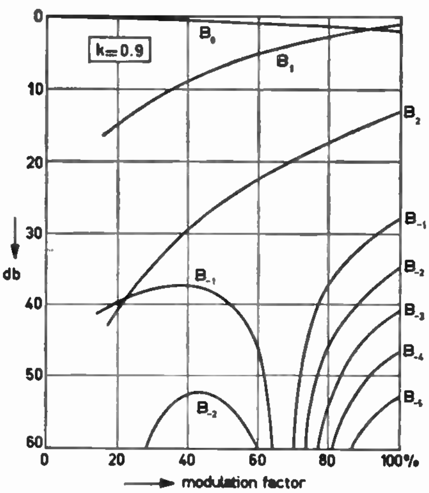

Fig. 1—Amplitudes of various sideband components as a function of the depth of amplitude modulation for adjustment parameter $K=0.9$.

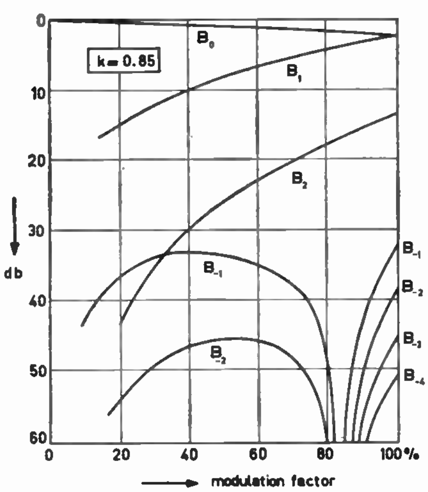

Fig. 2—Amplitudes of various sideband components as a function of the depth of amplitude modulation for adjustment parameter $K=0.85$.

It is very easy to extend the calculations in the case of an audio signal, built up of two cosine waves ($a\cos pt + b\cos qt$), by using the same method as has been shown for a single tone.

Now in the unwanted sideband a spectrum term $-ab\cos(\omega-p+q)t$ appears. Bearing in mind that $a+b$ does not exceed 0.5, it is easily seen the amplitude of this term is always below the level of -24 db.

In the laboratory an experimental transmitter, of which the block diagram is given in Fig. 3 below, has been in use for over a year and is shown to people interested in broadcasting.

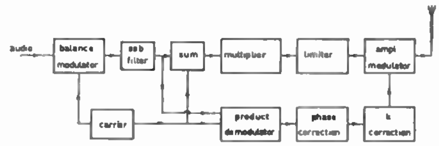

Fig. 3—Block diagram of the CSSB modulator.

The measurement results are in accordance with the theory.

T. J. van Kessel
F. L. H. M. Stumpers
J. M. A. Uyen
Philips Research Laboratories
N. V. Philips' Gloeilampenfabrieken
Eindhoven, Netherlands

* Received February 26, 1962.

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Author's Reply[3]

I would like to take this opportunity to comment on the above communication and on the paper[2] authored by van Kessel, Stumpers and Uyen.

Actually, the system they discussed was developed at the Kahn Research Laboratories as a first step in a program to determine the optimum phase stretch technique and is covered by our patent. Theoretically, it was obvious that this technique would not approach the optimum CSSB arrangement, but it was so simple to set up experimentally that we conducted tests and the results of van Kessel, Stumpers and Uyen are very similar to these early experiments. The reason that it was felt that this squared technique was not as satisfactory as the later developed 1.4 phase stretch method was that it was incapable of 100 per cent envelope modulation without serious degradation of the undesired sideband characteristic. (See Fig. 5 of their paper.[2])

The authors point out that a K or correction factor is required in the amplitude modulation branch if the undesired sideband radiation is to be kept at a fairly low value. They indicate that a factor of 0.85 would be about optimum. This means that the system's peak amplitude modulation, *i.e.*, the envelope modulation, is limited to 85 per cent and would produce a loss in signal-to-noise of approximately 1.5 db. While many may feel that this is not of great moment, AM broadcasters are continuously striving to produce full modulation and in many cases it is unfortunately true that even overmodulation is common. In any case, 100 per cent modulation capability is necessary and this requirement is fully met by modern CSSB equipment. (Actually, measurements of 120 per cent modulated CSSB waves show relatively good spectrum characteristics.) The reason for the modulation limita-

[3] Received April 16, 1962.

tion in the squared technique is that the desired phase modulation curve is not closely approximated by the squared phase modulation function (see Fig. 9 of Reference 1).

As to the method of elimination of intermodulation distortion, this technique is identical to the technique used and discussed in my early papers. That is, one either derives the envelope function from the input signal or from a product modulator. In this manner, both the techniques described by van Kessel, Stumpers and Uyen and our modern CSSB systems are theoretically free of envelope distortion. Measurements of equipment indicate that an over-all harmonic distortion figure of less than 1 per cent is easily obtained.

I am, of course, pleased to see that such distinguished engineers have independently derived similar positive conclusions about the CSSB advantages and I appreciate the very kind comments made by these authors in their full paper and also by the Editor of the *European Broadcasting Union Review*. May I suggest that the reader secure the complete paper by van Kessel, Stumpers, and Uyen,[2] as it is most interesting and important.

LEONARD R. KAHN
Kahn Research Laboratories, Inc.
Freeport, N. Y.

A Discussion of Recently Proposed Aether Drift Experiments*

It has been suggested in recent correspondence that the hypothesis of an aether, in addition to providing a satisfactory interpretation of relativistic phenomena, predicts a number of first-order effects which are not predicted by special relativity. On this basis a number of experiments have been proposed to measure the aether drift; these include Rapier's atomic clock experiment [1], Ruderfer's proposed observations on planets and satellites [2], and Carnahan's aberration experiment [3].

It is the purpose of this note to point out that the theory given for some, and probably all, of these proposals is fallacious, and that experiments of a considerably more recondite nature may be necessary to demonstrate the existence (or otherwise) of an aether.

The simplest of the proposals was that of Rapier [1], who suggested making a direct measurement of the one-way passage of a light signal between two synchronized atomic clocks. It is insufficient to refute this with the traditional argument that it is impossible to show that the clocks remain synchronized on separation [4], for if a positive effect were observed its magnitude would depend on the orientation of the light path relative to the aether and would presumably show periodic variations correlated to the earth's rotational and orbital motion; any desynchronization resulting from the separation process could be detected by varying the speed and path of separation.

Unfortunately, however, the aether hypothesis predicts a precisely null result. The fallacy arises from neglect of the second-order frequency change experienced by a moving clock; during the separation of the clocks their difference in frequency builds up into a first-order phase shift which exactly compensates for the effect of aether drift on the transit time of the light signal. In the simplest case, if the velocity relative to the aether is v and one clock is taken to a distance L with velocity u, the absolute time lost by the transported clock is

$$\frac{L}{u}\left\{\left[1-\frac{v^2}{c^2}\right]^{1/2}-\left[1-\left(\frac{v+u}{c}\right)^2\right]^{1/2}\right\}$$

$$\simeq \frac{Lv}{c^2}\ \text{(to first order in } v/c\text{)}.$$

Since this absolute time taken by the light signal is

$$\frac{L}{c}\left[1-\frac{v}{c}\right],$$

the measured time is L/c regardless of the value of v. An exact calculation (assuming the frequency to be a function of velocity but not of acceleration) shows that the result is correct to all powers of v/c and independent of the paths taken by the clocks. The only observable effect is the well-known "clock paradox" which depends only on the changes in speed of the clocks measured in any inertial system and is independent of the value of a superimposed aether drift [5], [6], [7].

The same argument applies to the second suggestion of Ruderfer [2], requiring observations of a stable oscillator in an earth satellite. The anticipated phase-modulation will be exactly compensated by the periodic phase shift of the clock itself resulting from its motion through the aether. This prediction is, in fact, consistent with the results of phase comparisons between highly stable frequency transmissions made daily over distances of thousands of miles. No periodic phase modulations (other than those caused by spurious propagation variations) appear to have been observed [8].

Ruderfer's first proposal, however, involving detailed observations of planetary motion, has to be examined with more caution, since an exact solution of the problem is impossible without knowing the mechanism of gravitational interaction or the likely effect on it of motion through the aether. Nevertheless it is difficult to see how Ruderfer's first-order "timing error" could manifest itself; simple position-time measurements of a planet would yield no information at all since, for example, a straightforward calculation shows that a circular orbit traveling through the aether at speed v would have the same appearance as an elliptical orbit with eccentricity $v_0v/2c^2$, where v_0 is the planet's orbital velocity. This eccentricity cannot be independently checked by distance measurements since it is of the same order (probably 10^{-6} or less) as the intractable second-order relativistic effects. Attempts to detect the "timing error" by observations of a natural satellite of the planet should also give a null result, for the same reason as the earth satellite experiment discussed above, assuming that gravitational clocks undergo the same frequency shifts as atomic and electromagnetic clocks.

Finally we can consider the possibility of aberration experiments. As Carnahan points out [3] these depend for their success on finding a source of radiation whose propagation direction, as measured by an observer at rest in the aether, is unaffected by the motion of the source through the aether. Since the fundamental mechanism of the emission of radiation (or photons) is unknown, it would be premature to assert that such a source cannot exist. But in so far as radiating systems can be analyzed in terms of simple dipoles, or in terms of simple interactions between fundamental particles, elementary considerations of energy and momentum balance indicate that the direction of radiation as seen by the source is fixed, and no aberration effects are therefore observable.

In view of the advantages of an aether hypothesis in interpreting relativistic phenomena [7], it is of interest to inquire whether there might be any other ways of detecting the aether. It is conceivable, for example, that experiments might reveal a slight variation in the velocity of photons with frequncy [9]; the principles of special relativity would then become merely good approximations, and comparisons between high-energy γ rays and light rays would probably enable an aether drift to be detected. Other possibilities could arise from departures from conventional electrodynamics in strong electric or magnetic fields [10], or from detailed studies of the properties of fundamental particles in motion.

In the absence of experimental proof of the absence of such effects it is not permissible to assert on philosophical grounds that attempts to measure the aether drift are meaningless. Nevertheless it seems unlikely that this will be achieved by experiments of the type suggested in recent correspondence.

P. F. SMITH
National Institute for Research
in Nuclear Science,
Harwell, England

* Received February 26, 1962.

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Some Remarks on "Penetration of the Ionosphere by Very-Low-Frequency Radio Signals"*

In a recent article,[1] Leiphart, Zeek, Bearce, and Toth have published some interesting data furnished by the Navy satellite "LOFTI I" on the propagation characteristics of VLF radio signals through the ionosphere. Their conclusions, based on the experimental results, seem to confirm early theoretical investigations done by the writer.[2] Although it is likely that the authors of the article have carried on similar investigations, it may be of interest to communicate the writer's results.

1) The authors have recorded a propagation loss of 38 db by day. They have calculated the absorption loss to be 27 db; inter-reflection losses were not calculated. The writer used the "LOFTI I" data to calculate inter-reflection as follows: Fig. 2 of the article in question shows that the lower boundary of the ionosphere can be approximated by a sharp density interface of 2×10^5 electrons per cubic centimeter. The wave is assumed to be linearly polarized; the interface reflection is then calculated on the basis of a gyrotropic semi-infinite uniform plasma.[2] A curve giving reflection losses at a sharp interface between free space and such a plasma is shown in Fig. 1 of this communication. In accord with the "LOFTI I" experiment, a gyrofrequency of 1 Mc per second and a signal frequency of 18 kc per second were used in the calculations. The plasma frequency for the assumed electron density at the interface is 4×10^6 cps; the corresponding reflection loss read off from Fig. 1 is 12 db.

The total propagation loss, reflection loss (12 db), plus absorption loss (27 db) amounts to 39 db which is in close agreement with the 38 db figure obtained in the "LOFTI I" experiment.[1] The additional inter-reflections above 100 km created by the electron density gradient of the medium have been neglected because the gradient is small over several wavelengths except for a narrow region about 200 km. Calculations show that this small region introduces an additional reflection loss less than 4 db.

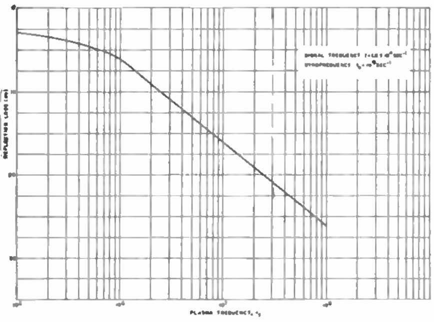

Fig. 1—Reflection loss vs plasma frequency at the interface between free space and a semi-infinite lossless uniform gyrotropic plasma.

2) The assumption of a sharp interface does not hold well at night as seen from Fig. 2 of Leiphart, *et al.* However, if the electron density for an equivalent uniform ionosphere is chosen at 200 km (the electron density gradient above this point begins to get small over a length comparable to a wavelength), one obtains an electron density of 10^5 electrons per cubic centimeter and a corresponding plasma frequency of 2.8×10^6 cps. The reflection loss read off Fig. 1 of this communication is about 10.5 db. The absorption loss calculated by the authors is 2 db.[1] The total loss amounts to 12.5 db, which again is in close agreement with the 13 db figure quoted in the "LOFTI I" experiment.

In neither calculation were spreading losses taken into account. The writer is intending to extend a previous analysis[3] of trapped modes in gyrotropic plasma slabs bounded by sharp interfaces in electron density to the case of varying density at the interface in order to account for the spreading loss.

3) A transmission delay as great as 30 times free space delay was recorded by the authors. This order of magnitude is also in agreement with some of the writer's earlier investigations.[4] The results of this latter investigation shown in Fig. 2 of this communication give group and phase velocities vs altitude; the plot is based on a gyrofrequency of 1 Mc per second, a signal frequency of 18 kc and on the data of Fig. 1 in the article discussed. It is seen that the phase and group velocities of the VLF signals are between 1/10 and 1/100 of the speed of light in vaccum over the major part of their path which accounts for a transmission delay of the order of 10 to 100 times the free space delay.

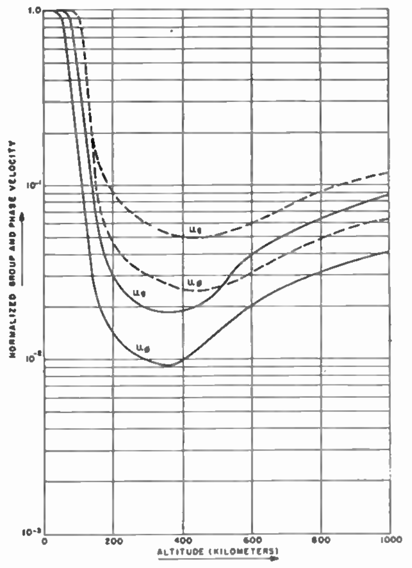

Fig. 2—Group and phase velocities of a VLF electromagnetic wave guided by the earth's magnetic field.

— — — — RH wave velocity-night
———— RH wave velocity-day
V_g = group velocity
V_ϕ = phase velocity

$$\mu_g = \frac{\sqrt{1+\dfrac{\nu_p^2}{\nu_b-1}}}{1+\left[\dfrac{\nu_b/2}{\nu_b-1}\right]\dfrac{\nu_p^2}{\nu_b-1}} = \frac{V_g}{C_v} = \text{normalized group velocity}$$

$$\mu_\phi = \frac{1}{\sqrt{1+\dfrac{\nu_p^2}{\nu_b-1}}} = \frac{V_\phi}{C_v} = \text{normalized phase velocity}$$

$$\nu_p = \frac{f_p}{f}$$

$$\nu_b = \frac{f_b}{f}$$

C_v = speed of light in vacuum
f_p = plasma resonant frequency
f_b = gyrofrequency $= 10^6\ \text{sec}^{-1}$
f = signal frequency $= 1.8\times 10^4\ \text{sec}^{-1}$

H. Hodara
Research and Development Div.
The Hallicrafters Co.
Chicago, Ill.

* Received February 23, 1962.

[1] J. P. Leiphart, R. W. Zeek, L. S. Bearce, and E. Toth, "Penetration of the ionosphere by very-low-frequency radio signals—interim results of the LOFTI I experiment," Proc. IRE, vol. 50, pp. 6–17; January, 1962.

[2] H. Hodara, "The use of magnetic fields in the elimination of the re-entry radio blackout," Proc. IRE, vol. 49, pp. 1827–1830; December, 1961.

[3] H. Hodara and G. I. Cohn, "Wave propagation in magneto-plasma slabs," IRE Trans. on Antennas and Propagation, vol. AP-10, pp. 452–459; July, 1962.

[4] H. Hodara, "A new approach to space communications during re-entry," presented at the ARS 15th Annual Meeting, Washington, D.C.; December 5–8, 1960.

Contributors

Jean A. Develet, Jr., (S'51–A'55–M'59) was born in St. Albans, N. Y., on February 5, 1933. He received the B.E.E. and M.E.E. degrees from the Polytechnic Institute of Brooklyn, N. Y., in 1953 and 1954, respectively.

While a graduate student at Polytechnic he was a Junior Research Fellow engaged in pulse-echo fault location studies for the Signal Corps. He entered the U. S. Army in 1954 and served with the Corps of Engineers at Fort Belvoir, Va., as Chief of the Utilities Section, The Engineer Test Unit, where he was responsible for the planning and reporting of field tests on such diverse items as liquid oxygen generators for guided missile support, electrical power generators of 3-kw to 5000-kw capacity, fire-fighting equipment, and water purification plants for field armies.

In 1956 he joined the technical staff of Ramo-Wooldridge Corporation, Los Angeles, Calif., where he was employed in multi-megawatt modulator construction, video receiver design, high-power millimicrosecond pulse circuitry, and traveling-wave tube and microwave measurements, in addition to power supply design for traveling-wave tubes. He became a member of the technical staff of Space Technology Laboratories, Los Angeles, in 1959, working in system analysis pertaining to space communications, tracking, and telemetry; he also worked on the ground station evaluation studies for the Pioneer V space probe experiment, remote tracking station time synchronization, and atmospheric and ionospheric refraction, as well as communication system synthesis and modulation techniques for NASA's Project Relay.

Mr. Develet is a member of Sigma Xi, RESA, Tau Beta Pi, and Eta Kappa Nu.

❖

John H. Gebhart was born in Lakeland, Fla., on December 20, 1929. He received the B.S.E.E. degree (*magna cum laude*) from the University of Miami, Coral Gables, Fla., in 1951.

He was then employed as a Test Planning Engineer with the Western Electric Company, Winston-Salem, N. C. After military service as an Instructor in radar theory and maintenance at the Signal Corps Engineering Laboratories, Fort Monmouth, N. J., he did graduate work at the University of Illinois, Urbana, and received the M.S.E.E. degree, in 1956. From 1956 to 1959 he was an Electronics Engineer in the Laboratory of Biophysics of the National Institute of Neurological Diseases and Blindness.

Mr. Gebhart was a charter member of the Engineers' Honor Society at the University of Miami.

❖

James P. Gordon was born in New York, N. Y., on March 20, 1928. He received the B.S. degree from the Massachusetts Institute of Technology, Cambridge, in 1949, and the M.A. and Ph.D. degrees from Columbia University, New York, N. Y., in 1951 and 1955, respectively.

Dr. Gordon joined the staff of Bell Telephone Laboratories, Murray Hill, N. J., in 1955, and is presently Head of the Quantum Electronics Department there. He has written several technical articles in the field of quantum electronics. His research in quantum electronics has involved work on molecular beam masers, paramagnetic resonance and solid-state masers.

He is a member of the AAAS, the American Physical Society and Sigma Xi.

❖

Christian Guld was born in Vigsnæs, Denmark, on September 9, 1920. He received the M.S. degree in electrical engineering from the Royal Technical University, Denmark, in 1946.

In 1946 and 1947 he was Research Assistant at the Laboratory for Telegraphy and Telephony, Royal Technical University, and in 1948 he worked at the Radio Receiver Research Laboratory of the Danish Academy of Technical Sciences on a research grant. Since 1948 he has worked at the Institute of Neurophysiology, University of Copenhagen, and has headed its Division of Instrumentation since 1952. He has been engaged in basic research in nerve and muscle physiology and has designed general purpose measuring equipment for work in this field. In physiology his main studies concerned the motor unit potential in normal human muscle and the link between membrane potential and contraction in striated muscle. His main interest in instrumentation was the development of input amplifiers and problems of interference rejection in electrophysiological recordings.

Mr. Guld is a member of the Institution of Danish Civil Engineers, Danish Biological Society, Scandinavian Physiological Society, a Danish council representative of the IFME and a corresponding member of the Biological Engineering Society.

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Hsien Ching Ko (S'53–M'56–SM'60) was born in Formosa, China, on April 28, 1928. He received the B.S. degree in electrical engineering, in 1951, from the National Taiwan University in China, and the M.S. and Ph.D. degrees from The Ohio State University, Columbus, in 1953 and 1955, respectively.

He worked on ionospheric physics and propagation at the Radio Wave Research Laboratories in Taipei, Formosa, from 1951 to 1952. Since 1952 he has been associated with the Radio Observatory, Department of Electrical Engineering, at The Ohio State University where he is now Assistant Director of the Radio Observatory. In 1955 he joined the faculty of The Ohio State University, where he is currently Associate Professor of Electrical Engineering and Associate Professor of Physics and Astronomy. He has been engaged in research in the fields of radio astronomy, upper atmosphere physics, electromagnetic theory, radio propagation, antennas, and low noise receiving systems.

Dr. Ko is a member of Sigma Xi, Pi Mu Epsilon, Eta Kappa Nu, the American Astronomical Society, and Commission V on Radio Astronomy of the U. S. National Committee of URSI. He is secretary of the Columbus chapter of the IRE Professional Groups on Antennas and Propagation and Microwave Theory and Techniques.

❖

Edward F. MacNichol, Jr., (A'46–M'55) was born in Toledo, Ohio, on October 24, 1918. He received the B.A. degree in physics from Princeton University, N. J., in 1941. His graduate study at the Johnson Foundation of the University of Pennsylvania, Philadelphia, in 1947 and 1948, was supported by a predoctoral fellowship of the American Cancer Society. He received the Ph.D. degree in biophysics from The Johns Hopkins University, Baltimore, Md., in 1952.

He was a staff member of the Radiation Laboratory, Massachusetts Institute of

Technology, Cambridge, Mass., from 1941 to 1946, and worked on precision radar ranging equipment, especially automatic range tracking, radar navigation and bombing devices, relay radar and an early missile-guiding system. He was also a contributor and volume editor of volumes 19 and 20 of the Radiation Laboratory Series. After receiving his doctorate, he remained at The Johns Hopkins University, and is now an Associate Professor of Biophysics. He is currently doing research on the physiology of the retina of the eye and is active in the design of instruments for biological research. He is also teaching physiology of vision and elementary electronics for biologists.

Dr. MacNichol is a member of the American Physical Society, the Biophysical Society and the American Physiological Society. He is also Editor of the IRE TRANSACTIONS ON BIO-MEDICAL ELECTRONICS.

❖

John W. Moore (SM'56) was born in Winston-Salem, N.C. on November 1, 1920. He received the Bachelor's degree in physics from Davidson College, Davidson, N. C., in 1941, and the Ph.D. degree in physics from the University of Virginia, Charlottesville, in 1945.

After a year at the RCA Laboratories, Princeton, N. J., he took a teaching and research position at the Medical College of Virginia, Richmond, where he made electrical impedance measurements on the human body. In 1950 he joined the Biophysics Division of the Naval Medical Research Institute, Bethesda, Md., and studied impedance and potentials in nerve and muscle membranes. In 1954 he transferred to the Biophysics Laboratory, NINDB, National Institutes of Health, Bethesda, where he did research on the electronic control of nerve membranes and separation of the resulting current into specific ionic components. In 1961 he joined the faculty of Duke University, Durham, N. C., in the Department of Physiology where he continues to work in this area of research.

Dr. Moore is a member of Sigma Xi, the American Physical Society, the Biophysical Society, Marine Biological Laboratory and AAAS.

❖

Morio Onoe (M'58) was born in Tokyo, Japan, on March 28, 1926. He received a degree in electrical engineering, in 1947, and the Ph.D. degree, in 1955, both from the University of Tokyo.

Since 1956 he has been an Assistant Professor of Electrical Engineering at the Institute of Industrial Science, University of Tokyo. From 1956 to 1958 he was a Visiting Scholar at Columbia University, New York, N. Y., under the Fulbright Exchange Scholar Program. Since 1961 he has been a member of the technical staff of Bell Telephone Laboratories, Whippany, N. J., on leave of absence from the University of Tokyo. He has been working in the fields of piezoelectric vibrators and ultrasonic delay lines.

Dr. Onoe is a member of the Acoustical Society of America, the Association for Computing Machinery, the Institute of Electrical Engineers of Japan, and the Institute of Electrical Communication Engineers of Japan.

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Mitsuo Sawabe was born in Narita, Chiba, Japan, on March 1, 1937. He received a degree in physics, in 1961, from Tokyo College of Science. A part of his graduate work on ultrasonic gyrators was done at the Institute of Industrial Science, University of Tokyo.

Since 1961 he has been with Nihon Radiator Company, Tokyo, Japan.

Robert L. Schoenfeld (A'52–M'57) was born in New York, N. Y., on April 1, 1920. He received the B.A. degree from New York University, in 1942, the B.S.E.E. degree from Columbia University, in 1944, the M.E.E. and D.E.E. degrees from Polytechnic Institute of Brooklyn, N. Y., in 1949 and 1956, respectively.

From 1944 to 1946, he served as a Lieutenant in the U. S. Army Signal Corps. After a brief period with a now defunct industrial electronics concern, he joined the staff of the College of Physicians and Surgeons, Columbia University, in 1947. He worked as a Research Associate in the Department of Neurology doing research and development in connection with the measurement and analysis of the electroencephalograph (EEG). In 1951 to 1952 he worked for the New York City Department of Hospitals at Francis Delafield Hospital, and from 1952 to 1957 at Sloan-Kettering Institute for Cancer Research in New York, doing instrumentation for radioactive counting and X-ray dosimetry. He completed his doctorate research at Sloan-Kettering Institute on the mathematical analysis of radioactive tracers in compartmental systems. Since 1946 he has taught at Polytechnic Institute of Brooklyn, part-time until 1952 and again after 1957, when he resigned as Associate Professor in the Department of Electrical Engineering to assume his present post. He is currently Head (together with J. P. Hervey) of the Laboratory of Electronics at the Rockefeller Institute, where he is also Assistant Professor of Electronic Engineering. His major interests encompass electronic instrumentation for biomedical applications, the use of engineering theory in biological measurements, models and data processing, and engineering education.

Dr. Schoenfeld served on the Editorial Board of the *Review of Scientific Instruments* from 1957 to 1960. He is currently a member and Vice Chairman of the Administrative Committee of the IRE PGBME. He is a member of the American Physical Society, Sigma Xi and Eta Kappa Nu.

Books

Extraction of Signals from Noise, by L. A. Wainstein and V. D. Zubakov

Published (1962) by Prentice-Hall Inc., Englewood Cliffs, N. J. 356 pages+6 index pages+xii pages+3 bibliography pages+17 appendix pages. Illus. 6¼×9¼. $14.00.

This is a welcome translation from the Russian, prepared by Richard A. Silverman. The English is clear and fluent. The authors present an extended mathematical treatment of theoretical topics and applications in detection theory. While it demands substantial professional maturity from its readers, the book will prove rewarding to careful study.

The book includes, in Part 1, "Statistical Theory of Optimum Linear Filters," a well-developed presentation of the concepts of the filtering of random processes leading to elaborations of the prediction theory of Wiener, matched filters, and related variants. Part 2, "Statistical Theory of Optimum Receivers," treats statistical detection problems in a radar context, which provides opportunities for considering the statistical properties not only of circuit noise, but also of the fluctuations in the signal that result from random processes in the reflecting target. Part 3, "Auxiliary Topics," provides discussions of the mathematical properties of the normal random process, modulation in several ways by such processes, and the process involved in reflection from complex targets. There are also five short appendices, commenting on certain critical points.

It will not seem remarkable to specialists in the field that these topics could form the bases for such a rich theoretical discussion as may be found in these pages, even though such topics as classical information theory and random walks have been left out. The specialists will want to own this book. Others will find the challenge this book presents more manageable if they are fortified with a good background in mathematical statistics and analysis.

As a textbook, the present volume would suffer by not providing that standard teaching aid, a set of exercises. Otherwise, it could form the basis for an advanced course at the graduate level.

Professor Silverman points out that translation provided an opportunity to correct many of the errors appearing in the Russian edition and that in this task he had the help of the authors. He was evidently quite successful; no errors were found, and only a few misprints were discovered. Some of these stemmed from the unfortunate choice of the vinculum to denote averages, since it requires scrupulous care to ensure that the proper terms are included. The review copy was moderately well manufactured. There was some evidence of carelessness in trimming the pages and a few of the pages were so lightly and unevenly inked as to be almost unreadable.

Many of the developments presented here have been worked out in the 40's and 50's in this country, but the reader may feel that the citations to the Western literature are quite limited. The translator has made some effort to correct these deficiencies. Of course, corresponding criticisms could be made of Western writing. These observations serve to emphasize the need for translations of important examples of technical literature from those languages which are still inaccessible to most of us. Professor Silverman is to be congratulated in his efforts in meeting a part of that need.

D. H. Cooper
University of Illinois
Urbana, Ill.

Elements of Infrared Technology, by P. W. Kruse, L. D. McGlauchlin, and R. B. McQuistan

Published (1962) by John Wiley and Sons, Inc., 440 Park Ave. S., New York 16, N. Y. 435 pages+10 pages+xxi pages+3 appendix pages+references by chapter. Illus. 6×9¼. $10.75.

Infrared radiation—its generation, propagation, and detection—has become a field of increasing importance during the last two decades, mainly because of its potential for military applications, and as a result of growing commercial and scientific needs. This book makes an excellent and needed contribution to the somewhat scarce literature on the subject of infrared technology. In their treatment, the authors follow an already well-established subdivision of the pertinent material.

After an introductory chapter on the history and present role of infrared, the book proceeds to develop the laws of thermal radiation, the geometrical relationships of photometry, and a brief outline of line and band spectra and of their origin in atoms and molecules. An enumeration of the practical sources for thermal radiation concludes the chapter on the generation of infrared.

This is followed by three chapters (3–5) on propagation phenomena and their practical implications: in Chapter 3, the derivation of metal and insulator optics by means of Maxwell's equations and a discussion of absorption, dispersion, interference, and diffraction; in Chapter 4, an outline of more practical considerations as they apply to windows, filters, and reflectors; and in Chapter 5, a description of atmospheric optics which contains a brief theory of molecular absorption and of various kinds of scattering.

The third and largest part of the book, chapters 6–10, is devoted to various aspects of detection. The authors approach the subject by giving a general treatment of semiconductor physics in Chapter 6, and of noise theory in Chapter 7. Only after this foundation is laid and the standards for performance and figures of merit have been defined, does the book proceed with the discussion of infrared sensing devices as such. Single element detectors and image converters are described together in the natural classification (not always consistently applied) into thermal effects, in which radiation is detected by a temperature increase, and photon effects, in which the electrons or ions of the detector experience quantized excitation. Chapter 9 presents a mathematical analysis of detector mechanisms and their performance. This treatment contains an application of noise theory to infrared detectors with particular emphasis on the ultimate limitation by photon noise. The last chapter, on comparative performance of elemental detectors, deals with the construction and figures of merit of single element detectors, most of which are now commercially available, at least in small quantities.

It will be apparent from this review that the book goes far more into theory than its title suggests. If this is unexpected, it is, to this reviewer at least, an agreeable surprise. The authors announce in the preface that a companion volume is now in preparation—"Elements of Infrared Technology: Systems and Applications." In view of the high quality of the first volume, one will await the arrival of the second with interest.

M. Garbuny
Westinghouse Research Labs.
Pittsburgh, Pa.

Superconductive Devices, by John W. Bremer

Published (1962) by McGraw-Hill Book Company, Inc., 330 West 42 St., New York 36, N. Y. 179 pages+4 index pages+xii pages+references by chapter. Illus. 6¼×9¼. $8.00.

This six-chapter book is unique in that it is the only book on the subject of superconductive devices. Three chapters on cryotrons, cryotron circuitry and computer memory devices follow an introductory chapter on the phenomenon of superconductivity. The remaining two chapters deal with other non-computer, superconductive devices and a summary of low-temperature technique.

The book should be of value to the engineer as an introduction to the subject of superconductive circuitry. Discussion of the important technical and economic problems encountered in making and using superconductive circuitry is lacking, because the book was written during a period when the field was undergoing its maximum rate of development.

The author has drawn rather heavily upon his own efforts for most of the cryotron and circuitry descriptive material with references to, but little discussion of, other published information. The 208 references cover a broad spectrum of the field of superconductivity and computer circuitry where it applies. The collection of widely scattered references to a specific subject in one volume is always a worthy contribution and the author has done a creditable job of this, thereby increasing the usefulness of the book immeasurably.

C. R. Smallman
Arthur D. Little, Inc.
Cambridge, Mass.

Communications in Space, by Orrin E. Dunlap, Jr.

Published (1962) by Harper and Brothers, 49 East 33 St., New York 16, N. Y. 170 pages+ix pages+5 index pages. Illus. 5¼×8¼. $4.95.

Into a small book of some 170 pages, Mr. Dunlap has packed an easy-to-read story of electronic communications from Maxwell to the 600-foot dish of the radio telescope at Sugar Grove, West Virginia.

The author, of course, has spent his life in communications, and in radio communications at that. A ham radio operator in 1912, then Radio Editor of the *New York Times* for 18 years, and then RCA's Vice President for advertising and press relations, author of twelve other books on radio, television and other phases of electronic communications, he knows the history he has written; and he knows how to write. Therefore this book is an excellent one for those who want to know how radio, television or radar all came about, who accomplished the wonders, and to some extent what is coming next.

The contributions of Maxwell, Hertz, Marconi, Kennelly and Heaviside, Paul Godley, Fessenden, De Forest, Frank Conrad, Armstrong, Alexanderson, Zworykin, and many others are cited and appraised, and the ways in which the individual inventions and discoveries were integrated into systems are described.

This should be a good book for the boy of today who will probably learn all about masers and lasers before he hears about coherers or audions. He will learn that electronics is still highly dynamic, that there are new worlds still to be conquered, inventions to be made; and it is well for him to learn too how limited has been the life of many of the marvels widely heralded and then rapidly replaced by new ones demanded by the Space Age.

KEITH HENNEY
Snowville, N. H.

Radio and Electronic Laboratory Handbook, by M. G. Scroggie

Published (1961) by Iliffe Books, Ltd., Dorset House, Stamford St., London, S. E. 1, England. 522 pages+15 index pages+xii pages. Illus. 5¼×8¾. 55s.

This book is the seventh edition of Mr. Scroggie's "Radio Laboratory Handbook," first published in 1938. The sixth edition was published in 1954, so the author had a lot of modernizing to do in preparing the seventh edition. He has done an excellent job, bringing in the new without sacrificing what has been tested by time. Many diagrams and discussions contain transistors and diodes, but the person who first looks only at the pictures must remember that the symbol used for a transistor is more like a tube symbol than is ours.

This book has value for all persons interested in making measurements through the use of electronic equipment. The theoretical aspects of measurements problems are discussed in terms of a solid understanding of electric circuit theory, but there are many practical points mentioned for which common sense provides adequate background for understanding.

The chapters cover fundamental principles of measurement, sources of power and signals, indicators, standards, composite apparatus, measurement of circuit parameters, signal measurements, measurement of equipment characteristics, and very high frequencies. In addition, it has chapters of general advice on such subjects as "The End and the Means," and "Dealing with Results." The last chapter is an 80-page handbook of reference information on material of interest to electronic engineers.

Many different types of measurement systems are described in the book, and the practical results are given, but for full comprehension the reader must fill in the steps in the development. References to the literature are given frequently, so the book can serve as a good starting point for detailed study of a measurement method.

A sound philosophy runs through the book. The reader constantly finds phraseology and short comments which enliven the writing and which bring things into good perspective. For example, in the section on receiver sensitivity we find: "Novices are sometimes perturbed by the thought that. . . . So if you, the reader, are worried, you know what you are!" Again, we find: "Suspect everything. The genuine experimenter has a permanently suspicious mind." And, we find the advice: "And if one intends to earn a living by technical work, it is important to know that in the long run reliability is more valuable than careless brilliance."

It is clear that the book is written by a man who is thoroughly familiar with electronic measurements. It shows the influence of the new developments in pulse systems, in the use of transistors. However, the author has dismissed the digital meter with only a page and some references. The next edition will probably have ten pages on this topic!

G. B. HOADLEY
North Carolina State College
Raleigh, N. C.

Forty Years of Radio Research, by G. C. Southworth

Published (1962) by Gordon and Breach Science Publishers, Inc., 150 Fifth Ave., New York 11, N. Y. 270 pages+xiii pages+4 index pages. Illus. 6¼×9¼. $6.50.

This volume comprises a most fascinating autobiographical account of the author's experiences and associations in the radio profession. The period extends from his college days in 1913 until his retirement from Bell Telephone Laboratories a few years ago. This account provides much historical background related to his monumental reference book, "Principles and Applications of Waveguide Transmission," published in 1950.

The first half of the story relates the author's experiences in the early days of radio until 1930. This reviewer found that narrative particularly interesting for its references to various workers within his own acquaintance, especially at the Radio Laboratory of the Bureau of Standards in Washington. The author participated also in radio activities at Grove City College and Yale University before joining Bell Telephone Laboratories for most of his life work.

The second half is devoted to waveguides and microwaves, the work for which Dr. Southworth is best known. This is a very human story of the satisfactions and frustrations that go together in any pioneering work. It is also a welcome introduction to his many colleagues and their part in the development of this branch of science and engineering. The beginning was the incidental observation of waveguide phenomena in a rectangular trough filled with water (used for its high dielectric constant). The climax was the accelerated development and intensive application of microwave techniques during World War II. The story closes a few years before Telstar, the first satellite serving as a microwave relay for intercontinental television.

The book is directed primarily to the author's fellow members of the profession. It will be most interesting to those who are working in the same or closely related fields, and especially to those who may be acquainted with many of the workers playing an active part in the story It contains sufficient detail to qualify as a historical reference. It is required reading for anyone studying the ways of pioneering in various fields of technology.

H. A. WHEELER
Wheeler Laboratories, Inc.
Great Neck, N. Y.

RECENT BOOKS

Bellman, R. E. and Dreyfus, S. E. *Applied Dynamic Programming.* Princeton University Press, Princeton, N. J. $8.50. Application of the theory of dynamic programming to the numerical solution of optimization problems arising in connection with satellites and space travel, the determination of trajectories, feedback control and servo-mechanism theory, and so on.

Evans, Walter, H. *Introduction to Electronics.* Prentice-Hall, Inc., Englewood Cliffs, N. J. $14.35. Written for an introductory course in electronics, both for the electrical engineering student and for others interested in electronic instrumentation. A background in circuit analysis is assumed.

Langmuir, D. B. and Hershberger, W. D., Eds. *Foundations of Future Electronics.* McGraw-Hill Book Co., Inc., 330 W. 42 St., New York 36, N. Y. Series of lectures given under the auspices of the University of California in 1959–1960, designed to introduce graduate engineers and non-specialists to some of the scientific background underlying modern electronic technology.

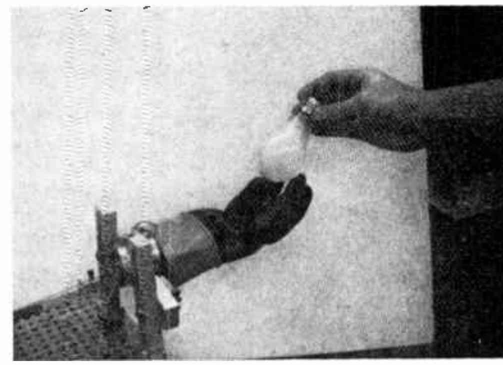
Osborn, J. A., Ed. *Magnetism and Magnetic Materials: Proceedings of the Seventh Conference.* Plenum Press, Inc., 227 W. 17 St., New York 11, N. Y. $12.50. Papers presented at the Conference on Magnetism and Magnetic Materials, Phoenix, Ariz., November 13–16, 1961.

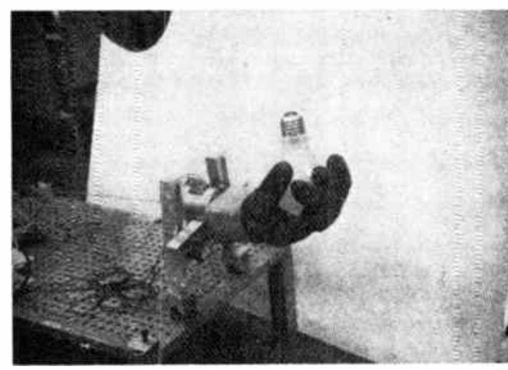
Scanning the Transactions

Electronics in Prosthetics. Although electrical engineers have a strong theoretical faith in the potentialities of their field in diverse areas, they are not immune to being fascinated when a startling or unique application is actually realized in the laboratory. A recent example is an extension of control theory into the physiological world. A rudimentary artificial hand has been constructed which in many ways resembles almost eerily its human counterpart and yet is elegantly simple in conception and construction.

The human hand can be considered as a control organ with two types of inputs: power to provide mechanical movement, and control signals to govern this movement. For a practical prosthetic device it is advantageous if the man can be relieved as much as possible of providing the control signals, that is, that the artificial hand act semi-automatically to external stimuli. The subject hand has sensing devices on the fingers, thumb, and palm which provide high-level electrical signals to a servo-motor so that there are no signal-to-noise ratio problems. The motor operates the jointed fingers and thumb by means of two flexible cables. Springs on each finger allow the fingers to adjust to the shape of the object grasped even though all four are operated by a single cable. The wrist joint consists of a ball bearing which allows angular motion of the hand in any direction for normally met angles. The hand is elastically suspended from the frame supporting the ball bearing with springs sufficiently strong so that, when an object is lifted, it acts as an ordinary scale with the point of rotation about the wrist.

The sensing devices are very simple and are first cousins to an ordinary carbon microphone. Each consists of a piece of artificial sponge filled with carbon granules with two electrodes and leads. The resulting pressure-sensitive element yields outputs from 4 to 10 volts when energized in its accompanying circuit. Besides in the fingers, thumb, and palm, these elements are used in the wrist to give an output proportional to the weight of the object held. With a minimum of transistorized electronics and two feedback loops, an experimental model of the hand has been found to operate quite well as illustrated by the two photographs showing it grasping a light bulb and adapting to its shape. The designers make no claim, but it is implied that a hale and hearty handshake would be returned in kind to those extending the same, while a limp, "cold fish" handshake would likewise be reciprocated. (R. Tomovic and G. Boni, "An Adaptive Artificial Hand," IRE Trans. on Automatic Control, April, 1962.)





Lasers have opened up vistas second to none in excitement for the communications engineer. Long-distance communications at prodigious information rates using optical frequencies may not be with us today, but with developments proceeding fast and furious in this field they are nearly as safe to predict for the near future as are taxes. Recall that it was only in late 1958 that Schawlow and Townes suggested the possibility of an optical maser, and that it was only the middle of 1960 that scientists at Hughes Aircraft Company announced that they had achieved the generation of coherent light.

To realize the full potential of coherent radiation in or near the optical spectrum, new modulation and detection techniques will have to be developed. But what can be accomplished in the way of long-distance digital communications with only slight improvements in the state of the art? As an example, consider an earth-moon link employing an optical-maser transmitter and a conventional photomultiplier detector in conjunction with a filter. The transmitter is located just outside the earth's atmosphere and the receiver is situated on the moon and is looking directly toward the earth. Assuming a CW optical-maser output of one watt and appropriate modulation techniques, calculations show that a postdetection signal-to-noise power ratio of approximately 135 at a bandwidth of 3 Mc obtains. This yields the capability of an information data rate of approximately 6 million bits per second in a binary system. The realization of this system is impeded by the facts that at present lasers with this output power are not available and the capability of modulating optical frequencies at megacycle bandwidths must be demonstrated. These obstacles are believed surmountable within the near future, and the tremendous potential advantages of digital communications at optical frequencies can soon be exploited. (Donald D. Matulka, "Application of LASERS to Digital Communications," IRE Trans. on Aerospace and Navigational Electronics, June, 1962.)

Modulation and Demodulation of Coherent Light are challenges which engineers and scientists are rising to meet. The achievement of extremely broad-band communications at optical frequencies awaits the fruits of their labors.

An FM phototube system has been proposed for demodulating light signals which are frequency modulated at microwave frequencies. The optical frequency modulation is converted into space modulation by using an optical dispersing element; the space modulation is then converted into transverse electron beam-wave excitation via a photocathode. A basic way of accomplishing this is to have the frequency-modulated light beam impinge upon a prism. The ray angle of the light emerging, and consequently the position of the light spot on the photocathode, will be correspondingly modulated. The configuration is such that the motion of the light spot on the photocathode will excite transverse position waves on the electron beam at the modulation frequency, or in microwave-tube parlance, synchronous cyclotron waves are set up. This transverse wave excitation can then be amplified and detected by a variety of means well known in the microwave-tube field.

The actual proposed system is somewhat more complex than this simplified description. Rather than an ordinary prism, two types of dispersing elements commonly used in interferometry are more suited for this purpose: the Michelson echelon and the Fabry-Perot etalon. A mathematical analysis of this demodulation process has been made and practical design formulas have been presented. Calculations indicate that it should not be difficult to demonstrate experimentally this type of FM light demodulation with strong output signals and large signal-to-noise ratio. (S. E. Harris and A. E. Siegman, "A Proposed FM Phototube for Demodulating Microwave-Frequency-Modulated Light Signals," IRE TRANS. ON ELECTRON DEVICES, July, 1962.)

Charge and Discharge. Probably every graduate of an electrical engineering curriculum remembers nostalgically his original exposure to the analysis of the charging and discharging of a capacitor. It was one of his first excursions beyond the bounds of steady-state Ohm's Law and nearly the first application of his newly acquired proficiency in the mysteries of differential equations. Armed with this exhaustive grasp of circuit theory and mathematics, he was ready to venture out and conquer forthwith, if not the world, at least any theoretical problem in the field of electrical engineering. Disenchantment came early. Yet without really articulating it, when faced with the apparently insoluble complexities of a new problem, who doesn't like to take a moment off to sketch the familiar RC circuit, facilely scribble down its differential equation, and restore self-confidence by quickly coming up with the reassuring exponential formula?

But life is never simple. For example, consider a capacitor consisting of a semiconductor diode biased in the reverse direction. Here we have a nonlinear capacitance whose value increases as the impressed voltage decreases. A general analysis of the charge and discharge of a junction-diode capacitor has been made. The resulting formulas for both the abrupt-junction and graded-junction cases differ significantly from that for a conventional capacitor: the current versus time curve deviates from the exponential since the charging and discharging proceed more slowly. The analysis has been utilized to study the operation of a cross-coupled multivibrator. Since a diode capacitor is constructed from the same material as are transistors, its exact behavior is of signal importance to the designer of molecular-electronics circuits. (H. C. Lin, "Step Response of Junction Capacitors," IRE TRANS. ON CIRCUIT THEORY, June, 1962.)

Pulse Measurement used to consist only of the drama played by the unhappy patient and his reassuring doctor, with a trusty pocket watch as a prop. Undoubtedly this ritual is still repeated innumerable times each day, but since the advent of radar and the computer the measurement of pulses has attained to new levels of importance and, incidentally, has taken on considerably different connotations.

A novel and useful technique has been developed for measuring pulse duration (electrical, not physiological) independent of pulse amplitude, an instrumentation requirement arising in many branches of technology. It is based soundly on a mathematical theorem and its corollary which show that the duration of a symmetrical pulse, regardless of its amplitude, may be determined from the measurement of two different types of time integrals. The mathematics is implemented in a practical transistorized circuit which is workable for pulse durations as narrow as 0.1 microsecond in the presence of additive Gaussian noise with a minimum signal-to-noise ratio of 18 db. (A. J. Rainal, "Integral Technique for Measuring Pulse Duration," IRE TRANS. ON INSTRUMENTATION, June, 1962.)

Electric Thrust Devices are conspicuous for the amount of attention they are receiving as promising means for the propulsion of space vehicles. Many desirable deep-space and interplanetary scientific missions are impossible with foreseeable chemical or nuclear power plants. Propulsion by electric means will be mandatory for such missions and will be preferable for many others. Feasible tasks for first-generation electrically-propelled spacecraft during this decade are Mars and Venus probes and orbiters; solar, Mercury, and Jupiter probes; and out-of-ecliptic probes. During the next decade, second-generation electric spacecraft will be assigned even more challenging missions. Detailed studies of the performance and power required for electrically-propelled vehicles have been made, and thrust and specific impulse requirements have been accurately defined. Three basic types of electric thrust devices are currently under development: the electrothermal, the electromagnetic (or magnetohydrodynamic), and the electrostatic. An example of the electrothermal type, the arc-jet motor, has reached an advanced state of development but is inherently unsatisfactory for interplanetary missions because of its low specific impulse. Magnetohydrodynamic or MHD devices of many types show promise, but many of the processes therein are complex and not fully understood. For high specific impulses, electrostatic propulsion devices or ion motors perform most efficiently at present. Several types of ion motors appear capable of development to the efficiencies and lifetimes required for interplanetary missions. Motors of the MHD type are considered to be a promising backup. (J. H. Molitor and D. G. Elliott, "Electric Thrust Device Requirements for Interplanetary Spacecraft," IRE TRANS. ON SPACE ELECTRONICS AND TELEMETRY, June, 1962.) Of course, if engineers are faced with exceptionally knotty technical problems for space travel, they can always take the following suggestion of the IRE STUDENT QUARTERLY's cartoonist:

"IT PROBABLY WON'T WORK, BUT WE'VE GOT TO TRY EVERY POSSIBILITY TO GET THERE FIRST."

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.

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Aerospace and Navigational Electronics

Vol. ANE-9, No. 2, June, 1962

The Editor Reports—(p. 52)

1962 Pioneer Award—(p. 53)

Frontispiece—Ludlow B. Hallman, Guest Editor—(p. 56)

Introduction to Aerospace Vehicular Digital Communications—Ludlow B. Hallman, Jr. (p. 57)

Digital Communications Between Aerospace Vehicles and Stations on the Ground—Klaus W. Otten (p. 58)

This tutorial paper presents an introduction to the problems of communications between aerospace vehicles and stations on the ground. The possible types of aerospace-to-ground communication links are described and classified according to the mode of signal propagation and according to obtainable range. The suitability of carrier frequencies throughout the electromagnetic radiation spectrum from very-low radio frequencies to light frequencies is discussed for the various types of links. Considerations of the information needed for the selection of modulation and demodulation techniques are presented, and factors which influence the practical value of digital coding are introduced. A discussion of the mutual influence of various parameters on the transmitter power requirement concludes the general survey.

Design of Reliable Long-Distance Air-to-Ground Communication Systems Intended for Operation Under Severe Multipath Propagation Disturbances—Klaus W. Otten (p. 67)

A survey of the principal methods of improving the reliability of long-range digital communication systems introduces the subject. Special modulation techniques as well as redundant transmission of pulses, redundant in time and in carrier frequency, are described as methods to counteract the effects of multipath distortion in connection with the proper special demodulation and diversity reception techniques.

It is shown that for an air-to-ground communication system, in which Doppler shift has to be considered in addition to the multipath distortion, high reliability can be obtained if quantized frequency modulation (QFM) and time redundancy are properly applied. To make the protection against multipath distortion most effective, the transmitted signal must have a wide frequency band. The optimum time diversity requires interleaving the redundant pulses and, therefore, requires complex logic circuitry for its instrumentation. The design characteristics of an "optimum" air-to-ground long-range digital communication system are described.

In the concluding section, it is indicated how one can apply the design principles for an "optimum" long-range communication system (as described) to the design of narrow bandwidth system with relatively simple instrumentation.

FAA Development in Aircraft Data Communications—Wayman R. Deal (p. 79)

The application of digital techniques to aeronautical air traffic control communications has been under study in the United States for more than fifteen years. During the past four years active development and evaluation programs have been conducted by the Federal Aviation Agency (FAA) on specific experimental equipment intended to provide guidance in the establishment of parameters for a universal system. Equipment has been developed using both low data rates and relatively high data rate approaches. Both systems utilize existing airborne voice communication equipment.

The major effort was spent on the development and evaluation of an experimental "Automatic Ground-Air Communication System" referred to as AGACS. This two-way time division multiplex system was designed primarily for operation at 750 bits per second, and provided both frequency-shift keyed carrier (FSK) and tone-shifted amplitude-modulation (FSK-AM) options.

The second equipment was developed under a program for "Analysis of Advanced Data Transmission Techniques" and utilized data rates in the range from approximately 30 to 200 bits per second. Audio tone-shifted (FSK-AM) Modulation was used.

The equipment developed is described and the results of the test evaluation is given. Guidelines are presented for future efforts.

Digital Selective Communications—Gerald A. Kious (p. 85)

Digital techniques promise to be a major factor in future aerospace communications systems. Voice is cumbersome, slow and redundant and does not make efficient use of the frequency spectrum. The digital approach permits the use of very narrow bandwidths, reduces redundancy and utilizes a constant loading factor. Selective addressing means that those not concerned with a message are not bothered by it, a very important factor in reducing distraction and fatigue of crew members in high performance aircraft. This paper describes terminal equipment developed for test and evaluation by the Air Force. The AN/URA-22 Control Monitor equipment provides for transmission of up to 456,976 different four-letter selective addresses, and recognition of individual, group and general calls, each of which can be any preassigned four-letter call. Remote switching is provided through the use of two mode characters, giving up to 676 posssible combinations. The fieldata code is used. Mode characters and addressor call are displayed at the receive end. The AN/URA-29 Digital Data equipment adds message capability, using both words and alpha-numeric characters to provide three word messages, followed by up to 48 alphanumeric characters in three lines of 16 characters each. Insertion is by simple push-button matrix and the message being composed is displayed. Received messages are displayed on receipt in plain language. The composed message is retained in storage until cleared. Operation of both equipments is explained in detail.

A Common System Approach for Aviation Data Communications—Evan L. Ragland (p. 91)

The present aviation voice communications link has evolved gradually with aviation. It was not originally designed to satisfy its current application as the primary link in military and civil control systems. As a result, as these systems have expanded the communications link has become less adequate. In the future, limitations of the communications link may become the controlling factor in the expansion of aviation. For this reason there is an urgent need for a clear definition of a general data communication subsystem approach for both civil and military application. For this definition to be complete all factors of aviation communication usage must be considered.

This paper carefully reviews 1) previous studies and development of civil and military data links, 2) published requirements of different aviation users, 3) evaluation of the future expansion of civil and military requirements, 4) current digital and communications technology, 5) transition from the present system to an improved system, and 6) the economic aspects of system design. From the considerations of this review a specification and approach to a general purpose aviation data communications subsystem is drawn. This approach provides for both voice and automatic communications and can be evolved gradually from the existing voice network.

Project Relay Digital Command System—S. H. Roth and T. A. Janes II (p. 100)

The National Aeronautics and Space Agency (NASA) has developed a coded message sequence consisting of discrete, pulse-duration-modulation (PDM) tone bursts for commanding satellites. The message consists of a sync pulse followed by some combination of six pulses, three each of zeros and ones. This code allows for 20 commands (the combination of six things taken three at a time).

This paper describes an equipment developed for the Relay satellite program which demodulates the tone bursts, converts the pulse-duration modulation into a binary code, and then decodes the message into twenty discrete commands. The demodulation and PDM-to-binary code conversion functions are accomplished by conventional transistorized circuitry which is described in a general manner. The circuitry for converting the code into command pulses is a novel utilization of magnetic circuitry and is described in detail. Magnetic cores are used to provide a shift register function and in addition perform the decoding, thus eliminating the conventional diode matrix usually employed for this function. A summary of the physical and electrical characteristics of the finished equipment is presented.

Application of LASERS to Digital Communications—Donald D. Matulka (p. 104)

Revolutionary developments taking place in the field of light generation show promise of providing a means for transmitting digital information over vast distances in space at extremely high rates. These developments stem from the generation of coherent light by devices called LASERS (Light Amplification by Stimulated Emission of Radiation).

This paper gives a brief description of LASER operation and discusses the applicability of the device to certain aerospace vehicular digital communications requirements. An earth-moon link is analyzed from the standpoint of beamwidth, power, and aiming requirements. It is shown that a system utilizing a coherent optical transmitter of less than 1 w and a conventional photodetector would be capable of transmitting digital information over this link at megacycle rates. The width band limitation here is imposed by lack of a suitable modulator rather than by any theoretical bound, and the power level is dictated by earth background noise and a usable transmitter and receiver beam angle.

Improvements which can be made on this rather simple system by increasing bandwidth and improving detection efficiency, tracking accuracy, and LASER techniques are pointed out. Curves showing basic limitations and interdependence of system parameters are plotted.

Automatic Height Transmission is One Step Closer—P. DeForrest McKeel (p. 110)

Use of automatic height transmission, the essential "third dimension" for use in the control of aircraft, has been brought closer by action taken recently at the Seventh Session of the International Civil Aviation Organization. The Session recommended adoption of technical standards that will: 1) update and improve the existing international standards for SSR, better known in the U. S. as the Air Traffic Control Radar Beacon System (ATCRBS); 2) provide specifications for 3-pulse sidelobe suppression (SLS); 3) provide standards for introduction of automatic height transmission; and 4) specify conditions under which SSR is to be implemented.

It was agreed further that the sensing element of the system should use a standard pressure setting of 1013.2 millibars and that it would be necessary to refer to the system as height transmission rather than altitude reporting.

Although a separate data link for height transmission is a possibility, the probable implementation data was considered too uncertain. Since air traffic is a heterogeneous mixture of civil and military aircraft, the existing SSR used by both civil and military aircraft was expanded to accommodate automatic height transmission features. It was agreed that SSR is needed to supplement the use of primary radar in air traffic control, but that each country would determine whether or not its use is required. The system specifications are appended to the paper.

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Automatic Control

VOL. AC-7, NO. 3, APRIL, 1962

What Has Become of Cybernetics?—Editorial—(p. 1)

The Issue in Brief—(p. 2)

An Adaptive Artificial Hand—R. Tomović and G. Boni (p. 3)

A review of the basic approaches to the artificial hand design is given. The importance of new ideas in this field using progress in automatic control theory is stressed. The most important feature of the new hand is the two level control. The movements of the hand can be controlled by signals produced by man as well as by external stimuli. For this purpose simple pressure sensitive transducers are placed in the hand to provide the control signals for reflexive-type movements. An adaptive control circuit is present for automatic weight adjustment. All these features are incorporated into the hand using simple and standard servocomponents. Use of appropriate feedback loops eliminates the need for complicated mechanical parts.

An Example of the Application of Dynamic Programming to the Design of Optimal Control Programs—Nicholas Kallay (p. 10)

The paper suggests a method for applying dynamic programming techniques to the design of optimal control programs. A step-by-step description of the method is given as it was applied to the problem of determining an optimal startup sequence for a satellite power plant.

In the present case the criterion of optimality is the amount of auxiliary energy required for the startup of the power plant in orbit. A phase space is defined in terms of pertinent power plant variables; the solution is then obtained by treating the problem as an N-stage decision problem consisting of selecting a sequence of phase space points such that the path so traced between the point representing the initial state of the plant and the point representing the final, operating state satisfies the criterion of optimality.

Numerical calculations were performed with the aid of the digital code described in the Appendix. The results of this parametric study indicate possible energy savings up to 25 per cent.

The paper is self-contained in the sense that very little if any contact with reactor technology or dynamic programming is required by the reader.

A Technique for the Adaptive Control of High-Order Systems—Edward A. Huber (p. 22)

A technique which is a modification of the model-reference method of adaptive control is developed to handle high-order systems. The transfer function of the model is the inverse of the desired transfer function of the closed-loop system insofar as it is practically realizable. Proper operation is obtained by adjusting the system until the poles of the closed-loop system are canceled by the zeros of the model.

Special filters are designed to aid in the detection of this cancellation. They have a narrow pulse for an impulse response and the dispersion of this pulse is used as the measure of error when an impulse is applied to the system. The criterion used for the design of the filter is that its impulse response should have a minimum second moment about the time of the maximum. The technique is applied to a fourth-order pitch-rate control system and the results of the computer simulation are given.

Stability and Convergence Properties of Model-Reference Adaptive Control Systems—Joseph J. Bongiorno, Jr. (p. 30)

An approach to adaptive control system design which does not involve a direct measurement or identification of the variable process parameters is investigated. Convergence-time relationships and sufficient conditions for system stability are developed when the process is characterized by a variable gain or variable time-constant, and fixed dynamics of any order. The adaptive capability is achieved by employing a model as a reference element and by means of appropriate adaptive circuitry. The adaptive circuitry is simple, easily instrumented, and does not require successive differentiations; the primary function of the adaptive circuitry is one of integration.

Verification of Aizerman's Conjecture for a Class of Third-Order Systems—A. R. Bergen and I. J. Williams (p. 42)

The second method of Lyapunov is used to validate Aizerman's conjecture for the class of third-order nonlinear control systems described by the following differential equation:

$$\dddot{e} + a_2\ddot{e} + a_1\dot{e} + a_0e + f(e) = 0.$$

In this case, the stability of the nonlinear system may be inferred by considering an associated linear system in which the nonlinear function $f(e)$ is replaced by ke. If the linear system is asymptotically stable for $k_1 < k < k_2$, then the nonlinear system will be asymptotically stable in-the-large for any $f(e)$ for which

$$k_1 < \frac{f(e)}{e} < k_2.$$

The Lyapunov function used to prove this result is determined in a straightforward manner by considering the physical behavior of the system at the extreme points of the allowable range of k.

Design of Multiple-Loop Feedback Control Systems—Isaac M. Horowitz (p. 47)

A single-loop design is basic for a two-degree-of-freedom plant, and it is theoretically able to achieve any desired insensitivity to plant variations or rejection of disturbances, if the plant is minimum-phase. In exacting feedback problems where the parameter variations or disturbances are large, the resulting single-loop transmission may require a larger gain and bandwidth than that of the plant. The added feedback compensation networks then have rising frequency characteristics and make the system very sensitive to HF noise in the feedback path. It is shown how a multiple-loop design permits the attainment of the same benefits of feedback, but with considerably less sensitivity to the HF noise. The basic problem is how to divide up the feedback burden most efficiently among the various loop transmission functions. Detailed procedures for this purpose are presented in the paper. The treatment is restricted to cascade-type plants.

On Errors Introduced by Combined Sampling and Quantization—Jacob Katzenelson (p. 58)

In digital computations, errors resulting from sampling and amplitude quantization (round off) are unavoidable. This work evaluates the mean-square error caused by sampling and quantization at the output of a linear network which contains a single quantizer. A detailed answer is given to the question, "Given quantized samples of a signal which is a sample function of a random process, what is the optimum linear filter for recovering the signal from its samples?" This filter is determined and its characteristics are summarized graphically for a specific example. A comparison with the conventional hold circuit shows that the optimum filter is much better if high accuracy is required and quantization is coarse. The difference in performance between the two filters is small when the accuracy requirement is low and the quantization is fine.

Also included as Appendix V is a survey of the general quantization errors problem, as it appears in the areas of digital computation and numerical analysis, and a study of multi-quantizer networks. It is found that extension of the method to networks which contain more than one quantizer is impractical, if not impossible.

A Contribution to Root-Locus Techniques for the Analysis of Transient Responses—B. J. Matkowsky and A. H. Zemanian (p. 69)

A method involving root-locus techniques is developed by which one can analyze transient responses by ascertaining bounds on them. In particular, the following questions are considered: 1) Given a rational system function $W(s)$, can another rational $M(s)$ (of simpler form) be constructed such that, for sufficiently large values of the constant multiplier B of $M(s)$, the corresponding transient responses satisfy the condition, $m(t) \geq w(t)$? 2) If so, can a range of values for B be determined for which the same condition holds?

Necessary conditions for an affirmative answer to question 1) are first developed and then sufficient conditions are obtained. Then, a general method using root-locus techniques is developed for answering question 2). Certain special cases are studied in detail and necessary and sufficient conditions are obtained, thus leading to the best possible bound for the given form of $M(s)$. Finally, a number of examples are given.

Circuit Theory

Vol. CT-9, No. 2, June, 1962

Step Response of Junction Capacitors—H. C. Lin (p. 106)

The large signal step response of a junction capacitor has been analyzed. Both abrupt and graded junctions have been considered. The analysis covers the charging and discharging of such capacitors through a linear resistance and through a battery. Expressions for the various responses have been derived. The results show that the charging and discharging waveforms are quite different as compared with the simple exponential rise and decay of linear capacitor. The analysis is utilized to study the operation of a cross-coupled multivibrator.

A Transistor Oscillator Frequency Stability Study—Robert Spence (p. 110)

The problem of oscillator frequency instability arising from environmentally induced changes in the active element is treated with respect to a particular transistor oscillator. Variations in frequency are related to those changes in collector capacitance and cutoff frequency which are due to temperature and supply voltage variations. Measurements show that the analysis is a useful basis for the estimation of frequency stability, and indicate correct identification of the major sources of instability. Consideration of the nonlinear aspects of operation is used to simplify the analysis and design of the oscillator. A useful order of frequency stability can be realized with the oscillator.

Impedance Transformations of Cascaded Active or Passive Identical Twoports—S. Eldring and R. A. Johnson (p. 116)

The number of iterative impedances for any active or passive twoport is one, two or infinite, the first and last cases being of only limited significance and application. For the second case, a previous article determined which of the two impedances (called stable) was ultimately approached by the sequence of input impedances to cascades of identical networks (active or passive) with ever-increasing numbers of sections. It was also noted that this convergence was in general nonmonotonic.

This article provides a solution to the problem of determining the impedance values which when used as terminations for such a cascade, will generate a sequence of input impedances which steadily approaches the stable iterative impedance. To facilitate the proofs involved, it also presents a resume of the relatively obscure geometric interpretation, the isometric circle method of construction, and the classification of the underlying bilinear transformations.

The theorems presented could have important application in the estimation of error in approximation problems involving artificial lines and filters working into varying terminating impedances and even, perhaps, over wide frequency ranges. Secondary results include 1) the passive iterative impedance for a cascade of passive networks is the stable one when the other is active, and 2) conditions on, and examples of, networks which generate periodic sequences of input impedances.

On Limitations of Broad-Band Impedance Matching without Transformers—S. Plotkin and N. Nahi (p. 125)

The basic work of Fano on the limitations for the synthesis of broad-band matching networks is extended to include some relations for determining resistance transformation ratio possible in the band-pass case. Fano's results are used to determine the load (or source) reactance possible and then to synthesize an optimum network. The relationships between the reflection coefficient, order of approximation, and the maximum theoretical resistance ratios are given for a number of cutoff-frequency ratios. These results for the band-pass case include absorption of ideal transformers with maximum turns ratios such that transformerless networks result. Appendixes include a design procedure, necessary polynomials for obtaining the immittance function up to an $n=6$ approximation (*i.e.*, a basic 6-section matching network), and one example using the results of this paper.

The *Z*-Matrix Parameters of Tapered Transmission Lines—J. L. Ekstrom (p. 132)

This paper presents an improved method of computing the *Z*-matrix parameters of 2-port tapered transmission lines. Linear second-order differential equations, in which the tapering information is contained in one variable coefficient, are derived for the transfer immittances; from the solutions of these the driving point impedances are found by differentiation and algebraic substitution techniques. As an example, the *Z*-matrix parameters for Jacobs' "generalized exponential" line are found.

Network Representation of Exponential Transmission Lines—V. Ramachandran (p. 136)

The purpose of the paper is to show that by the use of the rational fraction expansion of transcendental functions, the exponential transmission line can be represented at all frequencies by networks comprising lumped elements. The open-circuited and the short-circuited exponentiall lines are first considered; it is shown that there are three forms to represent each of these. In the cases of the open-circuited divergent line and the short-circuited convergent line, all the three forms yield physically realisable elements. But, in the cases of the open-circuited convergent line and the short-circuited divergent line, two forms can be physically realised when the impedance transformation ratio lies below 7.389 and only one form is realisable when it exceeds 7.389. It is shown that the exponential transmission line can be synthesized either by the use of the open-circuit impedance functions or by the short-circuit admittance functions, taking, in both cases, only the networks whose elements are physically realisable for any length of the line, thus constituting the two basic forms.

A Discussion on the Transient Analysis of Coaxial Cables Considering High-Frequency Losses—N. S. Nahman (p. 144)

The physical processes of HF loss are discussed. Two cases in transient analysis for coaxial cables are presented; the first considers skin effect in plated conductors while the second is an analysis based upon an attenuation approximation of the form (frequency)^m, $0<m<1$. A graphical transient analysis technique is described which allows one to easily analyze cables whose losses are due to a combination of physical processes. Generalized curves based upon the (frequency)^m approximation are presented by which the transient response of a cable can be rapidly evaluated for purposes of engineering design.

Approximation to a Specified Time Response—William C. Yengst (p. 152)

This paper presents a procedure by which specified data or a function of time $h^*(t)$ can be approximated by trigonometric and/or exponential functions of time $h(t)$ for which the Laplace transformations $H(s)$ are known and can be expressed in rational fraction form. The procedure is based on fitting $h^*(t)$ by an mth-order difference equation whose coefficients are determined by a least-squares technique. These coefficients are used directly to determine the poles of $H(s)$. The zeros of $H(s)$ are established by using the prescribed data or function $h^*(t)$ and the initial value theorem. The approximate function of time is obtained by taking the inverse Laplace transformation of $H(s)$. By this procedure not only is an approximation obtained for $h^*(t)$ in the time domain, but its transform is also found in rational fraction form suitable for realization as a driving point or transfer function. Furthermore, the least-squares technique used in determining most or all of the unknown parameters in this procedure tends to minimize the effect of random errors or noise present in the specified data.

Area Transforms—William M. Brown (p. 163)

The paper presents the theory of an operational calculus which is much more general than previously available theories. The theory is essentially all inclusive in that any operational analysis for which there is a multiplicative notion of transfer function for linear time-invariant systems is a special case of the general theory given here. The introduction reviews the foundations of operational analysis and then leads quite naturally to area transforms as the most general form. The general theory given here is conceptually very attractive in that it unites and extends all previous theories. Also, it is adequately general to include functions such as exp (t^2) in its domain of applicability. The basic idea of the theory is to express the functions of time as the superposition of exponential functions, exp (pt), with the superposition over the entire complex p plane rather than over only a line in the p plane as done with previously available forms of operational analysis (Fourier, Laplace, and Stieltjes transforms).

Simultaneous Flows Through a Communication Network—S. L. Hakimi (p. 169)

This paper presents a generalization of the results of Elias, Feinstein, and Shannon, and Ford and Fulkerson on the maximum rate of information flow through a communication network. The problem which is considered is the following: suppose a fixed rate of flow of information is being maintained between a pair of stations A and B of a communication network, then 1) what is the maximum rate of flow of information between another pair of stations C an d D of the same communication network, and 2) how can one allocate, among the channels, the original load on the communication network to obtain the maximal flow between stations C and D. It is shown that within certain determinable limits the sum of these two rates of flow remains a constant. A technique for attaining the maximal flow between stations C and D based upon the linear programming is described. A solution of the generalization of this problem to the case of k simultaneous flows is also presented.

Communication Networks with Simultaneous Flow Requirements—D. T. Tang (p. 176)

This paper considers the analysis and synthesis of communication networks with constant branch capacities, subject to a set of simultaneous flow requirements which are, in general, time-varying. Since the maxima of different flow requirements usually do not occur at the same time, savings in total capacity or total cost are possible through reassignments of flow paths.

A set of necessary and sufficient conditions for a communication network to satisfy a given set of simultaneous time-dependent flow requirements is obtained. This set of conditions, when applied to the case of constant simultaneous flow requirements, is shown to be more compact than those previously known. A synthesis procedure is then derived which enables one to obtain a communication network realization of minimum total branch capacity subject to the constraint that the network must assume a tree configuration.

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Electron Devices

Vol. ED-9, No. 4, July, 1962

A Proposed FM Phototube for Demodulating Microwave-Frequency-Modulated Light Signals—S. E. Harris and A. E. Siegman (p. 322)

A microwave phototube for demodulating frequency-modulated light signals is proposed. The demodulation is based upon the conversion of the frequency-modulated light into space-modulated light via an optical dispersing element. This space-modulated light is then incident on a photocathode where it is the source of transverse electron beam waves. A complete spectral analysis of the demodulation process is presented. It is shown that a quasi-steady-state viewpoint, *i.e.*, that of an optical signal with slowly varying frequency is permissible only if the optical resolution is sufficiently low. Design parameters for a phototube based on the use of a Michelson echelon are presented. A related scheme employing a Fabry-Perot etalon is also discussed.

Subharmonic Pumping of Parametric Amplifiers—Kenneth E. Mortenson (p. 329)

The purpose of this paper is to present in some detail the operation of a parametric amplifier pumped subharmonically as compared to being directly pumped. As considered here, subharmonic pumping does not involve harmonic pump power generation (external to the varactor) but the utilization of higher-order time-dependent capacitances to yield parametric amplification by employing, basically, only a three-frequency system.

The analysis given here is based on an evaluation of the Fourier series representation of the time-dependent capacitance resulting from large-signal ("hard") pumping of varactors. This evaluation indicates that significant values of higher-order time-dependent capacitances suitable for parametric amplification are obtained with relative pump swings in excess of about 90 per cent. Utilizing these higher-order time-dependent capacitances, the amplifier operation for various orders of subharmonic pumping is treated, including such factors as pump power requirements, gain, and noise figure. It is shown that, under certain conditions, less pump power is required to generate the same negative conductance than with direct fundamental pumping. Furthermore, for the same pump power and fundamental pump frequency, it is determined that significant improvements in amplifier noise figure are achieved by employing subharmonic pumping, provided varactor losses are small.

From the results obtained by both analysis and experiment, it is concluded that subharmonic pumping, even without harmonic power generation, is not only feasible but can be very useful up to C-band signal frequencies with existing varactors.

The DEMATRON—A New Crossed-Field Amplifier—G. E. Pokorny, A. E. Kushnick, and J. F. Hull (p. 337)

A nonreentrant beam, distributed-emission, crossed-field, forward-wave amplifier, the DEMATRON, is described. The difficulties encountered by early experimenters in achieving gain in excess of 6 db in nonreentrant, crossed-field amplifiers are overcome in the DEMATRON by use of either an electron velocity taper or circuit velocity taper.

A crossed-field amplifier design theory is given which is based on the use of equivalent magnetrons, and which takes into account the need for velocity compensation. In practice, electron velocity compensation is accomplished by either changing the sole-anode spacing, or by varying the dc magnetic field or a combination of both.

Experiments with the DEMATRON have yielded gains in excess of 10 db over a 15 per cent bandwidth. Power levels between 300 and 500 kw have been achieved at an operating voltage of 25 kv.

The design theory has been experimentally shown to be quite satisfactory in the large-signal, saturated gain region of operation. However, the lack of adequate small-signal theory has thus far prevented full optimization of the velocity compensation.

Prebunched Beam Traveling-Wave Tube Studies—Allan J. Lichtenberg (p. 345)

A traveling-wave tube with a prebunched beam is found to have a considerably higher efficiency than the same tube without prebunching. For the particular tube tested the efficiency is increased from 20 to over 35 per cent at a gain of 8. Computer calculations using a discrete disk model give similar results. The beam is bunched tightly in energy at the position of highest efficiency, indicating that very high efficiency could be obtained with a depressed collector. Both current and velocity modulation are required for prebunching, and are obtained by means of a current grid followed by an inductively tuned velocity modulation cavity. The requirements of the current grid are not great so that operation should be possible at frequencies well above the normal operating region of a microwave triode. The operation of the tube is sensitive to the output match, and it appears to be difficult to obtain a good match with the beam on.

A Small-Signal Analysis of the Electron Cyclotron Backward-Wave Oscillator—K. K. Chow and R. H. Pantell (p. 351)

In an earlier report on the construction and performance of the electron cyclotron backward-wave oscillator, it was shown, through physical arguments, that in an unloaded waveguide supporting the dominant mode, an electron having transverse rotation at its cyclotron frequency will interact with RF fields of approximately equal frequency. This transverse motion will deliver energy to the RF E fields and interact with the RF H fields, thus producing longitudinal bunching. A small-signal analysis is presented in this paper.

With the use of the normal mode expansion analysis, the circuit equation is obtained by considering the normal mode in approximate synchronism with the beam. The RF current is computed by considering electron motion under the dc and circuit fields, but neglecting RF space-charge fields. Combining these equations leads to a sixth-order equation of propagation constants. Two waves are far from synchronism and are therefore neglected; the remaining four are two waves which originate from the "fast cyclotron waves" and two waves which originate from the forward and reflected circuit waves. The "fast cyclotron wave" so obtained has a different meaning from the usual definition and is discussed in detail. Theoretical start-oscillation current is found to depend critically on the reflection coefficient at the electron gun end. Proper adjustment of this parameter leads to excellent agreement between the theoretical and experimental start-oscillation currents.

Analysis of Tunnel-Diode Converter Performance—Donald J. Hanrahan (p. 358)

Three tunnel-diode converter circuits—the Storm and Shattuck circuit, a push-pull version, and the Marzolf circuit—are analyzed graphically to obtain waveforms for both inverter and dc converter operation. Simple expressions are found for diode efficiency in ideal dc converter operation. The efficiencies are the same except that the efficiency of the Marzolf circuit is reduced by the magnetizing current required for the square-loop core. However, the Marzolf circuit has a more nearly square waveform and would require less filtering for dc conversion. The results point up the importance of developing tunnel diodes with high peak-to-valley ratios for converter application.

The Noise Performance of Negative Conductance Amplifiers—Bernard C. DeLoach (p. 366)

The noise performance of negative conductance amplifiers with outputs via circulators or isolators is developed using a traveling-wave approach. Included is a treatment of a quasi-degenerate parametric amplifier. It is shown that this amplifier has severe limitations as a low-noise amplifier.

Correction to "Single-Transit, Large-Radius E-Type Devices"—W. M. Nunn, Jr., and J. E. Rowe (p. 372)

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Information Theory

Vol. IT-8, No. 4, July, 1962

On the Statistical Detection Problem for Multiple Signals—J. B. Thomas and J. K. Wolf (p. 273)

The problem of detecting signals in noise is reviewed for the multiple input model, where each of the inputs can contain one of the many possible signals. The detection procedure for this model becomes, in general, the testing of multiple hypotheses. Two detection criteria are examined for choosing among multiple hypotheses and it is found that, for both criteria, the decision is based on the likelihood functions for the various signals.

Systems for computing likelihood ratios are examined in detail for the multiple input case. A multidimensional matched filter is considered and its relationship to the likelihood ratios is shown. Optimum signals are determined for the two-hypotheses problem.

Learning Filters for Optimum Pattern Recognition—David Braverman (p. 279)

An optimum adaptive system is obtained for the identification of pattern samples which are the sum of a fixed unknown signal, determined by the pattern of the sample, plus Gaussian noise. The system learns the unknown signals from a set of pattern samples, called learning samples, which have been identified with absolute certainty. The adaptive system is optimum in the sense that it computes the *a posteriori* probability of each pattern, given the sample to be recognized and the learning samples.

The rate at which the probability of misrecognition of the learning system approaches the probability of misrecognition of the *a posteriori* probability computing system with *a priori* knowledge of the fixed signals is derived, for binary recognition, as a function of the number of learning samples.

The Analysis of Certain Nonlinear Feedback Systems with Random Inputs—H. E. Henry and P. M. Schulteiss (p. 284)

A method is developed for the determination of the probability density function of the output of a nonlinear feedback system whose input is a random voltage of known statistical properties. The method of analysis is based upon the establishment of a mathematical model of the feedback system in such a way that the output is a Markov process. The transition probabilities of the Markov process are determined from the open-loop nonlinear characteristics of the system. From this model, the closed-loop output probability density function can be determined by the solution of an integral equation or, equivalently, by the solution of a set of simultaneous linear equations. As a consequence of the properties of stationary Markov chains, the same result can also be obtained by a process of successive matrix squaring operations.

The method is then applied to a complex nonlinear feedback system, a frequency tracking loop whose function is to follow the center frequency of a narrow-band random signal in the presence of wide-band noise. In addition to the study of this system with a stationary input, a simple extension of the method is made which allows the effect of a particular time-variation of the input statistical properties to be studied. The results of a digital computer study of this system are presented and discussed.

Efficient Error-Limiting Variable-Length Codes—Peter G. Neumann (p. 291)

Error Probabilities for Equicorrelated *M*-ary Signals Under Phase-Coherent and Phase-Incoherent Reception—Albert H. Nuttall (p. 304)

Formulas for the error probabilities of equicorrelated M-ary signals under optimum phase-coherent and phase-incoherent reception are derived in the form of previously untabulated single and double integrals. These integrals are amenable to computer evaluation for arbitrary M.

Two modes of reception are considered. In the first, one of M equal energy equiprobable signals is known to be transmitted during a symbol interval of T seconds through a nonfading channel with additive white Gaussian noise. The receiver is assumed to be synchronized in time and frequency with the incoming signal, and reception is on a per-symbol basis. Furthermore, the cross-correlation coefficients between all the signals are equal. The probability of correct decision in both phase-coherent and phase-incoherent reception is derived exactly, as a function of the signal-to-noise ratio, the common cross-correlation coefficient, and the size of the signal set M.

In the second mode of reception the only difference is that a threshold is incorporated in the receiver. The probability of false detection and the probability of detection and correct decision are derived exactly for both phase-coherent and phase-incoherent reception as a function of the threshold level, the signal-to-noise ratio, the common cross-correlation coefficient, and the size of the signal set M.

The method of reduction of multiple integrals presented here can be generalized, and may find application in other statistical studies in which the Gaussian density form is encountered under an integral.

Statistical Properties of the Contours of Random Surfaces—Peter Swerling (p. 314)

A random surface is a sample function of a random process $\{f(x, y)\}$ depending on two real parameters. Examples of random surfaces would be: photographs or television pictures; topographic maps; atmospheric pressure charts; and the like.

A number of statistical properties of the contours of such surfaces are derived. An application of the results to the problem of obtaining bounds on the information content of quantized random surfaces is outlined.

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Instrumentation

Vol. I-11, No. 1, June, 1962

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Microwave Theory and Techniques

Vol. MTT-10, No. 4, July, 1962

Injection Locking of Klystron Oscillators—Richard C. Mackey (p. 228)

If certain criteria are met, a microwave oscillator may be synchronized by the injection of a controlling signal into the oscillator cavity. Synchronization is dependent upon oscillator circuit parameters, the ratio of injected power to oscillator power, and frequency difference between the free-running oscillator and the injection signal. Locking has been observed with injection signals 70 db below the oscillator output and 30-db ratios have been demonstrated to be easily realizable. Injection locking may be considered a form of amplification that permits taking advantage of the fact that microwave oscillators are smaller, lighter, less expensive and more efficient than amplifier devices.

The low-frequency theory of Adler is shown to describe accurately the locking phenomena in reflex klystron oscillators and the transient response is extended to determine limitations on the amplification of modulated signals. Experimental verification of the theory is shown for 180° phase modulation of the locking signal at rates up to 100 kc for a VA-201 klystron. Design relations and curves are presented and applications and improvements are discussed.

The Application of Hypercomplex Matrix Analysis to Variable Parameter Networks—S. Krongelb, J. J. McNichol, and N. Kroll (p. 236)

The hypercomplex matrix methods developed to treat variable parameter elements are reviewed. The application of these techniques to the linear analysis of networks of variable parameter elements is demonstrated by considering a specific problem. A network containing two resonated variable capacitors separated by one-eighth wavelength of transmission line is first considered by the phase dependent admittance method. A partial treatment of the subharmonic case is given by this method to provide a physically plausible understanding of the network behavior. The complete problem is treated by the hypercomplex matrix methods. The discussion of the results illustrates how the network properties are determined from the mathematical formalism. Calculated characteristics of the two-capacitor network are given for several values of circuit parameters.

Broad-band Directional Couplers—E. A. Marcatili and D. H. Ring (p. 251)

It is shown how to connect two identical hybrids to obtain a directional coupler of arbitrary power division that operates over a broader band than that of the components. The broad-banding technique is possible with a certain kind of hybrid that includes Riblet couplers, multihole hybrids, coaxial hybrids and semioptical hybrids, but excludes T hybrids and ring hybrids.

Riblet couplers have a geometry particularly adaptable to the broad-banding technique. Where the balance of one of these couplers is better than 1 db, the balance of the broad-band hybrid can be made better than 0.16 db.

The broad-banding technique is particularly effective in the case of the 100 per cent transfer directional coupler type of circuit used for band separation filters and radar duplexers. In the semioptical waveguide band-splitting filters the bandwidth can be increased from about one to about four octaves (35–75 kMc to 35–580 kMc).

Noise Output and Noise Figure of Biased Millimeter-Wave Detector Diodes—K. Ishii and A. L. Brault (p. 258)

The behavior of a dc biased millimeter-wave detector diode was investigated by theoretical analysis and experimental measurement. The results indicate that because of the nonlinearity of the diode, shot noise appearing across the diode increases with dc biasing. For the same reason conversion gain of the detector increases with bias. The increase in gain is faster than the increase in noise for a certain range of bias current. Thus the noise figure of the diode detector and its minimum detectable signal are decreased.

Microwave Variable Attenuators and Modulators Using *PIN* Diodes—J. K. Hunton and A. G. Ryals (p. 262)

The *PIN* diode is a double diffused junction with an intrinsic layer separating the P and N regions. At frequencies above 100 Mc, the diode ceases to be a rectifier because of carrier storage and transit time effects. Its shunt capacitance is quite small because of the separation of the P and N regions by the I layer. Conductivity of the I region can be varied by a dc bias current and the device becomes an electrically variable resistor which can be used for microwave attenuators and modulators up to frequencies as high as 20 Gc.

The *PIN* junctions are mounted on posts which are inserted in a 50-ohm strip transmission line as shunt elements, and a number of these elements, spaced a quarter wavelength apart at midband, are used to form an attenuator. At the appropriate bias current, yielding 50-ohm junction resistances, the diode elements are reactively compensated by choice of post dimensions so that they are effectively pure resistances, yielding an image attenuation of 4.2 db per element. Many elements can be used to attain any desired total attenuation and higher impedance end elements can be used to improve the SWR. Bandwidths of 4 to 1 with low SWR in both ON and OFF conditions are achievable. Maximum attenuation of 60 db, insertion loss of 1 db, and SWR of 1.5 are typical for a 12-diode attenuator and powers of the order of watts can be handled with negligible harmonic generation. When used as a pulse modulator, rise times of the order of 10 nsec are achievable.

Analysis of Waveguide Modes by Standing-Wave Pattern Measurements—H. B. Dave (p. 274)

A method of analyzing a multimode transmission system is described, which is based upon the measurement of mean-square electric field along a line parallel to the waveguide axis. Although the analysis given is for a rectangular waveguide, the method has the advantage that it can readily be adapted to all types of transmission structures.

Excitation of Surface Waves on a Unidirectionally Conducting Screen—S. R. Seshadri (p. 279)

The excitation of plane surface waves by a line source on a unidirectionally conducting 1) infinite and 2) semi-infinite screen is considered. The conditions for the existence of the surface wave and the optimum location of the line source for obtaining the highest efficiency of excitation is determined.

On a Class of Multiple-Line Directional Couplers—C. R. Boyd, Jr. (p. 287)

Multiple-line directional couplers that utilize only two linearly independent modes of propagation are possible, provided certain restrictions on the maximum coupling are not exceeded. This paper discusses a class of multiple-line directional couplers which may be considered as a generalization of the familiar double-stub four-port directional coupler. The basic design relations for the symmetrical case are developed, and frequency behavior is investigated. Experimental results for an L-band six-port hybrid junction are presented and compared with theoretical curves.

Limitations on maximum coupling and fabrication problems will probably confine the number of lines in practical circuits to a small value. Bandwidths of couplers of this class tend to become narrower for the small total coupling off the main line as the number of lines is increased; however, standard techniques for broadbanding are applicable. Experimental results on the six-port hybrid junction agreed well with theory, and the circuit proved to be relatively compact.

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Nuclear Science

VOL. NS-9, NO. 3, JUNE, 1962

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Recent Developments in Alkali Halide Scintillation Crystals—W. W. Managan (p. 1)

The intrinsic resolution or minimum line width attainable in NaI(Tl) spectrometers can now be set at 5–6% for 662 KeV gamma-rays from Cs-137. Work leading to this conclusion is reviewed.

Pressure effects and directional effects in sodium iodide which support the exciton energy transport theory (Seitz model) are reported. Also, decay time, temperature coefficient ($-0.12\%/°C$), and non-proportional scintillation response in NaI(Tl) are reported.

Studies of the CsI(Tl) emission spectra and decay time constants as functions of temperature and de/dx are reported. Separation of alpha from gamma-ray events by color discrimination as well as by pulse shape discrimination is described.

A statistical method of measuring decay time constants over a wide range of intensity during decay is presented in an appendix.

The Scintillation Bubble Chamber—Richard H. Milburn (p. 16)

An additive has been found which converts liquid propane into a liquid scintillator solution. This solution has been used in a bubble chamber to associate photographed interactions in the chamber with externally developed coincidence pulses. Techniques and applications are discussed.

Bubble chambers have proved themselves very capable of recording precise geometrical information concerning high energy interactions involving charged particles. Provided that such particles have ranges of more than a few millimeters in the chamber liquid, their directions of flight over a 4π solid angle, their ranges, and in favorable cases their magnetic curvatures and ionizations may all be determined from measurements of chamber photographs. Secondary interactions, multiple scattering, and knock-on electrons are also in general quite apparent. In so far as one may choose the chamber liquid to be hydrogen, deuterium, helium, hydrocarbon, or one containing heavy nuclei one can in many sorts of experiments also control the entire input to an observed high energy reaction in the chamber.

A major chamber limitation is in the time coordinate. A bubble chamber photographs in one, perhaps cluttered, picture everything that occurs over a period of the order of milliseconds, and it can do this at most only a few times a second, more commonly every couple of seconds. Moreover one must decide to expand the chamber and use up the relatively expensive opportunity to photograph before it can be determined whether an interesting particle entered or an interesting interaction occurred in the chamber. One solution is to take as many pictures as possible as rapidly as possible and to rely upon the scanning room to sort out the wheat and the chaff. For many types of event this introduces a new variable called "scanning efficiency" which, in its subtle dependence upon psychology and physiology is often difficult to evaluate. It would frequently be desirable to have a non-visual determination as to whether a given photograph contains an interesting, particularly a *rare* and interesting, interaction and, if so, where in the chamber it occurred. If the interesting event is the passage into the chamber of a rare charged particle amidst a heavy background (for example, high energy antiprotons in a negative-pion beam) such a particle may be electronically labeled in flight and its position of entry into a bubble chamber recorded by a counter hodoscope, as has been studied by Kadyk and by v. Dardel or in an end-on spark chamber, such as have been developed by groups at CERN and the Brookhaven Laboratory. However, the passage of a neutral particle into a bubble chamber is less easily recorded and it is an attack on this problem that we shall now discuss.

A neutron, photon, neutral meson or neutral hyperon can only make its presence known by an interaction which destroys the momentum state if not the identity, of the very particle being detected. If the ionizing products used to detect this interaction are of short range relative to the chamber dimensions, then a signal indicating their presence must be generated from within the chamber sensitive volume itself. If an electronic signal could be generated in this manner, and if a similar signal can be derived from a process that is outside the chamber and which produces the neutral particle, then the two events could be correlated using standard counting techniques. Provided that a single chamber photograph selected on the basis of such a correlation does not contain too many additional uncorrelated interactions, the desired event may be picked out on the basis of kinematics and other criteria. In any case the photograph in which the desired event occurs may be selected from among perhaps many other photographs containing only spurious interactions which might otherwise be confused with the real one. Besides this reduction in the "noise" in a set of bubble chamber photographs one could in principle utilize timing signals from the chamber volume to measure lifetimes or to determine the time-of-flight of the incident neutral or charged particle whose interaction is to be studied.

Response of NaI(Tl) to X-rays and Low Energy Gamma Rays—W. C. Kaiser, S. I. Baker, A. J. MacKay, and I. S. Sherman (p. 22)

A thorough investigations of the response of NaI(Tl) in the region below 50 kev was made using a continuous distribution of X-rays from a tungsten target as well as monoenergetic gamma rays from radioactive sources. The results of this investigation agree well with those of Engelkemeir. (The investigation also showed that below 20 kev NaI(Tl) is sensitive to surface treatment. Above 20 kev the response is independent of surface treatment as long as the surface is free of moisture. An electron response curve is shown which, when used with the decay scheme of the excited iodine atom, predicts the observed nonlinear response NaI(Tl) to low energy X- and γ-rays.

Studies of the Scintillation Process in CsI(Tl)—R. Gwin and R. B. Murray (p. 28)

This paper reports a series of experiments on the pulse-height response of CsI(Tl) crystals to monoenergetic gammas, protons, and alphas, in an effort to examine some predictions of a previously proposed model of the scintillation process. Analysis of the data to date indicates the following: The pulse height per unit energy for gammas is an energy-dependent function which reaches a maximum near 20 kev, and the shape of this function is sensibly independent of Tl concentration. The pulse height per unit energy for protons and alphas is found to be energy dependent throughout the energy range studied. For alphas this function passes through a distinct minimum near 1 Mev. Effects dependent on pulse clipping time are discussed.

Effect of Delta Rays on the Response of Inorganic Scintillators to Heavy Particles—R. B. Murray and A. Meyer (p. 33)

Experimental results from Yale indicate that the scintillation efficiency of inorganic crystals to heavy particles is not a function of dE/dx alone but is instead composed of a series of discrete functions, one for each incident particle. This paper presents an analysis of these experimental results with attention to the effect of delta rays produced by the primary particle. In this treatment the total light emitted per unit path length of the primary is the sum of two contributions: one from the saturation column of the primary particle, and one from those energetic secondary electrons (delta rays) which escape the saturation column and produce light with a high efficiency.

Resolution and Line Shape in Scintillation Counters—J. B. Prescott and P. S. Takhar (p. 36)

An experimental study has been made of some factors that determine the pulse height resolution of scintillation counter assemblies. A statistical model is developed from which an analytic expression for the ideal line shape is obtained. Excellent agreement is found with observations using an artificial light pulser.

A comparison is made between noncrystalline organic scintillators and a sodium iodide crystal of similar size. It is shown that for gamma-ray detection, an important contribution to line-width originates with variations in the light collection efficiency from different regions of the scintillator.

The Effect of Dislocation Density on the Energy Resolution of Sodium Iodide Scintillators—I. Cooke and L. D. Reed (p. 46)

Sodium Iodide activated with thallium was used as the detector in a scintillation counter. The scintillating crystals were stressed with the intention of introducing dislocations and observing whether or not there was any measurable effect on the scintillation energy resolution intrinsic to the crystal. The photopeak was fitted to a gaussian curve to determine accurately the full width at half maximum. The results suggest that dislocations have a significant and detectable effect on the resolution.

Scintillation Response of Alkali Iodides to Alpha Particles and Protons—J. W. Blue and D. C. Liu (p. 48)

The relative pulse height for all alkali iodide crystals, with and without activation, in response to alpha particles and protons was measured as a function of the incident particles energy at both liquid nitrogen and room temperatures. In all cases the response curve is linear for protons of an energy range from 1 to 10 Mev, linear for alpha particles of high energies, but clearly nonlinear generally below 15 Mev. The response curve of activated crystals at liquid nitrogen temperature is generally flatter than that at room temperature. Unactivated crystals have a unique response only at liquid nitrogen temperature. For all crystals the proton scintillation efficiency curve lies slightly below the linear portion of the alpha particles curve. A sharp change in dL/dE is observed in the region of $130 < dE/dx < 160$ kev/mg-cm^{-2}, indicating a threshold for a process which has a decreased probability of exciting luminescence.

Computation of Efficiencies of Organic Liquid Scintillators—Adam Heller (p. 52)

The results of the computation of the scintillation and fluorescence properties of 22 aromatic conjugated systems are presented. Some of these are expected to have superior properties to those in use at present. A new scintillation solute, 1,4-bis-(p-isopropylstyryl)-benzene, when used as a secondary solute in a p-terphenyl solution in xylene, has an efficiency higher than 80% on the anthracene scale.

Progress on Photomultiplier Tubes at EMI Electronics, Ltd.—J. Sharpe (p. 54)

Progress in Photomultiplier Tubes for Scintillation Counting and Nuclear Physics—G. Pietri (p. 62)

This paper is primarily concerned with the development of high speed photomultiplier tubes designed for nuclear physics research applications. The significant characteristics and parameters influencing the performance of high speed tubes are discussed with particular emphasis to rise time, transit time fluctuations, high gain and linearity throughout a wide range of anode currents.

The performance of the well-known types 56AVP and 58AVP are described and in addition, a new tube type XP1020 which provides many outstanding features with regard to high speed performance.

Finally, standard types for gamma ray spectrometry, low energy beta particle and soft X-ray detection are briefly discussed.

Some Recent Developments in Photomultipliers for Scintillation Counting—R. M. Matheson (p. 73)

A family of three photomultipliers designed for good pulse height resolution has been developed. Photomultipliers for scintillation counting under severe environmental conditions have been produced and additional types are under development. The introduction of new window materials and sealing techniques had led to tubes sensitive in the far ultraviolet. Several types giving improved high speed performance are under development.

What's New at DuMont—J. G. Koosman and H. Timan (p. 78)

Direction of multiplier phototube activity is described, indicating a variety of special products. Pulse height resolution with related parameters are reviewed, and the performance spread in production runs of types 6292 and 6363 is discussed. Characteristics of a new high gain tube type are presented in some detail.

New Developments in Photoemissive Tubes—F. W. Schenkel (p. 83)

A discussion on a variety of special types of photomultipliers shall be presented. Among these will be the CBS type CL-1090, 14-stage, high speed, low dark current type and the 5-inch ruggedized photomultiplier. Also to be presented are new developments in special type dynode multipliers and large area, high current photomultipliers. Recent developments in vaccuum type neutron detectors shall be discussed briefly.

Status Report on the Development of the ASCOP Photomultipliers—Jean-Pierre Causse (p. 90)

Slightly over a year ago, Electro-Mechanical Research, Inc., introduced commercially a new line of photomultiplier tubes derived from the work done at the Research Center of its parent company, Schlumberger Well Surveying Corporation in Ridgefield, Connecticut. The new phototubes are marketed under the trade name "ASCOP."

Individually manufactured and tested under very stringent specifications, they fulfill the need for high performance, high reliability photomultipliers for both scintillation counting and the detection of light. Thus far they have proven especially useful in difficult environmental conditions such as those encountered in space exploration. Several of them have orbited or are now in orbit around the earth without failure. We will review only briefly the principle of their construction, which was described at the last Symposium, in order to dwell in more detail on new developments.

Field-Enhanced Transmission Secondary Emission for High-Speed Counting—E. J. Sternglass and G. W. Goetze (p. 97)

The time-response of high-gain transmission-type dynodes using KCl in the form of a low density layer is investigated. Both the results using light-pulses and charged particles directly incident on these dynodes indicate no detectable time lag within the limits of the experimental method ($<0.5\times10^{-9}$ sec). The number of emitted secondaries under the impact of α-particles and electrons confirms the theoretical expectation that the yield is proportional to the rate of energy loss per unit distance, dE/dx. The results indicate that transmassive dynodes of the low density type with yields between 50–100 secondaries per incident 5 kev electron can be used both for high-speed electron multiplication and direct detection of nuclear particles without the use of scintillators.

Electron Multipliers Utilizing Continuous Strip Surfaces—W. C. Wiley and F. C. Hendee (p. 103)

Thin films of semiconducting materials with secondary electron emission ratios greater than one have been used for some years in the Bendix crossed electric and magnetic field electron multiplier for ion detection. In these devices the electric field to accelerate the electrons is established by the potential gradient created by external current flowing through the resistive strip. A low background counting rate, insensitivity to radiation in the visible and near ultraviolet, and stability on exposure to atmospheric gases are properties which permit this device to have interesting photon detection applications in the extreme ultraviolet in dynamic vacuum systems and outer space environment.

A new form of electron multiplier utilizing continuous thin film surfaces, but not requiring a magnetic field, has been developed. Small diameter tubes (a few mils I.D.) coated on the inside surface give gains up to 10^6. This "channel" multiplier has been operated singly as a miniature electron or photon detector, and in arrays in image intensifiers. The principle of operation and some applications of these channel amplifiers will be discussed.

The Further Development of a Transmission Type Image Intensifier by 20th Century Electronics, Ltd.—D. L. Emberson (p. 107)

A large number of transmission secondary emission image intensifiers to the basic design, described by Wilcock et al., have now been constructed and the performance data obtained from these tubes is summarised.

Based on this information a tube of improved performance has been designed incorporating a larger dynode area and most recently a multi-alkali photocathode. The preliminary performance data on a number of these tubes with respect to resolution and dark current is reported. The results of measurements of the operational fatigue of this type of tube are also described.

Transmission Secondary Emission Image Intensifiers—N. A. Slark and A. J. Woolgar (p. 115)

The tubes to be described are of the transmission secondary emission type. Electrons from a tri-alkali photo-cathode are accelerated and focused on to a series of dynodes where multiplication takes place by transmitted secondary emission. The electrons from the final dynode are then focused on to the output phosphor.

Photomultipler Single-Electron Statistics—Robert F. Tusting, Quentin A. Kerns and Harold K. Knudsen (p. 118)

Our measurements of the amplitude distribution of photomultiplier anode pulses due to the emission of single-electrons from the cathode consistently show a peak. It is significant that the peak position agrees with that of a calculated distribution based on a Poisson distribution of secondary electrons at each dynode. The integral distribution, obtained by counting single-electron pulses, tends to show a plateau.

In low-light-level counting applications, one can set the dscriminator so that a majority of photomultiplier single-electron pulses will be counted. Further increase in the sensitivity will eventually increase the noise rate faster than the counting efficiency.

The techniques for measuring photomultiplier single-electron statistics are useful for obtaining comparative collection efficiencies. By single-electron measurements, one can adjust focusing-electrode potentials to maximize overall collection efficiency.

It is believed that there is some correlation between the amplitude and time distributions. Further work is necessary to show the extent of the correlation.

The "State-of-the-Art" in Nuclear Particle Detectors—James W. Mayer (p. 124)

Significant progress has been made during the past year toward realizing the very great potential of semiconductor detectors in nuclear research; this paper will review this progress, together with brief reference to relevant earlier work. Representative fields where major advances have been recorded are those of electron and high energy heavier charged particle detection using Lithium drifted and other thick depletion layer devices, neutron detection studies, nuclear astrophysical research where the low mass, size, and power requirements are of paramount importance, focal plane studies with magnetic spectrographs involving multichannel semiconductor systems, nuclear reaction product isotope identification with both hybrid gas and semiconductor detectors and with all semiconductor systems, high efficiency particle detection with large area mosaic detectors, and fission and other studies in high ambient neutral radiation fields where the junctions have unique advantages. Further applications are suggested.

Nuclear Experimentation with Semiconductor Detectors—D. A. Bromley (p. 135)

The theoretical and experimental characteristics of semiconductor particle detectors have been discussed at several conferences. The purpose of this paper is to discuss some of the recent results obtained with conventional p-n and surface-barrier detectors as well as to indicate the performance of ion-drifted p-i-n and high-resistivity gallium-arsenide and silicon "conductivity" detectors. The operation of conventional devices in regard to resolution and stability measurements, the existence of a fission fragment energy defect, irradiation effects, the base resistivity variations is discussed. The response, resolution, and rise time of p-i-n detectors (depletion widths from 1 to 5 mm) are presented along with some of the factors influencing the characteristics of the ion-drift process. Work with high-resistivity gallium-arsenide and silicon detectors has shown the influence of recombination processes. The limitations and further improvements in these devices are indicated.

Properties of an $n^+i\ p^+$ Semiconductor De-

tector—F. P. Ziemba, G. Pelt, G. Ryan, L. Wang and R. Alexander (p. 155)

A semiconductor nuclear particle detector fabricated by alloying or diffusing p^+ layers and n^+ layers into opposite faces of a slab of very high resistivity p-type silicon results in a device with several advantages over the conventional semiconductor detector structure. In the n^+ i p^+ structure, the space charge region extends throughout the entire silicon wafer upon application of a few volts bias and since further increases in bias do not appreciably change the width of the space charge region, the capacitance approaches a constant value. The series resistance of such a structure is very low and the collection time is exceedingly short, limited only by the applied field and the high field mobility of the carriers. Experimental data on collection time versus applied bias for several different nuclear particles as well as the characteristics of the unit in a low noise detector-amplifier system are presented.

The Window Thickness of Diffused Junction Detectors—R. L. Williams and P. P. Webb (p. 160)

The detector window thicknesses for carrier gas diffusions have been found to be very close to $\frac{1}{2}$ the diffusion depths. This result is interpreted in terms of the concentration profile of the diffusant. In the surface region, where the impurity concentration is greater than 10^{18} per c.c., the carrier lifetime becomes so small that ionized charges recombine before they can diffuse to the junction.

With reduced surface concentrations, window thicknesses of less than $\frac{1}{2}$ the diffusion depth have been fabricated, but good voltage rating diodes with a window thickness of less than 0.1 micron (.025 mg/cm²) are yet to be realized. For all diffusion depths a non-noise resolution limit of devices has been observed which corresponds to approximately $\frac{1}{3}$ the energy loss of an incident particle in the window layer of diodes.

For paint-on diffusion detectors, the nominal 900°C − 10 minutes − 0.1 microns diffusion units have measured window thicknesses of about 0.3 microns. For the process studied it appears that an excessive amount of phosphorus is incorporated into the surface layer.

Considerations in the Development and Use of Very Thin Junction Counters Suitable for $\Delta E/\Delta x$ Detection—C. N. Inskeep, W. W. Eidson and R. A. LaSalle (p. 167)

Thin diffused-junction counters have been developed suitable for use as $\Delta E/\Delta x$ detectors. Counters ranging in thickness from 0.0005 inch to 0.0025 inch and in area up to 1.0 inch diameter have beeen fabricated by a process allowing a variety of counter shapes as well as good dimensional control. Tests with the 22-Mev alpha beam of the Indiana University cyclotron indicate satisfactory $\Delta E/\Delta x$ detection over a range of alpha particle energies from 22 Mev to below 7 Mev. Fabrication techniques will be discussed. Some results will be presented to illustrate the application of these $\Delta E/\Delta x$ detectors to nuclear spectroscopy.

Semiconductor Fast Neutron Detectors—G. Dearnaley, A. T. G. Ferguson and G. C. Morrison (p. 174)

The application of semiconductor junction detectors to spectroscopy and flux monitoring for fast neutrons is discussed in terms of energy resolution and efficiency. The most useful forms of counter appear to be the proton-recoil and He^3 filled detector arrangements, and the construction, performance and applications of such devices are described. A He^3 filled detector has given an energy resolution of 150 keV for neutrons of 2–4 MeV with an efficiency approaching 10^{-5}. Measurements are being made of the cross-sections for reactions induced in the silicon of the detectors by fast neutrons, an effect which limits the range of neutron energies for which the counters can be used. Radiation damage to detectors is briefly discussed.

Improved Techniques for Producing P^+-I-N^+ Diode Detectors—J. L. Blankenship and C. J. Borkowski (p. 181)

Techniques for achieving a thin dead layer exhibiting low sheet resistance on the N^+ side of a P^+-I-N^+ diode made by the lithium drift process have been developed. A controlled quantity of lithium was diffused through a 1- to 2-micron phosphorus doped layer on the silicon diode. The phosphorus doped layer provide low sheet resistance. Because most of the lithium diffused layer was drifted into the bulk material, dead layers of less than 7 microns thickness were achieved.

Detectors made by this technique have given 23 kev (fwhm) resolution for gamma rays and mono-energetic electrons at room temperature, limited by diode noise. Detectors cooled to 78°–195°K gave 6.5 kev resolution for Cs^{137} conversion electrons (625, 655 kev) and Pb^{207} x-rays (74, 90 kev). Detectors stored without bias voltage at room temperature did not change performance over a four-month period.

An analysis of the drift parameters has shown that the lithium drift rate depended upon the power dissipated in the diode during drift. An automatic control system was developed which allows the lithium drift operation to proceed at power dissipations in excess of 50 watts.

Pulse Shape Discrimination in P-N Junction Detectors—Herbert L. Funsten (p. 190)

A pulse shape discrimination circuit for solid state particle detectors has been developed that will distinguish between ionizing particles stopping completely within the junction region and those passing beyond. It is based upon the fact that the first kind of particles produce a fast rising pulse in the order of nanoseconds whereas the second kind have an added diffusion component in the microsecond region. A circuit description and representative spectra are given.

The Operation of Solid State Alpha Counters in Chemical Process Streams—L. Cathey and W. J. Jenkins (p. 193)

The alpha activity in a chemical process stream may be monitored by placing an etched N-type silicon wafer in the stream to operate as a surface barrier alpha counter. The counters are constructed by the Blankenship technique with the gold cover omitted. The bare silicon surface is immersed in the sample stream. Since the liquid of the stream forms the front contact to the silicon the stream conductivity must be high.

The detectors have been operated in solutions of the nitrate salts of thorium, uranium, neptunium, plutonium, and americium. The effect of solution conductivity and acidity has been investigated. The effects of aluminum and fluoride ions have been studied also.

The detector will cease to operate if the oxidation of the silicon surface is allowed to run to completion. The silicon dioxide layer will insulate the "P" inversion layer on the surface of the silicon from the liquid contact. If the bias is operated in the forward direction for short periods, the silicon dioxide insulator will be punctured and contact between the liquid and "P" inversion layer will be restored. A similar result can be achieved by illuminating the silicon surface for short periods to induce a photo-etch process. Both of these methods have been used to maintain detectors in a counting condition for periods of up to nine months.

B^{10} Diffused Junction Detectors in N-type Silicon—H. M. Mann and F. J. Janarek (p. 200)

Silicon diffused junction detectors were prepared by diffusion of boron-10 into N-type silicon of resistivity 1000–3800 ohm-cm at a temperature of 1000°C. Detector diameters ranged from 3 mm to 2 cm. With the junction edge exposed in air or in vacuum, these detectors were extremely unstable. The use of organic silanes and ordinary waxes as a coating on the exposed junction edge provided stability. For some detectors the treatment with silanes was followed by a decrease in leakage current during subsequent storage. For alpha particles a resolution width (FWHM) of 40–45 kev was obtained for 6 Mev alpha particles in a detector of diameter 3–4 mm. For 2-cm-diameter detectors the resolution width at this energy was 80–100 kev, but reduced to 40 kev for alpha particles at a higher energy of 9 Mev. The peak-to-valley ratio for fission fragments from Cf^{252} was constant and equal to 1.40 at a detector reverse voltage of 15 volts or higher. The use of boron-10 allows identification of the energies of the products of fission reactions that occur within the detector, although similar detectors may be prepared using ordinary boron. For the present detectors the energies of the reaction products from the reaction $n(B^{10}, Li^7)\alpha$ and their sum were resolved with a resolution width of approximately 250 kev, and with energy corresponding to the known energies to within 5%.

Response of Silicon Surface Barrier Detectors to Hydrogen Ions of Energies 25 to 250 kev—R. I. Ewing (p. 207)

Commercially produced silicon surface barrier counters have been exposed at room temperature to hydrogen ion beams varying in energy from 18 to 225 kev and the pulse spectra analyzed. The output has been found to be linear with ion energy. A dead layer equivalent to about 4 kev loss by a 100 kev proton is indicated, suggesting that the gold surface layer is the only dead region. The width of the pulse spectrum is found to be about 9 kev for protons independent of ion energy, and limited by system noise. Multi-atomic ions produce an output a few per cent below that expected of an equivalent number of protons of the same velocity. The width of the pulse spectrum is also found to be greater for the multi-atomic ions than for protons.

Gold-Doped Silicon Detectors for the Control of Bubble Chamber Photography—C. R. Sun (p. 211)

The properties of a gold-doped silicon detector have been studied with β-ray sources. The observed peak of 0.6 Mev from a Sr^{90}-Y^{90} source agrees well with the most probable energy loss in the detector, calculated for minimum ionizing electrons. Our ultimate aim is to use a matrix of 20×24 of such detectors to control our bubble chamber photography.

Silicon Surface Barrier Detectors With High Reverse Breakdown Voltages—R. J. Fox and C. J. Borkowski (p. 213)

Surface barrier detectors with essentially no dead layer and with depleted depths up to 1.5 mm have been achieved by the combination of an inverted edge and a guard ring. The ability of these devices to withstand high reverse bias voltages insures rapid collection of the charge carriers and consequently fast rising pulses even for thick depletion regions. The energy resolution was 9.2 kev for 1-Mev electrons and was 7.5 kev fwhm at 285°K for 0.6-Mev electrons.

Encapsulated Surface Barrier Particle Detectors, Some Methods and Techniques—Niels J. Hansen (p. 217)

Techniques are described with which inexperienced workers have been able to produce detectors with almost 100% yield. Results are presented of measurements made at 78°K and at 300°K.

Application of Silicon Semiconductor Detectors to Measurements on Monoenergetic Neutron Beams—M. G. Marcazzan, F. Merzari and F. Tonolini (p. 234)

The Status of the Scintillation Chamber—Martin L. Perl (p. 236)

Several successful experiments carried out with the scintillation chamber are summarized to show what can be done with the systems which can be built now. There have been no developments in scintillation chamber components in the last year which allow radical improvements of these present systems. Many of the high energy nuclear experiments envisaged for the scintillation chamber can be done better with spark chambers. As a result of these two factors, there are only a few areas where a scintillation chamber is now the best instrument to use. Examples of such areas are space physics experiments, high energy beam imaging, high energy gamma ray detection and particle decays.

Some Measurements of the Efficiency for Observing Photoelectrons in Image Intensifiers—J. R. Waters, G. T. Reynolds, D. B. Scarl and R. A. Zdanis (p. 239)

A method is described for measuring the efficiency with which photoelectrons can be detected in an image intensifier system. A weak light in shone onto the first cathode and the photoelectron current measured. The light is then attenuated by a large known factor and compared with the number of spots observed in a given time. A simple system was used to demonstrate this method; a maximum efficiency of 85% was achieved. The results of experiments with a filament scintillation chamber and a Cerenkov detector using image intensifiers are discussed and yield some information about the electron detection efficiency. A simple method of finding the detection threshold is also mentioned with two typical examples of its results.

Observation of the Focused Ring of Cerenkov Light from a Single Particle Using an Image Intensifier System—S. K. Poultney and J. R. Waters (p. 243)

An image intensifier system has been used to observe the Cerenkov light emitted by a single charged particle traversing a Lucite radiator. The light, which was emitted in a forward cone of half angle 48°, was focused by a lens into a 3.5-in. circle on the photocathode of an RCA C70035 single stage image intensifier. The intensifier chain also included a three stage RCA C73491 intensifier and an RCA C74036 single stage intensifier orthicon in a closed circuit television system. A well collimated beam of negative II-mesons of 820 Mev/c went through the Lucite radiator in the shape of a narrow cone. A circle of about 50 spots should have been seen if all the photoelectrons emitted by the light at the first cathode had been detected. The actual number seen varied between 4 and 6 depending on the gain of the system.

Spark Chambers—State of the Art—James W. Cronin (p. 247)

A review of the presentstate of spark chamber technology is presented. The properties of spark chambers are reviewed and a summary of recent developments is given. An effort is made to make the survey of recent developments as complete as possible. Emphasis is given to technical problems which remain to be solved.

Ionization Density Effects in Spark Chambers—E. Engels, D. Roth, J. Cronin, and M. Pyka (p. 256)

During the course of an experiment on the decay properties of Λ^0 particles we observed qualitatively an ionization effect. The sparks of the heavily ionizing proton were bright and had 100% gap efficiency. The sparks of the minimum ionizing pion are weak and have a low gap efficiency. These observations are studied quantitatively in this paper.

Spark Chamber for Electronic Data Retrieval—Michael J. Neumann and Herrick Sherrard (p. 259)

In preparation of an information retrieval system system for experiments in nuclear physics a digitized spark chamber has been constructed and operated. Magnetic tape has been used for direct recording. It is planned to provide a high speed buffer storage.

Reduction of Delay Between Particle Passage and Spark-Chamber Spark—Joachim Fischer and Gus T. Zorn (p. 261)

The delay time between the passage of an ionizing particle and the spark-chamber spark is one of the limiting factors in the time resolution of spark-chamber systems. The usual resolving time is of the order of 0.5 μsec. The delay time in photomultipliers, cables, coincidence mixers, high voltage pulse generators, spark gaps, spark chamber voltage rise, and spark formation is investigated. Elements of a system are described which result in a total delay of about 33 nsec, including the photomultiplier. This may permit a reduction in spark chamber memory time and time resolution by about one order of magnitude. The reduction in delays and rise times also decreases the displacement of tracks in spark chambers operating in a magnetic field.

Recent Developments in Multichannel Pulse-Height Analysis—R. L. Chase (p. 275)

The state of the pulse-height analyzing art will be reviewed with particular emphasis on the developments of the past two years. The discussion will include consideration of multidimensional instruments, calibration-stabilizng techniques and some of the auxiliary features that are becoming increasingly available on commercial instruments. Possible future developments with respect to resolving time and memory organization will be discussed briefly.

Survey of Nanosecond Techniques—Albert L. Whetstone (p. 280)

In their application to high energy physics, nanosecond electronic techniuqes offer the experimenter the possibility of accumulating large amounts of highly selected data in the presence of intense backgrounds of uninteresting events. In this paper selected examples of recent contributions to nanosecond electronics are reviewed. Attention is given in fast logic circuits, pulse measuring the transmission techniques, and to the new components which have made the circuit developments possible. A temerous look at some future prospects is attempted.

Pulse-Shape Discrimination—A Survey of Current Techniques—R. B. Owen (p. 285)

A feature of scintillation counters is their ability to provide information on the type of exciting particle from the characteristics of the scintillation decay. The current state of knowledge on the decay of organic and inorganic phosphors is reviewed, and the main features of the many circuits devised to utilize the effect are discussed.

Limitations to the discrimination technique may be expected at high rates due to pulse pile up. The degrees of discrimination possible to low particle energies appears to be worse than may be expected on the basis of the photoelectron information available. The implications of this are examined.

Recent Developments in Scintillation and Semiconductor Spectroscopy—R. L. Heath (p. 294)

The most important area of development in gamma-ray scintillation spectrometry at the present time is the adaptation of digital computer techniques to the analysis of pulse-height spectra. The use of multi-dimensional pulse analyzers capable of storing 256 spectra, each containing up to 256 channels of digital information, has greatly increased the need for automated data processing techniques. As a first step, logical adders and tape input-output systems have been added to modern analyzers to permit some reduction of data within the machine itself. The development of analytical techniques adaptable to machine programming, however, appears to represent the only reasonable solution of the problem. A series of computer programs are described which have been successfully applied to the reduction of pulse-height data using digital computers. The calculations carried out by these programs include: determination of pulse-height *vs* energy response of a NaI detector, calculation of coincidence sum spectra; and the analysis of complex pulse-height spectra to obtain energies and relative intensities of individual gamma rays or relative abundances from a mixture of radionuclides. Measurement problems and instrumental requirements are discussed in detail.

Nuclear Instrumentation for Scintillation and Semiconductor Spectroscopy—Thomas L. Emmer (p. 305)

This paper describes four "modular" nuclear instruments: a double delay-line linear amplifier, an antiwalk single channel analyzer, a fast-coincidence unit, and a biased amplifier and linear gate for semiconductor detectors.

Time Resolution of a Scintillation Counter System—Arthur E. Bjerke, Quentin A. Kerns and Thomas A. Nunamaker (p. 314)

Time resolution of a scintillation counter system composed of scintillator, light pipe, multiplier phototube, discriminator, and coincidence circuit is discussed. The discriminator and coincidence circuits can produce coincidence curves with edges as steep as 10 psec per decade. Multiplier phototubes are capable of similar resolution, providing they view sufficiently bright, well-defined light flashes.

In a time-of-flight measurement at the Berkeley 184-inch Cyclotron, the coincidence curve slope, measured from a counting rate of 50% to 5%, is 400 psec per decade. The light level corresponds to 2000 electrons reaching the first dynode of the multiplier phototube. A slope of 200 psec per decade would be expected if only the multiplier phototube and electronics were considered. Geometry and rate of light output of the scintillator are believed to limit system resolution at present.

Rejection of Gamma Background Radiation Pulses in Hornyak Buttons—A. De Volpi and K. G. Porges (p. 320)

The nature of pulses resulting from alpha, beta, gamma, and neutron irradiation of Hornyak buttons has been examined under various conditions, including the use of light filters and both fast and slow electronics. Tentative conclusions about the mechanism of their origin are reached. Pulses from Compton recoils originate in the matrix and light transmitting media of the Hornyak button and phototube. Such pulses are of short duration, being similar in shape to photomultiplier noise pulses, in contrast to the 70-nanosecond decay of the fastest component of ZnS(Ag) fluoresence.

A relatively simple, passive anticoncidence network is described which allows fast cancellation of the gamma-induced short pulses while passing the zinc sulfide scintillation with adequate efficiency. Rise time of the pulses emerging from the background suppression circuit is about 7 nsec. This network, in conjunction with a distributed amplifier, tunnel diode discriminator, and other fast electronic equipment is used to select fission neutrons in coincidence with fission events.

Fast Pulse Multiplication by Logarithmic Attenuators—C. H. Vincent and D. Kaine (p. 327)

When a thin (dE/dx) transmission counter is used in front of a thick (E) particle counter, so that the particle may be identified by the two signals obtained, a pulse voltage multiplier of moderate accuracy and input range, but high speed, is required to effect the identification. A simple circuit of a new type, requiring no special components, and referred to as a logarithmic attenuator, has been developed for this purpose. With pulses of a microsecond or more in duration, it gives an immediate output for identification.

The Application of Automatic Testing to

Complex Nuclear Physics Experiments—Frederick A. Kirsten (p. 333)

The complexity of the data-acquisition systems required for advanced nuclear physics experiments is increasing. The difficulties involves both in setting up these systems and in detecting failures or drifts in the associated electronic equipment increase rapidly with the complexity of the system. To alleviate these difficulties, some automatic test routines for checking a complete data-acquisition system from phototube to scaler or analyzer have been developed.

One technique involves the use of the nanosecond light pulsers described by Kerns. These are mounted so as to illuminate the scintillators. Relay matrices for routing the electrical triggers to the pulsers have been developed. The routing is programmed to sequentially activate various combinations of light pulsers, thereby simulating the nuclear events under investigation, as well as accidental events.

Control systems are provided to perform the programming, with either automatic or manually controlled sequencing.

Methods of checking or recording the results of the test routine are discussed. Two applications of this technique to actual experiments are described.

Design of a Scintillation Counter K^+ Detector for a Bubble Chamber—T. Bowen and R. F. Roth (p. 340)

A K^+ detector is being designed for the Princeton-Pennsylvania Accelerator Rapid Cycling 15" Hydrogen Bubble Chamber which will trigger the photography whenever a K^+ is produced in the chamber. For the reactions of interest within our energy range, the K^{++}'s produced in the chamber will come to rest in the magnet coils needed for the field in the chamber and in the absorbers which will be placed in the gap between the coils. Plastic scintillators will be sandwiched between coil pancakes and absorber layers, and viewed by 96 2" photomultiplier tubes. We expect to be able to detect all delayed pulses due to K^+ decay secondaries triggering at least two layers with delays greater than 7 nsec., which corresponds to 20 per cent of all K^+'s produced. The major design problems are connected with economical design of the magnet-scintillation counter system and detection of the pulses delayed only a few nanoseconds in the possible presence of sometimes larger "prompt" pulses in the same counters.

Liquid Scintillation Counter Design Parameters—Richard B. Frank (p. 345)

The development of a liquid scintillation counting system has led to an evaluation of several multiplier phototube types, the comparison of a number of counting chamber surfaces, and an investigation of thermoelectric photocathode cooling.

The results obtained, including the more important performance and electronic circuit characteristics of the system, are described.

A Nearly-Four-Pi Liquid Scintillation Counter for Bursts of Fast Neutrons—Robert J. Lanter and Daniel E. Bannerman (p. 352)

This liquid scintillation counter detects 14 mev neutrons by both proton-recoil and neutron-proton capture processes. It is designed to measure fast neutrons which occur in a single burst of a few microseconds duration, a burst which may contain anywhere from 100 to 100,000 neutrons. The counter is 36 inches in diameter, 36 inches long, and has an axial access cylinder 12 inches in diameter. It contains 475 liters of triethylbenzene, 2.5 grams per liter of p-terphenyl and 0.02 grams per liter of POPOP. Scintillations are detected by 20 Dumont Type 6364 multiplier phototubes. Four tubes are used as proton-recoil detectors and 16 are used as neutron-proton capture detectors. The counter was completed in January 1959, and has shown high reliability and only a relatively small change in sensitivity since it was first calibrated.

Nuclear Particle Spectrometers for Satellites and Space Probes—E. L. Hubbard (p. 357)

In its program for research into the nature of the cosmic ray and Van Allen radiation the University of Chicago (the Enrico Fermi Institute for Nuclear Studies and the Laboratories for Applied Sciences) has been several particle telescopes and spectrometers. These instruments have taken various forms from a combination of semi-proportional counters to a telescope that used a gold-silicon surface barrier diode in combination with a CsI(Tl) scintillator. The CsI(Tl) uses a photodiode as a light sensor. The various forms of the telescopes will be discussed. It should be pointed out that this paper is a review only of some of the detectors used in the cosmic ray program at the University of Chicago.

Practical Aspects of a Scintillation Spectrometer System for Use at High Altitudes—John Gilroy (p. 366)

This paper discusses a system designed for obtaining γ-spectra from high altitudes, using a light weight balloon-borne device transmitting pulse information to ground equipment employing a standard multichannel analyzer. The sensing unit is a thallium activated cesium iodide crystal, surrounded with pilot B scintillator for discrimination against charged particles, observed by a 6199 photomultiplier tube. Packaging arrangements to minimize temperature effects and prolong battery lifetime are described. The resolution and drift recorded during operation of the system compared favorably with those observed in the laboratory without telemetering.

A Double Gamma-Ray Spectrometer to Search for Positrons in Space—T. L. Cline, E. W. Hones, Jr., and P. Serlemitsos (p. 370)

The advantages of the scintillation technique allow the development of experiments which can be made to respond preferentially to chosen types of particles. One such series of experiments requires the development of detectors which will identify single positrons mixed in a relatively high flux of other radiations. Directional detectors of positrons of energy between a few ev and a few Mev have been designed to search for admixtures of these particles in the cosmic radiation near the top of the atmosphere, in the electron population of the trapped radiation zone and in the solar particle streams and plasma clouds in interplanetary space. These detectors, which are scheduled for flight on balloons, sounding rockets and the Eccentric Geophysical Observatory satellite, are described. The scintillation techniques involved are discussed, and the analog and digital data-recovery instrumentation for each experiment is also outlined.

A Scintillation Counter Telescope for Charge and Mass Identification of Primary Cosmic Rays—D. A. Bryant, G. H. Ludwig and F. B. McDonald (p. 376)

A cosmic ray telescope has been developed to study the charge and energy spectra of primary cosmic radiation. The two major objectives of this program are:

1. A determination of the amount of interstellar material through which primary cosmic rays have passed between their source and the vicinity of the earth. This amount of material can be deduced from the shape of the low energy spectra of the primary nuclei helium to oxygen measured at solar minimum.

2. A study of the rigidity dependences on the various forms of modulations of hydrogen and helium nuclei.

A secondary objective is a study of the charge and energy spectra of cosmic rays produced by the sun.

Detector for Low Energy Gamma-Ray Astronomy Experiment—K. J. Frost and E. D. Rothe (p. 381)

A detector developed specifically for rocket and satellite borne low energy gamma-ray astronomy experiments is described. The detector consists of a 2" long×1" diameter CsI(Tl) crystal viewed by a single photomultiplier tube together with a 8.75" long×5.75" diameter CsI(Tl) crystal viewed by four photomultiplier tubes. The small crystal is inserted into a well of the larger crystal, the output of the smaller crystal being run in anticoincidence with the large crystal. The output from the small crystal surviving the anticoincidence circuit is fed into a pulse height analyzer. This detector and its mode of operation provides a gamma-ray spectrometer which has: angular collimation, low background sensitivity, suppression of the compton continuum and relatively high photopeak efficiency. Data on the photopeak efficiency, angular response, and suppression of the compton continuum at various energies are presented. Reasons for using anticoincidence shielding rather than bulk shielding with lead to contend with background problems are discussed.

Use of Scintillation Counters for Space Radiation Measurements—R. V. Smith, J. B. Reagan and R. A. Alber (p. 386)

Scintillation detectors used in space applications must meet rigid criteria for high performance at reduced pressure, at variable temperatures and must withstand extremes of shock and vibration. A large number of tubes have been subjected to shock and vibration tests. Ruggedized tubes withstood the tests and showed little change in gain. Measured temperature coefficients of gain varied from $-.19\%$ to $-.53\%/°C$, from 25°C to 70°C and may be circuit compensated. Battery and inverter type high voltage supplies have shortcomings which have been overcome in a high frequency voltage inverter circuit which supplies potential to each dynode directly. A preamplifier is described which has a voltage gain of one, a high input impedance and wide dynamic range such that further amplification is not required. The techniques described have been tested in several space flights.

Application of Nuclear Semiconductor Detectors for Space Spectrometry—S. S. Friedland, F. P. Ziemba, R. M. Olson and H. DeLyser (p. 391)

The proton energy spectrum in cosmic rays, trapped radiation belts, and solar flares is of considerable interest in space physics studies. The development of the semiconductor nuclear particle detector and the lithium ion drifted semiconductor detector offers a convenient means for studying proton spectra over a wide range of energies. The small size, ruggedness, stability, energy resolution and low power requirements of the semiconductor detector make it compatible with the requirements on space vehicle instrumentation.

Various systems to perform proton spectrometry are to be described. These include—

1. A single detector with a discriminator and counting circuitry and a single detector with pulse height analysis.

2. A dual detector to remove the ambiguity of the double value of pulse height vs. energy.

3. An array of detectors each with a different thickness degrader to cover the additional energy ranges.

4. A stack of detectors in multi-coincidence forming a telescope.

A brief desciption of the principles of operation of the semiconductor detector, the lithium drift detector and the response of these devices to charge particles is included for completeness. The reader familiar with this material may advance to the following paragraphs.

High Energy Gamma Ray Satellite Experiment—C. P. Leavitt (p. 406)

In order to locate possible discrete sources of high energy cosmic rays as well as to gain information related to cosmic ray origins a detector of 100 to 1000 Mev gamma rays has been constructed for inclusion in the NASA Orbiting

Solar Observatory, Satellite S-17. Both direction and energy are determined by a Cerenkov coincidence arrangement together with a total absorption lead glass Cerenkov counter. Solid state pulsed light sources will be used to provide calibration during flight.

Lunar Composition by Scintillation Spectroscopy—M. A. Van Dilla, E. C. Anderson, A. E. Metzger, and R. L. Schuch (p. 405)

Information on the composition of the moon and its past history is being sought through measurements of the spectrum of the gamma rays emerging from the lunar surface. The detector is a 3×3 in. CsI-plastic scintillator phoswich feeding a 32-channel analyzer weighing 5 pounds. Energy range is to 3 Mev; in-flight calibration is provided by a small Co-57 and Hg-203 source fixed to the detector. Energy resolution is 12 ± 2 per cent at 0.66 Mev. The spectrometer, and indeed the entire spacecraft, is biologically sterile. The spectrometer aboard Ranger 3 (launched January 26, 1962; missed the moon by about 20,000 miles) transmitted back spectra currently being studied; Rangers 4 and 5 are repeats scheduled for later in 1962.

A New Technique for Recording Heavy Primary Cosmic Radiation and Nuclear Processes in Silver Chloride Single Crystals—Charles B. Childs and Lawrence Slifkin (p. 413)

A new technique for recording heavy primary cosmic radiation and nuclear processes has been developed through decoration of dislocations formed by these processes in large silver chloride single crystals. The crystals are prepared and exposed to cosmic radiation during high altitude balloon flights. The crystals are then treated according to the Haynes-Shockley photoelectric technique which results in silver collecting at the dislocations and thereby decorates the dislocations at room temperature within two hours. Since silver chloride is transparent to visible light, the decorated dislocations are observable with an optical microscope at a magnification of about 150–200. A one-to-one correspondence has been established between tracks in the crystals and those in photographic emulsions accompanying the crystals during balloon flights. Examples of heavy primary cosmic radiation and nuclear processes in crystals are shown.

Notice of PGNS Annual Meeting, 1962—(p. 415)

Prices of these Transactions—(p. 416)

Space Electronics and Telemetry

Vol. SET-8, No. 2, June, 1962

Foreward—Fred T. Sinnott—(p. 70)

Introduction to Special NASA Issue—Hugh L. Dryden—(p. 71)

Frontispiece—Hugh L. Dryden—(p. 72)

NASA Space Electronics and Telemetry Systems Development—R. D. Briskman (p. 73)

A Skin Tracking Radar Experiment Involving the COURIER Satellite—M. Easterling (p. 76)

Range measurements are a very powerful type of tracking data for spacecraft operating at interplanetary distances. As a step in the development of a technique for making such measurements, an experimental system was devised in which the technique was adapted to the radar tracking of artificial Earth satellites. The system was tried out by tracking the Echo balloon and, after modifications, proven by tracking the Courier satellite. This paper describes the system that was used and the experiments that were conducted. Evaluation of the range data was made by comparing it with angle and Doppler data. The comparison was made by computing an orbit from angle and Doppler data obtained on three successive passes. The range was computed from the orbit for comparison with the measured range. The ranges measured were between 1,900,000 and 2,500,000 meter, and the largest disagreement between the measured and computed range during these three passes was 36 meters which is only about twice the estimated uncertainty in the range measurement. The average disagreement was approximately 20 meters or one part in 10^5. At the present it is not possible to state what part of the disagreement should be attributed to the ranging system and what part should be attributed to the orbit computation. The direction to be taken by further investigation is given.

The 1961 JLP Venus Radar Experiment—W. K. Victor and R. Stevens (p. 84)

The feasibility of exploring the solar system by radar was demonstrated on March 10, 1961, when a radio signal was beamed at the planet Venus, and for the first time in history the return echo was detected within a few minutes. The JPL Venus radar experiment has resulted in 1) an improvement of the accuracy of the Astronomical Unit by more than two orders of magnitude, 2) a determination that the dielectric constant and apparent roughness of its surface material are not unlike surface materials commonly found on Earth, and 3) a determination of the rotation rate of Venus for its most probable axis of rotation. In addition, the experiment verified that reliable interplanetary UHF communication is possible over ranges of 50 to 75 million miles and that planetary radar observatories are both practical and useful. Many improvements in the state of the radar and communications art are noted.

Analysis of the Range and Range Rate Tracking System—F. O. Vonbun (p. 97)

The "*range and range rate*" $(r_j+\dot{r}_j)$ system in its very simplest form is described. In particular, the errors in position and velocity are treated using pessimistic values of the measured quantities r_j and $\dot{r}_j$. Thus, a realistic evaluation of tracking qualities can be made for different orbits over certain tracking stations. The range and range rate system briefly described in this paper is a high-precision tracking system.

Knowledge of the uncertainty in position δ_{ri} is important, but knowledge of the uncertainty of the velocity vector $\dot{\delta}_j$ is of the utmost importance. Thus the use of coherent Doppler measurements to determine the velocity has a great advantage over any pulsed system and, in addition, permits extremely narrow frequency bandwidths (in the order of 10 to 100 cps) to be employed, reducing the power requirements considerably.

The basis for using range r_j and range rate $\dot{r}_j$ only is the fact that r_j and $\dot{r}_j$ can be measured to very high precision, thus furnishing r and $\dot{r}$ with low errors. The nature of these errors is discussed.

Pulse-Frequency Modulation—R. W. Rochelle (p. 107)

Pulse-frequency modulation (PFM) has been successfully employed as the encoding technique in a number of small U. S. Earth satellites where reduction of power and weight are prime considerations. This paper is concerned with the introduction of some of the more basic principles of PFM. Rather than presenting a rigorous proof on the orthogonality of this type of modulation, a comparison is made to the better-known characteristics of coded binary sequences. It is shown that PFM with quantized frequencies has the same communication efficiency in the presence of additive white Gaussian noise as a corresponding set of coded binary sequences with an equal number of quantized levels.

Synchronization of Telemetry Codes—J. J. Stiffler (p. 112)

A well-known means of efficiently transmitting information over the continuous, white Gaussian channel involves the encoding of successive *blocks* of data into sequences of binary digits (called *code words*). Efficient decoding of these sequences in turn necessitates a knowledge of the instants in time at which one code word ends and the succeeding word begins.

This paper presents a method for obtaining this synchronization which neither decreases the channel capacity nor increases the complexity of the encoding equipment. The method is to select, from the many encodings which are equally good for purposes of synchronous operation, that encoding for which the maximum absolute value of the correlation ρ_0 between any code word and any sequence formed from the overlap of two code words is a minimum. Thus a large correlation is observed only in the synchrounous phase position. This technique is applied to an important class of block codes, the *binary orthogonal codes*. An algorithm for constructing these codes with the desired self-synchronizing properties is presented, and upper bounds on the value of ρ_0 are thereby established.

Design of PM Communications Systems—R. L. Choate and R. L. Sydnor (p. 117)

A direct approach to the design of PM communications systems utilizing phase-locked demodulators is developed, and the design procedure is illustrated by example.

In addition to the new design techniques, two other ideas are emphasized. First, the output signal-to-noise ratio (SNR) is defined as the *ratio* of the mean information power output to the total mean-square closed-loop phase error, which includes the effect of signal distortion on the design. Second, the safety margin is defined as the ratio of the total mean-square phase error diminished by the mean-square phase error due to signal distortion to the quantity representing the nominal or expected mean-square phase error due to additive noise.

Although the design techniques developed in this paper are applicable for information waveforms having a flat power spectrum, a similar approach can be used for sine-wave modulation. The equations representing the output SNR for sine-wave modulation are included.

An experimental PM communications system has been constructed, and the results of initial tests show good agreement with the predicted performance.

Automatic Data Processing—C. J. Creveling, A. G. Ferris, and C. M. Stout (p. 124)

The problems of reducing telemetry signals from scientific earth satellites are assessed and an account is given of the factors governing the establishment of an automatic processing sequence. Some of the special features of the steps in this sequence are detailed, and an indication is given of the evolution in philosophy which has taken place since processing experience has accumulated. Final mention is given to some of the problems awaiting solution together with some of the plans under development for solving them.

Telemetry Considerations for Large Space Vehicles—J. E. Rorex and W. O. Frost (p. 135)

This paper discusses some of the problem areas and considerations unique to large space-vehicle telemetry. Some significant characteristics of specific telemetry multiplexing and modulation techniques are pointed out. The contributions to real-time data requirements such as vehicle prelaunch monitoring and checkout that can be made by telemetry techniques are described. Finally, some observations are made on the direction of future research and development (R&D) effort in the telemetry field.

Self-Erecting Space Antennas—W. F. Croswell, M. C. Gilreath, and V. L. Vaughan, Jr. (p. 139)

A self-erecting technique for erectable space-vehicle antennas is proposed. Working models of Yagi disk antennas utilizing this new principle are described.

Saturn Telemetry Antennas—J. W. Harper and U. Mrazek (p. 143)

The size and complexity of the Saturn vehicle influences telemetry antenna design in several respects. This paper describes the problems associated with size and shape of the vehicle body, bandwidth and power requirements, spurious frequency generation, and exhaust flame effects. Signal strength measurements recorded during the first Saturn flight are presented and compared with precalculated values.

Receiving System Design for the Arraying of Independently Steerable Antennas—J. H. Shrader (p. 148)

This paper considers the problem of arraying independently steerable antennas for use in deep-space communications systems. It describes the basic design of a receiver capable of utilizing all the signal power received on a number of antennas without requiring that the signals be phase coherent. The results of this study indicate that, for very large aperture systems, the array approach offers both economical and technical advantages over the single-reflector approach.

The Application of the Cassegrainian Principle to Ground Antennas for Space Communications—P. Potter (p. 154)

In the last few years considerable interest has arisen in application of the Cassegrainian principle to paraboloidal antenna systems. In the case of large ground-based tracking antennas, it appears that this type of feed system can offer significant performance and operational advantages over conventional systems. For this application, however, special sidelobe requirements are imposed on the Cassegrainian system. The forward sidelobe distribution must be controlled to reduce the effect of solar noise interference, and the backlobe level must be controlled to reject blackbody radiation from the antenna environment. It is shown that these considerations are the major factor in choosing the feed system configuration. An experimental system utilizing an 85-ft antenna operating at 960 Mc is described. This system has an aperture efficiency of approximately 50 per cent and a measured zenith noise temperature of 9.5° K.

The *X-Y* Antenna Mount for Data Acquisition from Satellites—A. J. Rolinski, D. J. Carlson, and R. J. Coates (p. 159)

Earth-orbiting satellite programs demand optimum performance from automatic tracking antenna systems. Certain criteria such as maximum drive shaft rates and best satellite data transmission conditions are considered in the design of the antenna mount. This paper describes the advantages of an *X-Y* antenna mount for performing data acquisition and satellite tracking functions. Maximum shaft rates of two-axis mounts are compared under similar satellite pass conditions, and the relative advantages of using an *X-Y* mount are discussed. A discussion of several of the general design considerations of the servo-control system is presented, and a discussion relating the error constants to the output rates is given. A few salient features describing the advantages of the *X-Y* mount, from the servo drive system designer's viewpoint, are also presented. The paper concludes with a brief description of NASA's 85-ft parabolic *X-Y* antenna located at the Data Acquisition Facility at Gilmore Creek, Fairbanks, Alaska.

An Operational 960-Mc Maser System for Deep-Space Tracking Missions—T. Sato and C. T. Stelzried (p. 164)

An operational 960-Mc low-noise receiving system for use in deep-space tracking missions is described. A ruby-cavity maser, low-noise antenna, low-loss transmission line connecting the antenna to the maser and associated instrumentation all combine to yield reliable and low-noise performance. Results of tests on this system in preparation for the Ranger RA-3 lunar probe are presented. The system was successfully used during operations with RA-3. A minimum system temperature of 47°K has been achieved.

A Technique for the Measurement of the Power Spectra of Very Weak Signals—R. M. Goldstein (p. 170)

A technique is described for the measurement of power spectra which is designed to distinguish the spectrum of a weak signal which is masked by strong background noise. The method was applied during the Jet Propulsion Laboratory Venus radar experiment to study spectral characteristics of the echo. An analysis is given which describes the performance of the system in such an application.

Microwave Measurements of Steady-State and Decaying Plasmas—Perry W. Kuhns (p. 173)

Some results are given of an experimental study of the loss mechanism for electrons in plasmas. Steady-state and decaying plasmas were studied by means of microwave interferometers. Values of the electron-ion recombination coefficient for nitrogen and argon are given. The effect of water vapor as an impurity upon the electron-ion recombination rate of air is demonstrated. Data is also given on the transmission of microwaves through an ionized layer about a body in a supersonic gas stream and the effect upon transmission of water vapor addition to the stream.

Nuclear-Electric Spacecraft Concepts for Unmanned Planetary Exploration—Robert J. Beale (p. 178)

The utilization of advanced, high-powered nuclear-electric spacecraft in the near future for the unmanned scientific exploration of space will provide significant weight allowances and power levels for scientific payloads and communication equipment. The electronic system designer, presently faced with minimum weight and power limitations imposed by today's chemically-propelled spacecraft, must develop new techniques to take advantage of the greater capabilities promised by nuclear-electric spacecraft.

Power levels approaching 1 Mwe may be expected. Systems capable of employing this power must operate under an intense nuclear radiation flux at elevated temperatures for long time periods. Development of such equipment must be initiated at an early date.

Electric Thrust Device Requirements for Interplanetary Spacecraft—J. H. Molitor and D. G. Elliott (p. 183)

This paper discusses the requirements that must be met by electric thrust devices in order to be used with interplanetary spacecraft. Two missions, a Mars orbiter and a Jupiter capture, chosen as representative of the time periods following 1965 and 1970, respectively, are analyzed to determine the thrust and specific impulse requirements of an electric propulsion system. The state of the art of electric thrust devices is discussed, and it is concluded that, with expected advances, ion motors can meet all of the requirements of interplanetary missions, with magnetohydrodynamic motors a promising backup.

Potentialities of Electron Bombardment Ion Engines for Electric Propulsion—Daniel J. Kerrisk (p. 188)

This paper discusses the expected performance of electron bombardment ion sources when used for electrostatic propulsion. Two particular sources, the duoplasmatron and the Penning Ion source, are examined in some detail and suggestions are made on how their performance for application with electric thrust devices could be improved. It is concluded that electron bombardment sources may offer some advantages over surface contact engines if ways can be found to significantly improve their propellant utilization.

Contributors—(p. 194)

Ultrasonics Engineering

Vol. UE-9, No. 1, June, 1962

Equivalent Electrical Circuits of the Quartz Crystal Transducer for Analysis of Ultrasonic Systems—Yujiro Yamamoto (p. 1)

In this paper three equivalent electrical circuits for the quartz crystal transducer are synthesized, two by using operational amplifier technique and one by linear network synthesis technique. The analog is based on consideration of one-dimensional vibration of the transducer only. The circuits are valid for any uniform terminations, complex frequency analysis, and higher harmonic excitation of the quartz crystal transducer.

Transmission Characteristics of Longitudinal-Mode, Strip Delay Lines Having Asymmetrically Tapered Widths—Allen H. Meitzler (p. 6)

A novel form of ultrasonic delay line has been developed which has two characteristic electrical properties: 1) a delay that is approximately a linear function of frequency and 2) a band-pass loss characteristic which is centered on the linear region of the delay curve. The delay medium of the line is a thin strip having an asymmetrically tapered width. The novelty of the line lies in the fact that by arranging the transducers so that a longitudinal wave motion interacts with one minor surface as well as the two major surfaces of the strip, a band-pass loss characteristic is obtained which is independent of the transducers and terminating circuits. Thus the delay line structure alone is one having properties of both a mechanical filter and a dispersive delay line.

Discussion of Time Delay in Reference to Electrical Waves—E. Howard Young, Jr. (p. 13)

The time delay of an electrical signal through an ultrasonic delay line may be uniquely expressed by either of the two terms "phase delay" or "group delay." There have been devised numerous ways of determining both of these values for use in the field of ultrasonics. Many of the measuring systems do not determine these constants directly but arrive at a value by using what appears to be, in certain instances, a close approximation to the mathematical definition. The purpose of this paper is to discuss some of these measuring systems as to the degree of the approximation involved. In doing so, the measuring system itself is analyzed together with the type of phase-vs-frequency curve for the delay line under consideration.

The Depletion Layer Transducer—Donald L. White (p. 21)

The depletion layer transducer is an ultrasonic transducer for use at UHF and microwave frequencies. Its potential advantages are high efficiency, large bandwidth, and comparative simplicity in fabrication. The region which generates or detects the ultrasonic waves is a thin flat high-resistance depletion layer, such as a *p-n* junction or a rectifying metal to semiconductor contact, in an extrinsic piezoelectric semiconductor. When an ac voltage is applied to the material, the depletion layer behaves in a manner similar to an extremely thin piezoelectric crystal bonded to the conducting substrate. Since the depletion layer can be generated on the surface of the semiconductor, the problems of handling extremely thin piezoelectric plates are avoided. Depletion layer transducers have worked at frequencies as high as 1000 Mc. It is anticipated that improvements in circuit and fabrication techniques will greatly extend their frequency range and efficiency.

Contributors— (p. 28)

Abstracts and References

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The Index to the Abstracts and References published in the PROC. IRE from February, 1961 through January, 1962 is published by the PROC. IRE, June, 1962, Part II. It is also published by *Electronic Technology* and appears in the March, 1962, issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extension and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands or from The British Standards Institution, 2 Park Street, London, W. 1, England.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears each June and December at the end of the Abstracts and References section.

ACOUSTICS AND AUDIO FREQUENCIES

534.13:551.507.362.1 **2515**

Preliminary Investigation of the Equivalence of Acoustically and Mechanically Induced Vibrations—L. Marin and R. C. Kroeger. (*J. Acoust. Soc. Am.*, vol. 34, pp. 674–678; May, 1962.) Damage due to vibration in rockets may be prevented by designing equipment structurally based on an equivalence relation between acoustically and mechanically induced vibrations.

534.133–8+621.372.412 **2516**

Degenerate Oscillations and Doublet Splitting of the Natural Frequencies of Piezoresonators—Finagin. (See 2570.)

534.14 **2517**

Wave Excitation on a Plane Comb Structure—M. D. Khaskind. (*Akust. Z.*, vol. 7, no. 3, pp. 366–369; 1961.) A method developed earlier (3059 of 1960) gives a simple and effective form of solution for the problem of surface-wave excitation in the two-dimensional and the three-dimensional case.

534.231 **2518**

The Field of a Point Source in a Layered Inhomogeneous Medium Bounded by an Uneven Surface—Yu. P. Lysanov. (*Akust. Z.*, vol. 7, no. 3, pp. 320–323; 1961.) An integral expression for the field is derived by a perturbation method.

534.231:551.508 **2519**

Contribution to the Study of Acoustic Zones of Silence and the Distant Focusing of Sound—J. Ecollan., J. Hieblot, and Y. Rocard. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3605–3607; May 30, 1960.) The focusing of sound in the Sahara at a point 200-km distant from its source suggests reflection at an equivalent height of about 120 km. The possibility of extending measurements of wind velocity and air temperature to such an altitude by this method is considered.

534.24 **2520**

The Total Reflection of Waves in Media in Movement—A. Metz. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3792–3794; June 8, 1960.) Theory is developed to determine the conditions necessary for total reflection in a medium in which one part is moving relative to another. See also *C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3591–3592; May 30, 1960.

534.241 **2521**

A Mechanism of Acoustic Echo Formation—A. Freedman. (*Acustica*, vol. 12, no. 1, pp. 10–21; 1962.) The combination of "image pulse" and "creeping wave" mechanisms is believed to account for the main scattering phenomena from rigid convex bodies, the former mechanism being paramount outside the shadow region at small wavelengths, the latter mechanism predominating at large wavelengths.

534.26 **2522**

Diffraction of a Cylindrical Sound Wave at a Cylinder—E. L. Shenderov. (*Akus. Z.*, vol. 7, no. 3, pp. 370–374; 1961.) The axis of the abstacle is displaced relative to the wave axis. Field calculations are compared with experimental data.

534.283–8 **2523**

Damping of Ultrasound in Metals, Carbon Steels and Nickel—P. J. Gellings. (*Acustica*, vol. 12, no. 1, pp. 29–32; 1962.) The resonant reverberation method is described. The results obtained are given together with a qualitative theoretical interpretation.

534.5 **2524**

Power Flow between Linearly Coupled Oscillators—R. H. Lyon and G. Maidanik. (*J. Acoust. Soc. Am.*, vol. 34, pp. 623–639; May, 1962.) The power flow between two independently and randomly excited harmonic oscillators is calculated assuming small linear coupling. The method is extended to the problem of the interaction between two multimodal systems.

534.52 **2525**

Spherical Aberration of Solid Acoustic Focusing Lenses—B. D. Tartakovskiĭ. (*Akust. Z.*, vol. 7, no. 3, pp. 349–357; 1961.) The dependence of the longitudinal spherical aberration on the index of refraction and the shape of the lens is investigated, and the optimum relations between them are determined. The role of transverse waves in forming an additional focus is explained.

534.61–14 **2526**

Measurement of the Thermal-Noise Spectrum of Water—D. H. Ezrow. (*J. Acoust. Soc. Am.*, vol. 34, pp. 550–554; May, 1962.) Experimental values in the frequency range 1.0–2.5 Mc determined from measurements of the radiation resistance of thin-disk transducers loaded by water, show good agreement with a theoretical spectrum.

534.614–8 **2527**

Analysis of the Pulse Superposition Method for Measuring Ultrasonic Wave Velocities as a Function of Temperature and Pressure—H. J.

McSkimin and P. Andreatch. (*J. Acoust. Soc. Am.*, vol. 34, pp. 609–615; May, 1962.) A method suitable for measurements on single crystals is described and data on some experiments with quartz are given.

534.62 2528
An Anechoic Chamber for Acoustic Measurements—A. N. Rivin. (*Akust. Z.*, vol. 7, no. 3, pp. 324–336; 1961.) The chamber is lined with fiberglass wedges and has a working volume of 370 m³. The acoustic field 3–4 m from a radiator is uniform to within ±0.5 db over a wide frequency range down to 70 cps.

534.76:621.396.97 2529
Stereophony—(*Onde élect.*, vol. 42, pp. 155–526; March, 1962.) Twelve papers covering various aspects of transmitting, recording and reproducing systems.

534.78:534.41 2530
The Investigation of Speech by means of the Tone-Pitch Recorder—W. Kallenbach. (*Frequenz*, vol. 16, pp. 37–42; February, 1962.) Applications of the recorder developed by Grützmacher and Loettermoser (see *e.g.*, 1226 of 1953) in phonetic and linguistic investigations are discussed.

534.86:534.6 2531
Electroacoustic Measuring Equipment and Techniques—L. L. Beranek. (PROC. IRE, vol. 50, pp. 762–768; May, 1962.) Sound-level measurements, calibration techniques and the determination of response chsracteristics are discussed.

534.874.1 2532
Model Experiments with Acoustic van Atta Reflectors—K. Walther. (*J. Acoust. Soc. Am.*, vol. 34, pp. 665–674; May, 1962.) An array of 36 conical horns on a flat surface 9 in.×9 in. was used in reflection experiments in the frequency range 2.5–9.0 kc/s. Back-scatter diagrams are shown.

534.88 2533
Range Dependence of Acoustic Fluctuations in a Randomly Inhomogeneous Medium—R. G. Stone and D. Mintzer. (*J. Acoust. Soc. Am.*, vol. 34, pp. 647–653; May, 1962.) Acoustic pulses are transmitted through a water tank in which turbulent mixing is produced by heating from below. The "microstructure" of the water temperature is examined with thermistors and a comparison made with the resulting acoustic fluctuations over various path lengths.

534.88 2534
Generalized Form of the Sonar Equations—R. J. Urick. (*J. Acoust. Soc. Am.*, vol. 34, pp. 547–550; May, 1962.) Sonar equations are derived for short transients or pulsed CW waveforms. For long pulses of constant intensity they reduce to the ordinary sonar equations.

534.88:534.417 2535
Effect of Correlated Phase Fluctuation on Array Performance—H. G. Berman and A. Berman. (*J. Acoust. Soc. Am.*, vol. 34, pp. 555–562; May, 1962.) The response of a uniform array of point detectors receiving CW signals is computed under the assumption that there are correlations in the phase fluctuations of the signals. General formulas are developed for various ranges of correlation.

621.395.625.3 2536
Current Problems in Magnetic Recording—M. Camras. (PROC. IRE, vol. 50, pp. 751–761; May, 1962.) Theories explaining the recording and playback processes are discussed.

681.844.08 2537
Some Aspects of Gramophone Pickup Design—R. W. Bayliff. (*Proc. IEE*, pt. B, vol. 109, pp. 233–243; May, 1962.) Analysis of the mechanical problems together with a new pickup design.

ANTENNAS AND TRANSMISSION LINES

621.315.2 2538
The Characteristic Impedance of Cylindrical Conductors in respect of Variously Arranged Plane Screens—W. Buschbeck. (*Telefunken Ztg.*, vol. 34, pp. 69–76; March, 1961.) English summary, p. 81.) The impedance is calculated for thin cylindrical conductors placed a) in a right-angled corner, b) in a tube of square section, c) in a rectangular U-channel, and d) between two parallel planes, all of conducting material. Design curves and formulas are given.

621.315.212 2539
Relation between Reflections in High-Frequency Cables with Static and with Pulse-Modulated Operation—H. E. Martin and D. Dosse. (*Arch. elekt. Übertragung*, vol. 16, pp. 56–66; February, 1962.) Under certain conditions a simple relation exists between the frequency-dependent input reflection coefficient and the pulse-echo voltage due to internal reflections along the cable, as obtained by the two standard methods of cable testing. The derived relation makes it possible to dispense with one of the tests.

621.315.212 2540
Reflection-Coefficient Curves of Compensated Discontinuities on Coaxial Lines and the Determination of the Optimum Dimensions: Part 2—A. Kraus. (*J. Brit. IRE*, vol. 23, pp. 365–371; May, 1962.) The optimum dimensions of supports compensated at their edges are derived from a series of measurements. Part 1: 1487 of 1960.

621.372.823:621.372.833 2541
A Broad-Band Waveguide Junction containing Dielectric—P. J. B. Clarricoats. (*Proc. IEE*, pt. C, vol. 108, pp. 398–404; September, 1961.) A method of obtaining an impedance match at the junction of two circular waveguides of different radius is described. It uses an axial dielectric rod partially filling the guide cross section.

621.372.823:621.372.853.2 2542
Microwave Propagation through Round Waveguide Partially Filled with Ferrite—A. J. Baden-Fuller. (*Proc. IEE*, pt. C, vol. 108, pp. 339–348; September, 1961.) Microwave field components and propagation constants are computed for four cases of circular symmetry. Good agreement is found between calculated and measured Faraday rotation at 9370 Mc.

621.372.824:621.372.831.4:537.226 2543
Action of a Dielectric Window, Metallized on One Face Placed in a Coaxial Waveguide—S. Lefeuvre. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 72–73; January 4, 1961.) A theoretical treatment using Smith diagrams.

621.372.825 2544
Ridged-Waveguide Receiver and Components—W. Shelton. (*Microwave J.*, vol. 5, pp. 101–107; April, 1962.) A summary of design data and performance figures for ridge-waveguide components.

621.372.826 2545
Design of Cylindrical Surface Waveguides with Dielectric and Magnetic Coating—T. Berceli. (*Proc. IEE*, vol. 108, pp. 386–397; September, 1961.) Propagation characteristics for a radially symmetric transverse magnetic wave are given, and used to discuss the design of surface waveguides.

621.372.833 2546
Wide-Band Coupling Systems between a Waveguide and a Transmission Line—B. Rogal and A. L. Cullen. (*Proc. IEE*, pt. C, vol. 108, pp. 433–437; September, 1961.) The system described is particularly suitable for coupling a klystron cavity to a waveguide. Almost constant power transfer is obtained over a bandwidth of about 25 per cent.

621.372.85:538.6 2547
Microwave Hall Effect and the Accompanying Rotation of the Plane of Polarization—H. E. M. Barlow. (*Proc. IEE*, pt. C, vol. 108, pp. 349–353; September, 1961.) The Faraday rotation for TEM waves in a medium, subject to the Hall effect arises from a displacement current, and the behavior at infrared frequencies is briefly discussed.

621.372.852.22 2548
The Part Played by Surface Waves on the Reflection at a Ferrite Boundary—L. Lewin. (*Proc. IEE*, pt. C, vol. 108, pp. 359–361; September, 1961.) Further examination of a problem previously studied (388 of 1960), to resolve an anomalous energy flux situation which arises in a particular case.

621.372.852.32 2549
Some Ferrite Components for Isolation and Controlled Coupling—R. J. Benzie, P. A. N. Kerrigan, and J. A. Penney. (*Proc. IEE*, pt. B, vol. 109, pp. 277–283; May, 1962.) Details are given of three types of current-controlled variable attenuators and switches which are based on microwave power division by the field-controlled microwave permeability of ferrites.

621.372.852.322:621.372.821 2550
Yttrium Garnet Isolator using a Strip Line—Viet Nguyen Tuong. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 686–688; January 30, 1961.) The principle of operation and experimental results are described for isolators using two strip-lines side by side.

621.372.852.5 2551
Mode Conversion in the Excitation of TE_{01} Waves in a TE_{01} Taper-Type Mode Transducer—S. Iiguchi. (*Rev. Elect. Comm. Lab., Japan*, vol. 9, pp. 725–732; Nov./Dec., 1961.) Concise formulas for determining the magnitudes of unused modes are derived from Reiter's telegraphist's equations for a nonuniform waveguide with a straight axis.

621.372.853.1 2552
Backward Waves in Waveguides containing Dielectric—P. J. B. Clarricoats. (*Proc. IEE*, pt. C, vol. 108, pp. 496–501; September, 1961.) Conditions are determined which ensure backward-wave propagation in dielectric-loaded inhomogeneous waveguide structures. An example given is the backward-wave propagation of the hybrid H_{11} mode in a circular waveguide containing an axial dielectric rod.

621.372.853.3:537.56 2553
Propagation in a Waveguide Filled Nonuniformly with Plasma—G. H. Bryant. (*J. Electronics and Control*, vol. 12, pp. 297–305; April, 1962.) By dividing the plasma radially into an inner attenuating region of high average electron density, and an outer region of lower average electron density, an approximate solution is obtained for the E_{01} mode in a plasma-filled cylindrical waveguide.

621.396.67 "313" 2554
The Future of Antennas—M. D. Adcock, R. E. Hiatt, and K. M. Siegel. (PROC. IRE, vol. 50, pp. 712–716; May, 1962.) Likely developments in data processing and electronics

scanning as applied to antenna systems are outlined. The discussion also covers antennas for space communication.

621.396.677:523.164 2555

The New Cambridge Radio Telescope—Ryle. (See 2624.)

621.396.677:523.164 2556

The Italian Cross Radiotelescope: Part 1—Design of the Antenna—Braccesi and Ceccarelli. (See 2623.)

621.396.677:621.396.43:551.507.362.2 2557

Ground Station Aerials for Satellite Communication—J. K. Todd. (*Point to Point Telecommun.*, vol. 6, pp. 30–42; June, 1962.) Mechanical problems of tracking communication satellites with large aerial systems are considered.

621.396.677.001.24:621.396.96 2558

Aerial Design with the Aid of Geometrical Optics—R. Maunz. (*Telefunken Ztg.*, vol. 34, pp. 51–57; March, 1961. English summary, p. 80.) A method of designing radar antennas is described which is based on the dissection of the paraboloidal reflector into elements large with respect to λ; only the vertical radiation pattern is considered.

621.396.677.71:621.372.826 2559

The Launching of Surface Waves by an Endfire Array of Slots—A. L. Cullen and J. A. Staniforth. (*Proc. IEE*, pt. C, vol. 108, pp. 492–495; September, 1961.) An endfire array of slots suitable for launching surface waves on a dielectric-coated metal sheet is described and analyzed. The slots are represented as magnetic dipoles in the plane of the sheet, their axes being at right angles to the line of the array. Experimental results support the theoretical conclusion that high launching efficiency is possible.

621.396.677.85:621.396.965 2560

An Electronically Scanning Radar Antenna—W. H. von Aulock. (*Bell Labs. Rec.*, vol. 40, pp. 118–123; April, 1962.) Problems of design are considered in particular for airborne equipment operating at 9375 Mc.

621.396.679.4:621.396.65 2561

The Effects of Several Reflection Points in Aerial Feeders for F.M. Radio-Link Systems—U. v. Kienlin and A. Kürzl. (*Frequenz*, vol. 16, pp. 49–58; February, 1962.) The frequency characteristics of the reflection coefficient at the input of feeders, and the group delay distortion caused by reflections are determined. The mean noise level of a transmission channel can be estimated from measurements of the maximum input reflection coefficient.

AUTOMATIC COMPUTERS

681.142 2562

The Impact of Hybrid Analogue-Digital Techniques on the Analogue-Computer Art—G. A. Korn. (PROC. IRE, vol. 50, pp. 1077–1086; May, 1962.)

681.142 2563

Quarter-Squares Multiplier—J. C. Cluley. (*Electronic Tech.*, vol. 39, pp. 225–229; June, 1962.) High-speed analog units using square-law diodes are described.

681.142:621.382.23 2564

Some New High-Speed Tunnel-Diode Logic Circuits—M. S. Axelrod, A. S. Farber, and D. E. Rosenheim. (*IBM J. Res. & Dev.*, vol. 6, pp. 158–169; April, 1962.) The circuits are designed for majority-type logic or majority-type logic with inversion and use a multiphase sinusoidal power supply to obtain signal directivity.

CIRCUITS AND CIRCUIT ELEMENTS

621.3:061.3 2565

International Conference on Components and Materials used in Electronic Engineering—(See 2767.)

621.3.004.6:621.39 2566

The Reliability of Components for Telecommunications Equipment—J. Katona. (*Nachrtech.*, vol. 12, pp. 68–72; February, 1962.) Life tests and methods of statistical analysis are outlined.

621.3.049.75 2567

The Micromodule Approach to Microminiaturization—S. F. Danko. (*Electronics Reliability Micromin.*, vol. 1, pp. 65–72; January–March, 1962.) Microminiaturization techniques are classified and the specifications and characteristics of micromodule designs are discussed.

621.3.049.75:621.382 2568

Semiconductor Networks—W. Adcock and J. S. Walker. (*Electronics Reliability Micromin.*, vol. 1, pp. 81–95; January–March, 1962.) Production techniques for integrated networks are described and design data for various circuits are given.

621.319.4.001.4 2569

Short-Period Tests and Long-Term Behaviour of Capacitors—P. Petrick, (*Radio Mentor*, vol. 28, pp. 206–209; March, 1962.) The difficulties of basing predictions of useful life of capacitors on the outcome of short-term tests are discussed. Results of ionization tests on various types of capacitor are analyzed.

621.372.412+534.138-8 2570

Degenerate Oscillations and Doublet Splitting of the Natural Frequencies of Piezoresonators—B. A. Finagin. (*Akust. Z.*, vol. 7, no. 3, pp. 358–365; 1961.) Doublet splitting can occur for certain oscillation modes which are normally degenerate as a result of the removal of degeneracy by a small perturbation.

621.372.413 2571

On Casting Plastic Microwave Cavities—I. M. Firth. (*J. Sci. Instr.*, vol. 39, pp. 131–132; March, 1962.) Epoxy resin microwave cavities are described which after machining, polishing and silver-plating had Q values of about 6000 at 1.2° K.

621.372.413:621.375.9:538.569.4 2572

Design of Multimode Waveguide Cavities for Solid-State Masers—G. J. Troup. (*Proc. IRE (Australia)*, vol. 23, pp. 166–171; March, 1962.) Conditions for the propagation of pure TE and TM modes in anisotropic dielectrics are determined and propagation constants for uniaxial and biaxial crystals are derived. Application of the conditions to maser cavity design is indicated.

621.372.44 2573

The Analysis of Nonlinear Resonant Circuits—C. B. Newport and D. A. Bell. (*Proc. IEE*, pt. C, vol. 108, pp. 374–385; September, 1961.) Two graphical methods are described for calculating the response to a sinusoidal input current of a resonant circuit containing a nonlinear inductor.

621.372.44 2574

Series Resonance with a Nonlinear Capacitor—J. C. Hoffmann. (*C. R. Acad. Sci. (Paris)*, vol. 125, pp. 2904–2905; December 19, 1960.) An improved method of constructing resonance curves taking account of Q-factor variations. See 3519 of 1955.

621.372.512 2575

The Impulse Response of a Number of Identical Circuits in Cascade—K. Tharmalingam. (*Proc. IEE*, pt. C, vol. 108, pp. 335–338; September, 1961.) Approximate analytical expressions are derived, which agree well with exact solutions for a special case. Rough approximations for peak-to-peak ringing are given.

621.372.54 2576

A Method of Calculating the Transfer Functions of Ladder Networks—N. Ream. (*Proc. IEE*, pt. C, vol. 108, pp. 354–358; September, 1961.) Recurrence relations are derived from the Kirchhoff equations and used to obtain transfer functions, working from either the input or output end of the network.

621.372.54 2577

A Note on Optimum Linear Multivariable Filters—R. J. Kavanagh. (*Proc. IEE*, pt. C, vol. 108, pp. 412–417; September, 1961.) A general method is given for factorizing any power matrix to obtain explicit solutions for the optimum filter.

621.372.54:621.372.414 2578

All-Pass Circuit consisting of Transmission-Line Resonators—P. Birgels. (*Arch. elekt. Übertragung*, vol. 16, pp. 149–157; March, 1962.) The use of the bridged-T network as all-pass filter is discussed. For higher frequencies the reactances of the network can be formed by the input impedances of short- or open-circuited transmission-line sections.

621.372.54:621.374 2579

Filters with Favourable Transient Characteristics—J. Jess and W. Schüssler. (*Arch. elekt. Über ragung*, vol. 16, pp. 117–128; March, 1962.) Further design data for pulse-shaping networks are given [see also 1751 of 1961 (Herrmann and Schüssler)].

621.372.6 2580

Sign Matrices and Realizability of Conductance Matrices—G. Biorci. (*Proc. IEE*, pt. C, vol. 108, pp. 296–299; September, 1961.) A procedure for solving some topological network problems is proposed.

621.372.6 2581

Travelling-Wave Analysis of Generalized Networks—J. Zawels. (*Proc. IEE*, pt. C, vol. 108, pp. 300–308; September, 1961.) The behavior of networks is discussed in terms of traveling waves, and generalized wave parameters are derived. A "wave matrix" treatment of multiport networks is given, and typical "wave flow diagrams" are shown.

621.372.6 2582

Topological Synthesis of Nonreciprocal Resistance Networks—R. Onodera. (*Proc. IEE*, pt. C, vol. 108, pp. 325–334; September, 1961.) A method of synthesizing resistance networks for a specified driving-conductance matrix is demonstrated: some examples are given. In general, ideal transformers are required.

621.373:621.391.822 2583

Electrical Generation of Random Processes with Predictable Statistical Properties—J. Swoboda. (*Arch. elekt. Übertragung*, vol. 16, pp. 135–148; March, 1962.) A system is described using a random-noise source for producing chance decisions, with facilities for controlling the probability of events. Applications of stochastic processes thus produced in the testing of communication systems are discussed.

621.373.421.1 2584

Frequency Variations of a Circuit caused by Changing the Loss Angle by means of Diodes—F. Zühlke. (*Nachrtech.*, vol. 12, pp. 78–80; February, 1962.) The tuning of a diode-loaded oscillator circuit by varying the working point of the diode is described.

621.373.431:531.76 2585

Crystal-Locked Blocking Oscillators for a Time Mark Generator—P. P. Petry and C. S. Müller. (*Electronic Eng.*, vol. 34, pp. 395–398; June, 1962.) The circuit comprises six cascaded blocking-oscillator frequency dividers, each with a division ratio of one tenth. The first stage is crystal-controlled. Output rates range from 1 μsec to 1 sec. Long-term stability within 1 part in 10^6 has been observed.

621.374.3:621.382.333 2586

High-Speed Transistorized Pulse Generator—D. Grollet. (*Mullard Tech. Commun.*, vol. 6, pp. 238–245; April, 1962.) The circuit was designed for measurement of semiconductor diode transient characteristics. Pulse amplitude is adjustable up to 50 ma with rise and fall times about 5 nsec.

621.374.32:621.317.373 2587

Frequency Dividing Circuit and Phase Comparator—E. Archbold. (*J. Sci. Instr.*, vol. 39, pp. 107–110; March, 1962.) "A pulse counting circuit is described which delivers every nth pulse to its output where n is any number up to the capacity of the circuit. The phase comparator accepts two pulse trains of the same frequency and measures the phase difference between them to an accuracy of better than four parts in ten thousand."

621.374.32:621.318.134 2588

Flux Counters using Ring Cores of Square-Loop Ferrite—W. Hilberg. (*Nachrtech. Z.*, vol. 15, pp. 88–94; February, 1962.) Several types of counter circuit are described including two suitable for operation within wide temperature limits.

621.374.32:681.142 2589

Nanosecond Coincidence Circuit using Tunnel Diodes—A. Whetstone and S. Kounosu. (*Rev. Sci. Instr.*, vol. 33, pp. 423–428; April, 1962.) A tunnel-diode circuit is described suitable for compounding into complex logic schemes.

621.375.13:517.3 2590

The Standardization of Valve Circuits for Distortionless Electrical Integration: Parts 1 & 2—K. H. Kerber. (*Arch. tech. Messen*, nos. 313 and 314, pp. 37–38 and 67–70; February and March, 1962.) The conditions for error-free integration of continuous signals are established, and the parameters and characteristics of several valve amplifiers with integrating quadripoles in the grid or anode circuits are given.

621.375.4 2591

Mismatch Design of Transistor I.F. Amplifiers—D. M. Duncan. (*Proc. IRE* (*Australia*), vol. 23, pp. 147–158; March, 1962.) A method of designing narrow-band amplifiers is described in which stability is achieved by mismatch of transistor terminations.

621.375.432 2592

Calculation of the Frequency Response of a Nonlinear Transistor Amplifier—I. Gumowski. (*C. R. Acad. Sci.* (*Paris*), vol. 250, pp. 4322–4324; June 27, 1960.) Poincaré's method is used to solve the equation derived in earlier work [1448 of 1961 (Gumowski, *et al.*)].

621.375.9:538.569.4 2593

A Molecular Amplifier for an Operating Temperature of 90°K—H. Reitböck and A. Redhardt. (*Z. Naturforsch.*, vol. 17a, pp. 187–188; February, 1962.) Synthetic ruby is used as active material; the signal frequency is 9730 Mc and the pump frequency 24240 Mc.

621.375.9:621.372.44 2594

Parametric Circuits with Particular Regard to the Characteristics of the Nonlinear Element—K. H. Steiner. (*Arch. elekt. Übertragung*, vol. 16, pp. 67–82; February, 1962.) Detailed discussion of the results given in 3303 of 1961.

621.376.233 2595

Frequency Response of Nanosecond Radio-Frequency Pulse Detectors—B. G. Whitford. (*J. Sci. Instr.*, vol. 39, pp. 303–305; June, 1962.) The pulse distortion caused by inadequate bandwidth in the video output circuit of a crystal detector is measured by using the detector as a mixer of two CW signals and metering the amplitude of the output at the difference frequency. The theory and experimental results for 800-Mc bandwidth are given.

621.376.332:621.385.623.5:621.316.726 2596

Microwave Discriminator for the Frequency Stabilization of a Reflex Klystron—M. J. A. Smith. (*J. Sci. Instr.*, vol. 39, p. 127; March, 1962.) The arrangement and operation of a two-cavity system are described. No microwave bridge is required and no sidebands of the modulation frequency are produced.

GENERAL PHYSICS

535.13 2597

Generalization of the Edge Condition—P. Poincelot. (*C. R. Acad. Sci.* (*Paris*), vol. 250, pp. 4316–4318; June 27, 1960.) The edge condition established in earlier papers (see *e.g.*, 813 of 1961) for an infinitely thin, infinitely conducting plane screen is generalized by considering a perfectly conducting solid with a curved smooth continuous edge. See also C. R. Acad. Sci. (Paris), vol. 251, pp. 2670–2671; December 5, 1960.

537.29+538.69 2598

Decoupling of Bloch Bands in the Presence of Homogeneous Fields—G. H. Wannier and D. R. Fredkin. (*Phys. Rev.*, vol. 125, pp. 1910–1915; March 15, 1962.) Extension of earlier theory [2289 of 1960 (Wannier)].

537.311.3 2599

Electrical Conduction Phenomena in Anisotrophic Media—D. Langbein. (*Z. Phys.*, vol. 167, pp. 83–95 and 96–99; February 28, 1962.)

Part 3—Galvanomagnetic Effects (pp. 83–95).

Part 4—A Further Model for the Electron-Phonon Interaction Matrix Element (pp. 96–99).

Part 2: 2233 of July.

537.311.33 2600

Mobility of Electrons in Intense Electric Fields—A. V. J. Martin and J. Le Mée. (*J. phys. radium*, vol. 23, pp. 7–11; January, 1962.) An introductory report of work on nonconstant-mobility theory, with application to the technetron, and a review of literature dealing with electron mobility at high field strengths. See also 3967 of 1961 (Martin and Le Mée).

537.312.62:538.221 2601

Superconductivity and Ferromagnetism—B. T. Matthias. (*IBM J. Res. & Dev.*, vol. 6, pp. 250–255; April, 1962.) A discussion of the relations between the two phenomena.

537.525.2:538.221 2602

Polarization of Field-Emission Electrons from Magnetically Saturated Iron—H. v. Issendorff and R. Fleischmann. (*Z. Phys.*, vol. 167, pp. 11–19; February 28, 1962.) Field-emitted electrons from longitudinally magnetized Fe were accelerated to 126 kev and a polarization value of 0.0001 ± 0.0019 was measured.

537.533 2603

Electronic Simulations of the Effect of Lens Aberrations on Electron-Beam Cross-Sections—P. W. Hawkes and V. E. Cosslett; J. K. Oxenham. (*Brit. J. Appl. Phys.*, vol. 13, pp. 272–279; June, 1962.) The aberrations produced by focusing or correcting elements without rotational symmetry are simulated up to the fifth order and implicitly to any order. Examples of simulated aberrations are illustrated.

537.533.33 2604

Electron-Optics in Multistage Lens System: Part 1—M. Morikawa. (*Z. angew. Math. Phys.*, vol. 13, pp. 167–175; March 25, 1962. In English.) An extension of the expressions for the cardinal elements of a lens system as derived on the basis of a simplification (to be published in *J. Appl. Phys.*).

537.533.74 2605

The Angle Dependence of the Characteristic Energy Losses on Al, Si, Ag—C. Kunz. (*Z. Phys.*, vol. 167, pp. 53–71; February 28, 1962.) Discrepancies between plasma theory and the results of loss measurements on electrons scattered in passing through metal foils are discussed. Some losses are attributed to the effects of surface excitation. See *e.g.*, 1104 of 1958 (Ferrell and Quinn).

537.56:538.63 2606

Magnetohydrodynamics of a Ternary Plasma and Associated Types of Wave—P. Cavaillès, R. Jancel, and T. Kahan. (*C. R. Acad. Sci.* (*Paris*), vol. 250, pp. 3789–3791; June 8, 1960.

537.56.08 2607

Errors Due to Diffusion in Drift Velocity Measurements—J. J. Lowke. (*Aust. J. Phys.*, vol. 15, pp. 39–58; March, 1962.) Errors in the measurement of electron drift velocity by shutter methods are discussed.

538.11 2608

Magnetic Disorder as a First-Order Phase Transformation—C. P. Bean and D. S. Rodbell. (*Phys. Rev.*, vol. 126, pp. 104–115; April 1, 1962.)

538.561.029.65 2609

Spanning the Microwave Infrared Gap—P. D. Coleman. (PROC. IRE, vol. 50, pp. 1219–1224; May, 1962.) The problems of producing coherent radiation in the 100–1000-μ wavelength range are reviewed and the probable direction of future developments is indicated.

538.566:535.42 2610

Diffraction by a Cylinder with a Variable Surface Impedance—L. B. Felsen and C. J. Marcinkowski. (*Proc. Roy. Soc.* (*London*) *A*, vol. 267, pp. 329–350; May 22, 1962.) The two-dimensional Green's function is evaluated for a cylinder whose surface impedance varies sinusoidally around its periphery.

538.569.4:535.21 2611

Broadening and Shift of Magnetic Resonance Lines caused by Optical Pumping—J. P. Barrat and C. Cohen-Tannoudji. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 255–256; January 9, 1961.) Using equations established in a previous note (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 93–95; January 4, 1961.) optical pumping is considered as a relaxation process in the fundamental state and is shown to cause shifts of magnetic resonance lines.

538.569.4:535.21 2612

Observation of the Shift of a Magnetic Resonance Line caused by Optical Pumping—C. Cohen-Tannoudji. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 394–396; January 16, 1961.) Experimental evidence supporting the theory previously developed (2611 above) is obtained using the isotope ^{199}Hg excited by radiation from a ^{204}Hg source.

538.569.4:538.221 2613

On the Physical Significance of Kittel's Formula in Ferromagnetic Resonance—R. Vautier. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3801–3803; June 8, 1960.) The imaginary demagnetizing factors of Kittel's formula are shown to play the role of supplementary elastic restoring couples which modify the period of natural precessional motion.

538.569.4:538.222 2614

Double Nuclear Resonance and Nuclear Relaxation—E. C. McIrvine, J. Lambe, N. Laurance, and T. Cole. (*Phys. Rev. Lett.*, vol. 8, pp. 318–320; April 15, 1962.) A double-resonance technique is described in which Zeeman transitions of distant nuclei are used to detect hyperfine transitions associated with nuclei at paramagnetic impurities.

538.569.4:621.375.9 2615

Excitation, Relaxation and Continuous Maser Action in the 2.613-Micron Transition of CaF_2:U^{3+}—G. D. Boyd, R. J. Collins, S. P. S. Porto, A. Yariv, and W. A. Hargreaves. (*Phys. Rev. Lett.*, vol. 8, pp. 269–272; April 1, 1962.) Details of experimental arrangements and discussion of results.

538.569.4:621.375.9:535.61–2 2616

Measurements of Optical Maser Oscillation in Ruby—K. Kubota. (*J. Phys. Soc., Japan*, vol. 17, pp. 570–571; March, 1962.) A report on measurements of the maser output for varying input energy.

538.569.4.029.64:537.226 2617

Mechanism of Dielectric Absorption in the Microwave Region—S. Dasgupta and G. Mohan. (*Proc. Roy. Soc. (London) A*, vol. 267, pp. 424–432; May 22, 1962.) A model is proposed to explain the behavior of long-chain aliphatic compounds, which is at variance with the conventional Debye-Fröhlich mechanism.

538.633:621.374/.376 2618

The Corbino Disc and its Self-Magnetic Field—H. H. Johnson and D. Midgley. (*Proc. IEE*, pt. B, vol. 109, pp. 283–286; May, 1962.) Possible applications for rectifiers, amplifiers and trigger circuits are discussed.

539.2:538.1 2619

Magnetic Configurations—F. Bertaut. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 76–78; January 4, 1961.) A general method for analyzing spin configurations theoretically is described. See 131 of 1961.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.12 2620

'Arc Welding' Theory of Cosmic Evolution—C. E. R. Bruce. (*Engineering (London)*, vol. 192, pp. 821–823; December 22, 1961.) Explanation of the formation of the universe on the basis of electric field breakdown leading to aggregation of matter along the discharge channels. See also 2271 of July.

523.164:523.75 2621

Sudden Cosmic Noise Absorption at 29 Mc/s—M. Krishnamurthi, G. S. Sastry, and T. S. Rao. (*Aust. J. Phys.*, vol. 15, pp. 20–24; March, 1962.) Absorption events are closely related to intense solar flares, but are governed by initial conditions in the terrestrial atmosphere and at frequencies above 25 Mc cannot be used for flare prediction.

523.164:621.396.62 2622

The Italian Cross Radiotelescope: Part 2—Preliminary Design of the Receiver—G. Gelato, C. Rosatelli, and G. Sinigaglia. (*Nuovo Cim.*, vol. 23, pp. 254–257; January 1, 1962.) A general description of the receiver to operate in conjunction with the antennas described in Part 1.

523.164:621.396.677 2623

The Italian Cross Radiotelescope: Part 1—Design of the Antenna—A. Braccesi and M. Ceccarelli. (*Nuovo Cim.*, vol. 23, pp. 208–215; January 1, 1962.) The telescope, which will operate at 408 Mc, is of Mills cross design with two antennas each about 1.2-km long and 30-m wide.

523.164:621.396.677 2624

The New Cambridge Radio Telescope—M. Ryle. (*Nature*, vol. 194, pp. 517–518; May 12, 1962.) Three 60-ft-diameter circular antennas, two fixed and one on a railway line running E-W give a comparable performance to that of a conventional paraboloid 1250 ft in diameter. It is planned to use the instrument initially on a frequency of 408 Mc/min. with an expected resolution of 1.5 arc.

523.164.3 2625

Radiation from Jupiter at 4.8 Mc/s—G. R. A. Ellis. (*Nature*, vol. 194, pp. 667–668; May, 1962.) A description of measurements made at Hobart, Tasmania, June–August, 1961.

523.164.32 2626

A Study of the Slowly Varying Component of the Sun at 327 Mc/s—G. Mannino, G. Setti, and S. Delli Santi. (*Nuovo Cim.*, vol. 23, pp. 923–926; March 1, 1962.) Particular attention is given to the relation between the radio flux and sunspot areas.

523.164.32:523.75 2627

Velocities of Shock Fronts in Solar Corona Generating Type II Radio Bursts at Metre Wavelengths—M. Krishnamurthi, G. S. Sastry, and T. S. Rao. (*Aust. J. Phys.*, vol. 15, pp. 120–122; March, 1962.) Intense solar flares produce X rays which cause sudden cosmic noise absorption and shock fronts which excite plasma oscillations in the solar corona and cause enhanced RF emission. Shock-front velocities are calculated from the time difference between these two events.

523.164.32:523.75 2628

The Decimetre-Wavelength Radiation Associated with Type IV Solar Radio Bursts—T. Krishnan and R. F. Mullaly (*Aust. J. Phys.*, vol. 15, pp. 86–95; March, 1962.) This radiation is produced simultaneously with m-λ emission, in widely separated regions of the solar atmosphere. It is not simply an extension of the Type IV continuum.

523.165 2629

Small Cosmic-Ray Increases Measured at Ground Level, September 3, 1960, July 18 and July 20, 1961—B. G. Wilson, D. C. Rose, and M. D. Wilson. (*Canad. J. Phys.*, vol. 40, pp. 540–549; May, 1962.)

523.165 2630

Energy Spectrum of Geomagnetically Trapped Protons—H. H. Heckman and A. H. Armstrong. (*J. Geophys. Res.*, vol. 67, pp. 1255–1262; April, 1962.) The measured energy spectrum of protons in the inner Van Allen radiation belt is discussed for energies >12 mev.

523.165 2631

Geomagnetically Trapped Protons from Cosmic-Ray Albedo Neutrons—A. M. Lenchek and S. F. Singer. (*J. Geophys. Res.*, vol. 67, pp. 1263–1287; April, 1962.) Albedo neutrons from galactic cosmic rays are shown to account for the gross features of the observed geomagnetically trapped protons. This leads to an equilibrium intensity that compares favorably with observations. 65 references.

523.165 2632

Outer Van Allen Belts and Neutral Points on Interface between Solar Wind and Geomagnetic Field—C. C. Chang. (*Nature*, vol. 194, pp. 424–426; May 5, 1962.) A possible trapping mechanism for high-energy solar particles in the formation of the Van Allen belt is suggested. Two neutral points, found by analysis on the interface between solar wind and geomagnetic field allow the plasma to seep into the field without reflection.

523.165:551.507.362.2 2633

Measurements of the Intensity and Spectrum of Electrons at 1000-Kilometre Altitude and High Latitudes—B. J. O'Brien, C. D. Laughlin, J. A. Van Allen, and L. A. Frank. (*J. Geophys. Res.*, vol. 67, pp. 1209 1225; April, 1962.) A preliminary report on measurements made by detectors on the satellite Injun I at magnetic latitudes between about 45° and 80°.

523.165:551.507.362.2 2634

Direct Observations of Dumping of Electrons at 1000-Kilometre Altitude and High Latitudes—B. J. O'Brien (*J. Geophys. Res.*, vol. 67, pp. 1227–1233; April, 1962.) A discussion of the detection by the satellite Injun I of the quiescent dumping of electrons with energy above 40 kev in intensities of the order of 10^1 particles per cm^2 sec. ster.

523.165:551.507.362.2 2635

Time Variations of Intensity in the Earth's Inner Radiation Zone, October 1959 through December 1960—G. Pizzella, C. E. McIlwain, and J. A. Van Allen. (*J. Geophys. Res.*, vol. 67, pp. 1235–1253; April, 1962.) Measurements made by Explorer VII indicate a time variation in the earth's inner radiation belt and suggest the existence of a strong impulsive source of trapped particles throughout the inner zone.

523.5:621.396.96 2636

Radar Observat ons of Meteor Echoes as a Means of Investigation in the Field of Meteor Astronomy and Physics—N. Carrara, P. F. Checcacci, and L. Ronchi. (*Nuovo Cim.*, vol. 24, pp. 145–155; April 1, 1962.) Methods of deriving velocity and radiant distributions of meteors and obtaining data on trail formation in the atmosphere are described.

523.5:621.396.96 2637

Radio-Echo Observations of Meteors in the Antarctic—C. S. Nilsson and A. A. Weiss. (*Aust. J. Phys.*, vol. 15, pp. 1–19; March, 1962.) The observed characteristic are in agreement with a three-source model for the distribution of the radiants.

523.75:550.38 2638

Propagation of Solar Particles and the Interplanetary Magnetic Field—C. S. Warwick. (*J. Geophys. Res.*, vol. 67, pp. 1333–1346; April, 1962.) The presence of an interplanetary magnetic field is inferred from effects on the propagation of energetic solar particles. This field is related to polar-cap absorption and the solar modulation of galactic cosmic rays.

550.38 2639

The Distant Geomagnetic Field: Part 1—Infinitesimal Hydromagnetic Waves—C. P. Sonett, A. R. Sims, and I. J. Abrams. (*J. Geophys. Res.*, vol. 67, pp. 1191–1207; April, 1962.) Comprehensive machine reductions of

Pioneer I magnetometer data are discussed in connection with the presence of infinitesimal hydromagnetic waves at geocentric distances 2.4–4.4×10⁴ km.

550.384.4 2640
Diurnal Variations of the Geomagnetic Field at Tamanrasset from 1948 to 1955 (Horizontal Component)—F. Duclaux. (*Ann. Géophys.*, vol. 18, pp. 110–115; January–March, 1962.

550.385.3 2641
Geomagnetic Bay-Like Disturbances before Geomagnetic Sudden Commencements of Sudden Impulses—T. Ondoh. (*Ann. Géophys.*, vol. 18, pp. 45–61; January–March, 1962. In English.) Solar RF bursts of Type IV and polar-cap blackouts are shown to bear a close relation to the geomagnetic bay-like disturbances.

550.385.37 2642
Geomagnetic Micropulsations—Y. Kato. (*Aust. J. Phys.*, vol. 15, pp. 70–85; March, 1962.) A detailed summary of world-wide observational results. 66 references.

550.385.37 2643
Geomagnetic Micropulsations with Periods 0.3–3 sec. ('Pearls')—J. A. Jacobs and E. J. Jolley. (*Nature*, vol. 194, pp. 641–643; May 19, 1962.)

550.385.4:551.510.53 2644
Relations between Atmospheric Ozone and the Earth's Magnetism—A. Vassy and I. Rasool. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3865–3866; June 8, 1960.) An analysis of two years' data shows a correlation between the occurrence of magnetic storms and increase of atmospheric ozone.

550.386:523.74 2645
On the Semiannual Variation of Geomagnetic Activity and its Relation to the Solar Corpuscular Radiation—W. Priester and D. Cattani. (*J. atmos. Sci.*, vol. 19, pp. 121–126; March, 1962.) Indirect support is given to the axial hypothesis of semiannual variation by a strong amplitude modulation of the variation during the solar cycle which is apparently related to the heliographic latitudes of the sunspot zones. Three models are derived to explain observed geomagnetic variations but show a phase lag of 20–30 days.

551.507.362.1 2646
Investigation of the Ion Composition of the Earth's Atmosphere on Geophysical Rockets 1957–1959.—V. G. Istomin. (*Isk. Sput. Zemli*, no. 7, pp. 64–77. 1961. *Planet-Space Sci.*, vol. 9, pp. 179–193; April, 1962.)

551.507.362.2 2647
The First Anglo-American Satellite—(*Electronic Eng.*, vol. 34, pp. 400–401; June, 1962.) A list of the experiments for which ARIEL was designed.

551.507.362.2 2648
Intermediary Equatorial Orbits of an Artificial Satellite—J. P. Vinti. (*J. Res. NBS*, vol. 66B, pp. 5–13; January-March, 1962.) Restrictions on the orbital inclinations specified in a previous paper (115 of January) are removed and simplifications of the periodic terms for approximately equatorial orbits are given.

551.507.362.2 2649
Luni-solar Perturbations of the Orbit of an Earth Satellite—G. E. Cook. (*Geophys., J. R. astr. Soc.*, vol. 6, pp. 271–291, April, 1962). The effects of the gravitational attraction of the sun and moon on the orbital elements of an earth satellite are investigated using Lagrange's planetary equations. Expressions are obtained for the change in the elements during one revolution of the satellite and for the rates of change of these elements. Corresponding expressions are obtained for the effects of solar radiation pressure, including the effect of the earth's shadow.

551.507.362.2 2650
The Contraction of Satellite Orbits under the Influence of Air Drag: Part 2—With Oblate Atmosphere—G. E. Cook, D. G. King-Hele, and D. M. C. Walker. (*Proc. Roy. Soc. (London) A*, vol. 264, pp. 88–121; October 24, 1961.) An extension of earlier work (Part 1: *Proc. Roy. Soc. (London) A*, vol. 257, pp. 224–249; Sept. 6, 1960) in which an analytical method was developed for determining the effect of air drag on satellite orbits of low eccentricity in a spherically symmetric atmosphere.

551.507.362.2 2651
The Contraction of Satellite Orbits under the Influence of Air Drags: Part 3—High-Eccentricity Orbits ($0.2<c<1$)—D. G. King-Hele. (*Proc. Roy. Soc. (London) A*, vol. 267, pp. 541–557. June 5, 1962.) Part 2: 2650 above.

551.507.362.2 2652
Atmospheric Density Measurements with a Satellite-Borne Microphone Gauge—G. W. Sharp, W. B. Hanson and D. D. McKibbin. (*J. Geophys. Res.*, vol. 67, pp. 1375–1382; April, 1962.) Ram pressure measurements of atmospheric densities agree with satellite orbital data.

551.507.362.2:621.391.812.3 2653
The Scintillation of the Radio Transmissions from Explorer VII: Parts 1 & 2—D. G. Singleton and G. J. E. Lynch. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 353–374; May, 1962.) The scintillations at 20 Mc are associated with both the frequency- and range-spread types of spread F. They can be interpreted in terms of irregularities at heights of 200–600 km, which are field aligned and have dimensions of the order of 1 km. A distinction can be made between the scintillations associated with the two types of spread F.

551.507.362.2:621.396.43 2654
The Orbits of Needle Satellites—G. E. Cook and J. M. Hughes. (*Planet-Space Sci.*, vol. 9, pp. 153–166; April, 1962.) The formation of orbital belts of needles and perturbations due to solar radiation pressure and the earth's gravitational potential are investigated.

551.507.362.2:396.9 2655
Navigational Satellites—(See 2684).

551.507.362.2:621.396.9 2656
Navigation Satellites with particular reference to Radio Observations—Johnson. (See 2685).

551.507.362.2:+629.19]:621.398 2657
Instrumentation for Space Research.—D. Brini, U. Ciriegi, A. Gandolfi and G. L. Tabellini. (*Nuovo Cim.*, vol. 22. suppl. no. 2, pp. 545–589; 1961. In English.) Detailed description of telemetry and tele-control systems and circuits.

551.510.535 2658
Atmospheric Blow-up at the Auroral Zone—K. D. Cole. (*Nature*, vol. 194, p. 761; May, 26, 1962.) As a result of Joule heating of the upper atmosphere by the eectric currents which cause geomagnetic variations (1942 of June) regions of the auroral ionosphere above 150 km may show scale height increases of the order of ten times.

551.510.535 2659
Atmospheric Composition and the F Layer of the Ionosphere—H. Rishbeth. (*Planet-Space Sci.*, vol. 9, pp. 149–152; April, 1962.) The photochemical processes which are believed to control the production and loss of F-region ionization are summarized. Seasonal anomalies and storm effects in the F_2 layer are discussed.

551.510.535 2660
The Geographical Distribution of Ionization and the Equator of the F2 Layer.—R. Eyfrig. (*J. Geophys. Res.*, vol. 67, pp. 1678–1682; April, 1962.) The existence of the equatorial trough in the western zone during the northern solstice month of 1953 is confirmed [see 3849 of 1960 (Minnis and Bazzard)]. An examination of data for July, 1953–1959 shows that in the western zone the minimum of the trough does not coincide with the magnetic-dip-equator.

551.510.535:523.164.4 2661
Ionospheric Focusing of Cosmic Radio Sources—M. D. Grossi, K. M. Strom and S. E. Strom. (*J. Geophys., Res.*. vol. 67, pp. 1672–1674; April, 1962.) The earth's ionosphere, acting as a spherical lens, can focus HF cosmic radio waves.

551.510.535:523.745 2662
Solar Activity and the Occurrence of Es—M. K. Das Gupta and R. K. Mitsa. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 408–411; May, 1962.) Statistical analysis suggests that there is a positive correlation between the occurrence of E_s with $fc>5$ Mc at Inverness, Scotland, and the Zurich sunspot number.

551.510.535:523.746.5 2663
Dependence of the Ionospheric F Region on the Solar Cycle—J. W. Wright. (*Nature*, vol. 194, pp. 461–462; May 5, 1962). Ionospheric electron density distributions are examined for typical quiet days in high, medium and low altitudes. Summer and winter distributions are illustrated for sunspot numbers 0, 100 and 200.

551.510.535:550.386 2664
The Effect of Geomagnetic Acitivity on the F2 Region over Central Africa—R. G. Rastogi. (*J. Geophys. Res.*, vol. 67, pp. 1367–1374; April, 1962.) The variations of the F2 layer with time of day, season, and magnetic activity are examined for stations in Central Africa where the geomagnetic and the magnetic equators are separated from each other. It is concluded that the variations at low and medium latitudes are determined by the magnetic and not by the geomagnetic latitude of the station.

551.510.535:551.507.362.1 2665
Rocket Measurement of the Electron-Density Distribution in the Topside Ionosphere—S. J. Bauer and J. E. Jackson. (*J. Geophys., Res.*, vol. 67, pp. 1675–1677; April, 1962.) Electron densities are measured by observations of the Faraday rotation at 73.6 Mc, and are compared with a theoretical model.

551.510.535:551.507.362.1 2666
Instrumentation for Rocket Measurements of the Free-Electron Density in the Ionosphere—K. I. Gringauz, V. A. Rudakov and A. V. Kaproskiĭ. (*Isk. Sput. Zemli*, no. 6, pp. 33–47; 1961, *Planet-Space Sci.*, vol. 9, pp. 247–257; May, 1962.) A description of the system of radio instrumentation used to investigate the electron density distribution in the ionosphere by geophysical rockets.

551.510.535:551.507.362.1 2667
Mass Spectrometer Investigations of the Structural Parameters of the Earth's Atmosphere at Altitudes from 100 to 210 km—A. A. Pokhunkov. (*Isk. Sput. Zemli*, no. 7, pp. 89–

100; 1961, *Planet-Space Sci.*, vol. 9, pp. 269–279; May, 1962.

551.510.535:621.391.812.63 2668
Polarization Characteristics of E_s Echoes at Waltair—C. A. Reddy and B. R. Rao. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 401–405; May, 1962.) At frequencies in the range 2–8 Mc, polarization measurements show that dense E_s reflects both magneto-ionic components while the semitransparent type reflects mainly the X component.

551.510.535:621.391.812.63 2669
Some Aspects of Gyro Splitting in the Ionosphere—G. G. Bowman. (*Aust. J. Phys.*, vol. 15, pp. 25–38; March, 1962, plates.) X-mode propagation below the electron gyro-frequency was observed in some ionograms from Hobart (geomagnetic latitude 52°S). A detailed study of a few examples is made on the basis of Ellis's theory (2888 of 1960).

551.510.535(98) 2670
The Ionospheric Triple Splitting in High Geomagnetic Latitudes—W. Becker and P. Schanz. (*Arch. elekt. Übertragung*, vol 16, pp. 100–104; February, 1962.) Quantiatative confirmation of the qualitative results given by Wright (4248 of 1960) is derived. Triple splitting is possible only in rhe E layer even at Thule. An interpretation is given of the third F-layer component occasionally observed and a closer analysis of ionograms is proposed.

551.510.535(98):621.391.812.631 2671
A Study of Solar Activity associated with Polar-Cap Absorption—C. S. Warwick and M. W. Haurwitz. (*J. Geophys., Res.*, vol. 67, pp. 1317–1332; April, 1962). Solar data are used to identify the solar event responsible for each of 43 polar-cap absorption events. The results show little evidence for the association of p.c.a. with places on the far side of the sun and there is no significant dependence of occurrence frequency or delay time on position of the flare. A relation is found between delay time and phase of the solar cycle, with long delay times occurring near the maximum of solar activity.

551.510.535(98):621.391.812.631 2672
Polar-Cap Absorption and the Magnetic Storm of February 11, 1958—W. L. Axford and G. C. Reid. (*J. Geophys. Res.*, vol. 67, pp. 1692–1696; April, 1962.) Interplanetary magnetic field changes, due to the solar corpuscular emissions associated with geomagnetic disturbance, are discussed in relation to polar-cap absorption.

551.510.62:551.508.8 2673
Radiosonde for Atmospheric Refractive-Index Measurements—A. P. Deam. (*Rev. Sci. Instr.*, vol. 33, pp. 438–441; April, 1962.) A description of a light-weight refractometer, operating at 403 Mc, with a ceramic coaxial cavity as the sensing device.

551.594.5 2674
The Occurrence of Aurora in Geomagnetically Conjugates Areas.—R. N. DeWitt. (*J. Geophys. Res.*, vol. 67, pp. 1347–1352; April, 1962.)

551.594.5:550.385.36 2675
An Auroral-Zone Electron Precipitation Event and its Relationship to a Magnetic Bay—R. R. Brown and W. H. Campbell. (*J. Geophys. Res.*, vol. 67, pp. 1357–1366; April, 1962.) Simultaneous observations of X rays at balloon heights, cosmic-noise absorption, geomagnetic micropulsations, and the geomagnetic elements H, D and Z have made it possible to follow the details of the event. It is suggested that the current system of the bay was initiated by a sudden increase in the flux of precipitated electrons.

551.594.5:550.386 2676
Magnetic Activity during Periods of Auroras at Geomagnetically Conjugate Points—E. Wescott. (*J. Geophys. Res.*, vol. 67, pp. 1353–1355; April, 1962.) The magnetic variations at Kotzebue, Alaska, and Macquarie Island, in the sub-Antarctic, show a high degree of correlation on two days when aurorae were present at the two stations.

551.594.5:621.396.96 2677
Radio-Echo Observations of the Aurora in Terre Adélie: Part 2—The Fading Characteristics of the Radio Echoes and their Frequency of Occurrence as a Function of Range, Amplitude, Height and the Aspects Effect—K. Bullough. (*Ann. Géophys.*, vol. 18, pp. 1–17; January–March, 1962. In English.) A more detailed study of several aspects of the radio-echo phenomena reported in Part 1 (see 4147 of 1961.)

551.594.5:621.396.96 2678
A Study of the Statistics of VHF Oblique Auroral Reflections—A. Egeland, J. Ortner, and B. Hultqvist. (*Ann. Geophys.*, vol. 18, pp. 23–44; January-March, 1962. In English) Measurements from October, 1957, to May, 1960 at 92.8 Mc at Kiruna, Sweden, have been correlated and data are given for diurnal, day-to-day and seasonal variations as well as duration and signal strength of echoes.

551.594.5:621.396.96 2679
Evidence of Low-Frequency Amplitude Fluctuations in Radar Auroral Echoes—R. S. Leonard. (*J. Geophys. Res.*, vol. 67, pp. 1683–1684; April, 1962.) The amplitude fluctuations of an auroral radar echo are compared with magnetic field micropulsations.

551.594.6 2680
Direct Observation of Correlation between Aurorae and Hiss in Greenland—T. S. Jørgensen and E. Engstrup. (*Nature*, vol. 194, pp. 462–463; May 5, 1962.) A close correlation is observed between 8-kc hiss and the occurrence of aurorae. See also 3869 of 1960 (Martin, *et al.*).

551.594.6 2681
Very-Low-Frequency Noise at Brisbane—H. E. Brown & G. G. Cairns. (*Nature*, vol. 194, p. 962; June 9, 1962.) Results obtained at Brisbane are illustrated. This is the lowest latitude at which noise in the 5-kc region has been recorded.

551.594.6 2682
The Ray Paths of Whistling Atmospherics: Differential Geometry—R. F. Mullaly. (*Aust. J. Phys.*, vol. 15, pp. 106–113; March, 1962.) Expressions are derived for the ray curvature in terms of electron density, magnetic field strength, and field direction. The rays do not generally follow the lines of force closely.

551.594.6 2683
Wide-Band Bursts of V.L.F. Radio Noise (Hiss) at Hobart—R. L. Dowden. (*Aust. J. Phys.*, vol. 15, pp. 114–119; March, 1962.) Because of attenuation, the ground intensities observed at several frequencies drop markedly with increase in frequency, while the deduced intensities above the ionosphere show a relatively flat spectrum.

LOCATION AND AIDS TO NAVIGATION

621.396.9:551.507.362.2 2684
Navigational Satellites—(*J. Inst. Nav.*, vol. 15, pp. 129–157; April, 1962.) The text of four papers presented at a meeting of the Institute of Navigation, September 26, 1961.
The Transit System—R. B. Kershner and R. R. Newton. (pp. 129–144).
A First Attempt to Obtain a Fix from Transit—W. Nicholson. (pp. 144–149).
Navigational Aids from Other Satellites—H. C. Freiesleben. (pp. 149–154).
Ionospheric Research through Transit—E. Golton. (pp. 154–157).

621.396.9:551.507.362.2 2685
Navigation Satellites with particular reference to Radio Observations—W. A. Johnson. (*J. Brit. IRE*, vol. 23, pp. 383–397; May, 1962. Description of work at the Royal Aircraft Establishment.

621.396.933 2686
The Air Traffic Control Equipment Subsystem—Present and Future—P. C. Sandretto. (PROC. IRE, vol. 50, pp. 663–672; May, 1962.) The technical aspects of the control problem are considered and forecasts are made regarding the equipment required for future systems. 79 references.

621.396.96.089.6:523.164.32 2687
Passive Radar Measurements at C-Band Using the Sun as a Noise Source—W. O. Mehuron. (*Microwave J.*, vol. 5, pp. 87–94; April, 1962.) An investigation of atmospheric refraction and attenuation, radiation characteristics of the sun, solar scintil ation, and the usefulness of the sun as a navigation reference was made as part of a development program for precision radar and other tracking systems.

621.396.963 2688
Limits of the Detectability of Targets in M.T.I. Rardar Systems—G. Kuhrdt. (*Telefunken Ztg.*, vol. 34, pp. 42–50; March, 1961. English summary, pp. 79–80.) A graphical method is given for determining the practical limits of a system allowing for the presence of fixed-target signals and the velocity of the moving target.

621.396.969.34 2689
Medium-Range Radar for Air Traffic Control—J. Mücke and K. Röhrich. (*Telefunken Ztg.*, vol. 34, pp. 5–12; March, 1961. English summary, pp. 77.) Operational requirements and design specifications of new German 23-cm radar equipment are outlined. For details of the system see 2690 below.

621.396.969.34 2690
Medium-Range Radar Equipment—(*Telefunken Ztg.*, vol. 34, pp. 13–41; March, 1961.) A group of papers describing various features of the new German radar system outlined in 2689 above.
a) The Transmitter of the Medium-Range Radar Installation—H. O. Voigt. (pp. 13–21, English summary, pp. 77–78.
b) The Display Equipment of the Medium-Range Radar Installation—K. Schluckebier. (pp. 22–27, English summary, p. 78).
c) Radar-Display Transmission for the Medium-Range Radar Installation—H. Hart. (pp. 27–32, English summary, p. 78).
d) The Aerial of the Medium-Range Radar Installation—J. Bartholomä. (pp. 33–41, English summary, p. 79).

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215:546.48'221 2691
On the Varied Shape of the Spectral Sensitivity Curves of the Photoconductivity of Pure Cadmium Sulphide Crystals. Influence of Crystal Thickness—E. Grillot, E. F. Gross, M. Bancie-Grillot, and B. Novikov. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 864–866; February 6, 1961.) The shape of the curve of the photo-effect at 77° K was found to vary with the thickness of the specimen. Crystals were prepared by sublimation, with a strong temperature gradient.

535.37 2692
Experimental Verification of Field Ionization of Traps in Luminescent Materials—D. E. Brodie, H. E. Petch and R. R. Haering. (*Canad. J. Phys.*, vol. 40, pp. 665–669; May, 1962.) Experimental investigation of the applicability of the field-ionization method for the measurement of trap energies.

535.37 2693
On the Luminescence of TlCl+KCl Phosphor—S. C. Sen. and H. N. Bose. (*Z. Phys.*, vol. 167, pp. 20–25; February 28, 1962. In English.) An analysis of X-ray diffraction spectra.

535.37:546.47'221 2694
Paramagnetic Resonance of Fe^{3+} Ions in Synthetic Cubic Crystals of ZnS—A. Räuber and J. Schneider. (*Z. Naturforsch.*, vol. 17a, pp. 266–270; March, 1962.)

535.37:546.47'221 2695
Piezoluminescence in Zinc Sulphide Phosphors—G. Alzetta, N. Minnaja and S. Santucci. (*Nuovo Cim.*, vol. 23, pp. 910–913; March 1, 1962.) The crystals were mechanically excited by hydrostatic compression. Both pressure steps and pressure pulses were used.

535.376:546.48'221 2696
Displacement of the Cadmium Atom in Single-Crystal CdS by Electron Bombardment—B. A. Kulp. (*Phys. Rev.*, vol. 125, pp. 1865–1869; March 15, 1962.) A threshold for the production of two fluorescence bands in CdS under electron bombardment at −196°C has been observed at 290 kev.

537.226 2697
Dielectric Properties of $Pb(FeTa)_{0.5}O_3$—$PbTiO_3$ and $Pb(FeTa)_{0.5}O_3$—$PbZrO_3$ Systems—S. Nomura and T. Kawakubo. (*J. Phys. Soc., Japan*, vol. 17, pp. 573–574; March, 1962.) The positions of the Fe and Ta ions in the $Pb(FeTa)_{0.5}O_3$ crystal are studied in X ray and dielectric measurements.

537.227 2698
Material Constants of Ferroelectric Ceramics at High Pressure—R. F. Brown and G. W. McMahon. (*Canad. J. Phys.*, vol. 40, pp. 672–674; May, 1962.) A continuation of previous work [3428 of 1961 (Brown)]. The effects of planar stress on the effective piezoelectric constants, elastic modulus, electromechanical coupling factor and high-field dielectric loss have been determined for (Ba, Ca, Pb) TiO_3, (Ba, Ca, Co)TiO_3, and the (Pb, Zr)TiO_3 compounds PZT4 and PZT5.

537.227 2699
Ferroelectric Properties of Mixed Titanates of the System $Bi_4Ti_3O_{12}$, n-$BaTiO_3$—P. H. Fang, C. Robbins, and F. Forrat. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 683–685; January 30, 1961.) An experimental investigation covering the temperature range −194° to 800°C for $n=1$ and 2.

537.227:538.569.4.029.6 2700
Microwave Losses in Strontium Titanate above the Phase Transition—G. Rupprecht and R. O. Bell. (*Phys. Rev.*, vol. 125, pp. 1915–1920; March 15, 1962.) Losses have been investigated as a function of frequency and temperature.

537.277:538.569.4.029.6 2701
Microwave Absorption in Cubic Strontium Titanate—B. D. Silverman. (*Phys. Rev.*, vol. 125, pp. 1921–1930; March, 15, 1962.) An analysis of the temperature dependence of the microwave loss in $SrTiO_3$ is presented using the linear chain model of a ferroelectric as a basis for discussion.

537.227:546.431'824–31:539.23 2702
On the Preparation of Thin Single-Crystal Films of $BaTiO_3$—R. C. DeVries. (*J. Amer. Ceram. Soc.*, vol. 45, pp. 225–228; May, 1962.) Crystals 300 A thick were evaporated on to Pt substrates. Uniform thickness was maintained over crystal areas 0.2 mm².

537.228.1:548.0 2703
Lattice Theory of the Piezoelectric and Elastic Properties of Crystals with Zinc Blende Structure allowing for Electron Polarization—L. Merten. (*Z. Naturforsch.*, vol. 17a, pp. 174–180 and 216–227; Febrlary and March, 1962.)
Part 1—General Relations based on Crystal Symmetry (pp. 174–180).
Part 2—Representation of the Piezo-electric Constant e_{14} by Microscope Parameters for the Model of Pure Field Polarization and for the Shell Model (pp. 216–227).

537.228.1:549.514.51 2704
The Growth Rate Dependence of the Internal Friction of Synthetic Quartz—C. K. Jones and C. S. Brown. (*Proc. Phys. Soc.* (*London*) *A*, vol. 79, pp. 930–937; May 1, 1962.)

537.311.32:546.86:538.63 2705
A New Effect in the Magnetoresistance of Antimony at 4.2°K—W. R. Datars and P. C. Eastman. (*Canda. J. Phys.*, vol. 40, pp. 670–672; May, 1962.) The effect has been observed in high-purity single crystals. The V/I characteristic shows a step decrease in voltage across the specimen followed by a nonlinear negative-resistance region.

537.311.32:546.87: 538.63 2706
Classical Explanation of the Anomalous Magnetoresistance of Bismuth—J. J. Hopfield. (*Phys. Rev. Lett.*, vol. 8, pp. 311–312; April 15, 1962.) Recent results of Esaki (1596 of May) showing anomalous breaks in the I/V relations for Bi, are explained by the existence of an extra force due to the onset of travelling-wave amplification, in the equation of motion of the carriers.

537.311.32:546.87:538.632 2707
Hall Coefficient for Thin Films of Bismuth—J. Salardenne. (*J. phys. radium*, vol. 23, pp. 133–134; February, 1962.) Possible causes of the varied values obtained for the Hall coefficient are examined.

537.311.33 2708
Surface Properties of Semiconductors—A. Frova and A. Stella. (*Nuovo Cim.*, vol. 22, suppl. no. 2, pp. 517–544; 1961.) Experimental methods for the determination of surface conductivity and mobility states are reviewed.

537.311.33 2709
Interband Electron-Electron Scattering and Transport Phenomena in Semiconductors—J. Appel. (*Phys. Rev.*, vol. 125, pp. 1815–1823; March 15, 1962.) The effect of interband scattering is investigated using a variational method based on a generalization of Kohler's variation principle to a multiband conductor. The combined influence of interband and intraband scattering on the transport coefficients of Ge is discussed.

537.311.33 2710
One-Phonon Transition Rate in Impurity Conduction—J. Mycielski. (*Phys. Rev.*, vol. 125, pp. 1975–1977; March 15, 1962.) The one-phonon transition rate for carrier transfer is calculated taking account of the dependence of the equilibrium position of the lattice atoms on the state of the carrier. See also 1978 of June.

537.311.33 2711
Warm-Electron Effect in Polar Semiconductor—R. Stratton. (*J. Phys. Soc. Japan*, vol. 17, pp. 590–591; March, 1962.) Correction and extension of earlier work (*Proc. Roy. Soc.* (*London*) *A*, vol. 246, pp. 406–422; August 19, 1958.) Relating to the field dependence of electron mobility in polar semiconductors.

537.311.33:534.283:538.6 2712
Magnetoacoustic Effects in Non-degenerate Semiconductors—H. N. Spector. (*Phys. Rev.*, vol. 125, pp. 1880–1892; March 15, 1962.) A self-consistent semiclassical treatment is given for the attenuation of a sound wave in a non-degenerate impurity conductor with spherical energy bands, using the formalism developed by Cohen, *et al.* (2585 of 1960).

537.311.33:537.323 2713
The Influence of Halogen Doping on the Thermoelectric Properties of the System $Bi_2Te_{3-x}Se_x$—U. Birkholz and G. Haacke. (*Z. Naturforsch.*, vol. 17a, pp. 161–165; February, 1962.)

537.311.33:538.614 2714
Note on the Faraday Effect in Anisotropic Semiconductors—B. Donovan and J. Webster. (*Proc. Phys. Soc.* (*London*) *A*, vol. 79, pp. 1081–1082; May 1, 1962.) Algebraic errors in a recent paper (1318) of April) are corrected and revised expressions are given.

537.311.33:538.632 2715
Measurement of the Hall Effect in Metal-Free Phthalocyanine Crystals—G. H. Heilmeier, G. Warfield, and S. E. Harrison. (*Phys. Rev. Lett.*, vol. 8, pp. 309–311; April 15, 1962.) A report of Hall-effect measurements on an organic semiconductor.

537.311.33:539.23 2716
Electron Tunnelling in Solids—L. W. Davies. (*Proc. IRE* (*Australia*), vol. 23, pp. 127–132; March, 1962.) A review of recent work on tunneling of electrons through insulating films between two metals in either normal or super-conducting states. 15 references.

537.311.33[546.28+546.289 2717
Band Structure of Silicon, Germanium and Related Semiconductors—J. C. Phillips. (*Phys. Rev.*, vol. 125, pp. 1931–1936; March 15, 1962.) A classification of energy levels based on a survey of recent optical and cyclotron-resonance experiments.

537.311.33:[546.28+546.289 2718
Valence Bands of Germanium and Silicon in an External Magnetic Field—V. Evtuhov. (*Phys. Rev.*, vol. 125, pp. 1869–1879; March 15, 1962.)

537.311.33:546.28 2719
Loop-Shaded Images Observed in X-Ray Diffraction Micrographs of Silicon Single Crystals—M. Yoshimatsu, T. Suzuki, T. Kobayashi, and K. Kohra. (*J. Phys. Soc. Japan*, vol. 17, pp. 583–584; March, 1962.) A new kind of lattice defect observed in a Si crystal is reported.

537.311.33:546.28 2720
Impurity-Induced Pipes through Diffused Layers in Silicon—A. Goetzberger. (*Solid-State Electronics*, vol. 5, pp. 61–70; March-April, 1962.) Most "pipes" appear to be caused by diffusion from localized sources of impurities on the surface. A theory is developed and compared with the results of observations.

537.311.33:564.28 2721
A Pulse Staining Method for Delineating n-n^+ and p-p^+ Junctions in Silicon—B. A. Joyce. (*Solid-State Electronics*, vol. 5, pp. 102–104; March/April, 1962.

537.311.33:546.28:539.23 2722
Epitaxial Growth of Silicon by Vacuum Evaporation—B. A. Unvala. (*Nature*, vol. 194, pp. 966–967; June 9, 1962.) Advantages of the method, which produces films comparable to those formed by chemical means, are cleanliness, the use of movable masks, the making of ohmic contacts, and the possibility of fabricating complete devices in a vacuum

537.311.33:546.289 2723
Change of the Debye Temperature of Germanium with Donor Concentration—J. Peretti. (*Nuovo Cim.*, vol. 23, pp. 1144–1146; March 16, 1962. In English.) An explanation is given for the variation in Debye temperature, and calculated values of temperature are compared with the results of measurements.

537.311.33:546.289 2724
The Determination of the Diffusion Coefficient of Arsenic at *p-n* Junctions in Germanium—R. Wölfle and H. Dorendorf. (*Solid-State Electronics*, vol. 5, pp. 98–102; March /April, 1962. In German.) Measurements of the diffusion coefficient at 580° and 650°C are reported and results are compared with those of other authors.

537.311.33:546.289 2725
Electron-Hole Scattering at High Injection Levels in Germanium—L. W. Davies. (*Nature*, vol. 194, pp. 762–763; May 26, 1962.) An expression is given for the mobility due to electron-hole scattering which gives results in agreement with experiment. Results are relevant to semiconductor rectifiers.

537.311.33:546.289 2726
Noise Temperature of Hot Electrons in Germanium—E. Erlbach and J. B. Gunn (*Phys. Rev. Lett.*, vol. 8, pp. 280–282; April 1, 1962.) The distribution function in energy of hot electrons is determined from results of measurements.

537.311.33:546.289 2727
Inelastic Scattering of E ectrons in Germanium—J. Callaway and F. W. Cummings. (*Phys. Rev.*, vol. 126, pp. 5–9; April 1, 1962.) An approximate calculation of the average rate of energy loss of electrons by a molecular excitation process in doped compensated Ge at low temperatures is given.

537.311.33:546.289 2728
Infrared Absorption in Heavily Doped *n*-Type Germanium—C. Haas. (*Phys. Rev.*, vol. 125, pp. 1965–1971, March 15, 1962.)

537.311.33:546.289 2729
The Dependence of Photovoltage on the Wavelength of the Photons Absorbed in Germanium—A. Surduts. (*C. R. Acad. Sci. (Paris)*, vol. 251, pp. 2665–2666; December 5, 1960.) For a given wavelength , which depends on the semiconductor and adsorbed molecules, in addition to the normal action of photons, variations of photopotential are observed, due to fast electrons which cross the surface potential barrier.

537.311.33:546.289:537.312.9 2730
Piezoresistance in *n*-Type Germanium at Low Temperatures—K. Sugiyama and A. Kobayashi. (*J. Phys. Soc. Japan*, vol. 17, p. 574; March, 1962.)

537.311.33:546.289:539.23 2731
Interaction of Gases with Evaporated Germanium Films between 78°K and 373°K—M. J. Bennett and F. C. Tompkins. (*Trans. Faraday Soc.*, vol. 58, pp. 816–828; April, 1962.) The results of experiments using O_2, H_2, N_2, CO and CO_2 are reported.

537.311.33:546.289:539.23 2732
Defects in Epitaxially Vapour-Growth Germanium Deposits—J. R. Dale. (*Solid-State Electronics*, vol. 5, pp. 108–109; March/April, 1962.) A new etch treatment for the study of epitaxial layers is described. Structural features of the deposits not usually seen are made clearly visible.

537.311.33:546.48'161 2733
Free-Charge-Carrier Effects in Cadmium Fluoride—J. D. Kingsley and J. S. Prener. (*Phys. Rev. Lett.*, vol. 8, pp. 315–316; April 15, 1962.) Optical absorption and conductivity due to free electrons have been observed in CdF_2 crystals doped with trivalent rare-earth ions.

537.311.33:546.48'221 2734
Hole Drift Mobility and Lifetime in CdS Crystals—J. Mort and W. E. Spear. (*Phys. Rev. Lett.*, vol. 8, pp. 314–315; April 15, 1962.) Measurements of mobility and lifetime made by fast-pulse techniques are described and briefly discussed.

537.311.33:546.48'231 2735
Exciton Structure and Zeeman Effects in Cadmium Selenide—R. G. Wheeler and J. O. Dimmock. (*Phys. Rev.*, vol. 125, pp. 1805–1815; March 15, 1962.)

537.311.33:546.681'19:538.569.4 2736
Spin Transitions Induced by External R.F. Electric Field in GaAs—E. Brun, R. Hann, W. Pierce, and W. H. Tanttila. (*Phys. Rev. Lett.*, vol. 8, pp. 365–366; May 1, 1962.

537.311.33:546.682'86 2737
Electrical Properties of InSb (*p-n*) Junctions—Y. Marfaing. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3608–3610; May 30, 1962.) An analysis of the I/V characteristics recorded at 80° and 100°K using 25-cps 200 µsec current pulses.

537.311.33:546.682'86 2738
The Galvanomagnetic Properties of InSb with High Magnetic Fields—H. Hieronymus and H. Weiss. (*Solid-State Electronics*, vol. 5, pp. 71–84; March/April, 1962. In German.) The Hall coefficient of intrinsic and heavily *n*-doped InSb at room temperature does not depend upon the magnetic induction up to 150,000 G. It grows for small *n*-doping, and becomes smaller for low *p* doping with increasing magnetic field. The hole mobility decreases from about 700 cm²/V sec with no field to 610 cm²/V sec at 150,000 G. The intrinsic electron concentration in InSb at room temperature is independent of the magnetic induction. From 10 G to 150,000 G, for the conduction band the Hall coefficient $R_n = -1/en$.

537.311.33:546.682'86:535.215 2739
Study of Photovoltaic Effects (Lateral and Transverse) in Indium Antimonide *p-n* Junctions—G. Courrier and Y. Marfaing. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3798–3800; June 8, 1960.) Measurements of infrared sensitivity and its variation with temperature are described. Results are given in terms of equivalent noise power.

537.311.33:621.382 2740
Contributions of Materials Technology to Semiconductor Devices—R. L. Petritz. (Proc. IRE, vol. 50, pp. 1025–1038; May 1962.) Purification techniques, crystal growth and diffusion and alloying processes are discussed and future developments outlined. 76references.

537.311.33:621.391.822 2741
Recombination-Generation Noise Theory for Semiconductors in Equilibrium—D. A. Evans and P. T. Landsberg, (*Proc. Roy. Soc. (London) A*, vol. 267, pp. 464–477; June 5, 1962.) A unified discussion is given of equilibrium recombination-generation noise in semiconductors, and its relation to the steady-state and transient lifetimes. It extends existing results and covers situations where certain types of Auger processes, as well as one-electron transitions, contribute to the noise.

537.312.62 2742
Superconductivity in Molybdenum—T. H. Geballe, B. T. Matthias, E. Corenzwit, and G. W. Hull, Jr. (*Phys. Rev. Lett.*, vol. 8, p. 313; April 15, 1962.)

537.312.62 2743
Superconductive Tunnelling—M. H. Cohen, L. M. Falicoy, and J. C. Phillips. (*Phys. Rev. Lett.*, vol. 8, pp. 316–318; April 15, 1962.) A Hamiltonian treatment of the process of tunnelling from a normal metal to a superconductor.

537.312.62:538.6 2744
Heat Capacity Evidence for a High Degree of Superconductivity in V_3Ga in High Magnetic Fields—F. J. Morin, J. P. Maita, H. J. Williams, R. C. Sherwood, J. H. Wernick, and J. E. Kunzler. (*Phys. Rev. Lett.*, vol. 8, pp. 275–277; April 1, 1962.) Results indicate that the sample contains a large number of filaments making most of it appear superconducting.

537.583 2745
Determination of Mechanical, Chemical and Thermionic Properties of Thoriated Tungsten—P. Schneider. (*Le Vide*, vol. 17, pp. 182–185; March/April, 1962. In English and French.)

538.221 2746
Directional Order Induced by Tension in an Equiatomic Ni-Co Alloy—R. Vergne. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 82–84, January 4, 1961.) Annealing of the alloy under tension creates an oriented superstructure. This is not in agreement with the theory of Neel (2983 of 1954.)

538.221 2747
Domain Structures and Coercivity of Alcomax III—L. F. Bates, D. J. Craik, and E. D. Isaac. (*Proc. Phys. Soc. (London) A*, vol. 79, pp. 970–976; May 1, 1962.)

538.221:537.533.7 2748
Antiparallel Weiss Domains as Biprisms for Electron Interference Effects: Part 2—H. Boersch, H. Hamisch, K. Grohmann, and D. Wohlleben. (*Z. Phys.* vol. 167; pp. 72–82; February 28, 1962.) The experimental investigations of Part 1 (2307 of 1961) were continued on thin permalloy films. The interference pattern is calculated allowing for the domain structure and the results are compared with those of the observations.

538.221:538.569.4 2749
Interpretation of NMR Spectra in Ferromagnetic Alloys—A. M. Portis and J. Kanamori. (*J. Phys. Soc., Japan*, vol. 17, pp. 587–588; March, 1962.) A note on the possible origin of satellite line observed in nuclear-magnetic-resonance spectra.

538.221 :538.652 2750
A Connection between Crystal Anisotropy and Magnetostriction in Nickel—R. Brenner. (*Z. Naturforsch.*, vol. 17a, pp. 150–154; February, 1962.)

538.221:538.662 2751
On the Variations of Magnetization due to Heating and Cooling in the Rayleigh Domain—O. Yamada. (*C. R. Acad. Sci. (Paris)* vol. 250, pp. 4313–4315; June 27, 1960.) Heating and cooling in the Rayleigh domain causes an irreversible increase of magnetization which is

dependent only on the amount of thermal change and not on its sign.

538.221:539.23 2752

Thin Magnetic Films—J. C. Anderson. (*Electronic Tech.*, vol. 39, pp. 230–234; June, 1962.) A review of experimental properties and an outline of problems in the theory of thin films.

538.221:539.23 2753

Magnetoresistance of Thin Nickel Films: Perpendicular Effect—G. Goureaux and A. Colombani. (*C. R. Acad. Sci.* (*Paris*), vol. 250, pp. 4310–4312; June 27, 1960.) Magnetoresistance is plotted as a function of field strength and of temperature well below the ferromagnetic Curie point, and the variation of magneto-resistance with film thickness for films of different dimensions is illustrated.

538.221:539.23 2754

On the Hall Effect, due to the Magnetization, of Thin Layers of Nickel—G. Goureaux and A. Colombani. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 517–518; January 23, 1961.)

538.221:539.23 2755

Magnetoresistance of Thin Films of Nickel: Longitudinal Effect—G. Goureaux. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 858–860; February 6, 1961.) Measurements were made as functions of the magnetizing field, the layer thickness and the temperature.

538.221:539.23:538.61 2756

Magneto-optic Detection of Ferromagnetic Domains using Vertical Illumination—A. Green and M. Prutton. (*J. Sci. Instr.*, vol. 39, pp. 244–245; May, 1962.) An improved version of the conventional Kerr-effect microscope is described in which a single lens serves as both capacitor and objective. Conventional microscope construction can be used and magnetic fields are more easily arranged than in the oblique case.

538.221:621.318.124 2757

New Remanent Structure of Magnetic Domains in $BaFe_{12}O_{19}$—H. Kojima and K. Goto. (*J. Phys. Soc. Japan*, vol. 17, p. 584; March, 1962.) An unusual domain structure exhibited by very thin single crystals of $BaFe_{12}O_{19}$ under certain conditions is illustrated.

538.221:621.318.134 2758

Creeping of the Hysteresis Loop in Ferrites Subjected to Two Magnetic Fields at Right Angles—G. Buttino, A. Cecchetti and A. Drigo. (*Nuovo Cim.*, vol. 24, pp. 324–333; April 16, 1962.) Experimental investigations were carried out on tubular ferrite specimens with superimposed longitudinal and circular magnetization. Results indicate that Néel's theory is applicable to ferrites under these conditions of magnetization.

538.221:621.318.134 2759

Infrared Faraday Effect in Yttrium Garnet—F. Gires. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 540–541; January 23, 1961.

538.221:621.318.134 2760

Magnetic Studies of some Orthoferrites—D. Treves. (*Phys. Rev.*, vol. 125, pp. 1843–1853; March 15, 1962.) A method is described for distinguishing between antisymmetric exchange and single-ion magnetocrystalline anisotropy as causes of weak ferromagnetism. Comparison of measured and calculated coefficients for Y, La and Lu orthoferrites shows the antisymmetric exchange mechanism to be predominant.

538.221:621.318.134 2761

Ferrimagnetic Structure of a Magnetite Crystal as Revealed by Electron Diffraction—S. Yamaguchi. (*Phys. Rev.*, vol. 126, pp. 102–103; April 1, 1962.)

538.221:621.318.134:534.283-8 2762

Temperature Dependence of Microwave Acoustic Losses in Yttrium Iron Garnet—E. G. Spencer, R. T. Denton, and R. P. Chambers. (*Phys. Rev.*, vol. 125, pp. 1950–1951; March 15, 1962.) Measurements on single crystal YFe garnet at 500 and 1000 Mc are described.

538.221:621.318.134:538.569.4 2763

Ferrimagnetic Resonance in Polycrystalline Nickel Ferrite—W. B. Nash and K. J. Standley. (*Proc. Phys. Soc.* (*London*) *A*, vol. 79, pp. 981–986; May 1, 1962.) A study of resonance in Ni ferrite spheres in the temperature range 20°–500°C at wavelengths close to 3.13, 1.25 and 0.87 cm.

538.221:621.318.134:538.569.4 2764

Spin-Wave Propagation and the Magneto-elastic Interaction in Yttrium-Iron Garnet—J. R. Eshbach. (*Phys. Rev. Lett.*, vol. 8, pp. 357–359; May 1, 1962.) Direct observation of the propagation of pulses of microwave energy via short-wavelength spin waves is reported. The experiments were carried out at 9420 Mc over a temperature range $\sim$0°–300°K.

538.221:621.318.134:538.614 2765

Study of Faraday Rotation in Ferrites in the Region of Ferrimagnetic Resonance—M. C. Vigneron. (*J. phys. radium*, vol. 23, pp. 129–130; February, 1962.) The equipment is described and some results are given in graphical form.

538.222:538.569.4 2766

Paramagnetic Resonance of Ni^{2+} and Ni^{3+} in TiO_2—H. J. Gerritsen and E. S. Sabisky. (*Phys. Rev.*, vol. 125, pp. 1853–1859; March 15, 1962.)

621.3:061.3 2767

International Conference on Components and Materials used in Electronic Engineering—(*Proc. IEE*, pt. B, vol. 109, suppl. nos. 21 and 22, pp. 1–294 and 295–630; June, 1961.) The text is given of the papers presented and the session discussions at the conference held in London, June 12–16, 1961.

621.318.2 2768

The Stability of Permanent Magnets—C. E. Webb. (*Proc. IEE*, pt. C, vol. 108, pp. 317–324; September, 1961.) A report of tests made to compare the magnetic stability of representative isotropic and anisotropic alloys.

621.771.8:621.382.032.27 2769

Trace-Plating—a New Semiconductor-Device Fabrication Technique—W. Rindner and J. M. Lavine. (*Solid-State Electronics*, vol. 5, pp. 85–88; March/April, 1962.) The technique involves the deposition of fine traces of soft metals by abrasion on semiconductor or insulating surfaces and their subsequent plating. Its feasibility for fabrication of small electrodes and transistors is demonstrated.

666.1.037.5:621.3.032.53 2770

A New Chromium-Free Refinable Iron-Nickel Fusible Alloy for the Construction of Microwave Valves—W. Düsing. (*Telefunken Ztg.*, vol. 34, pp. 64–68; March, 1961. English summary, p. 81.) A Fe-Ni-Mn alloy (47/48/5) is described with properties suitable for making glass/metal seals, particularly to lead glass. The characteristics of other Fe-Ni alloys are also given.

669.046.5 2771

Zone Melting—W. G. Pfann. (*Science*, vol. 135, pp. 1101–1109; March 30, 1962.) Details of the technique with examples of its application.

MATHEMATICS

519.621.372.63 2772

On a Generalization of Kron's Method for Active Networks—L. Castagnetto and J. C. Matheau. (*C. R. Acad. Sci.* (*Paris*), vol. 252, pp. 85–86; January 4, 1961.) A method is proposed for writing directly the solution of the most general active network.

MEASUREMENTS AND TEST GEAR

621.3.018.41.029.65(083.74) 2773

Simple Millimetre-Wave Frequency Standard—A. F. Dietrich. (*Rev. Sci. Instr.*, vol. 33, pp. 486–487; April, 1962.) A standard suitable for the calibration of wavemeters in the 50–60-Gc range.

621.317.1.029.6 2774

Microwave Measurements.—H. A. Wheeler. (Proc. IRE, vol. 50, pp. 1207–1214; May, 1962.) Techniques and equipment are classified and future developments outlined.

621.317.31.029.6 2775

Image Loop Probe for Measurement of Microwave Surface Currents—R. W. Dressel. (*Rev. Sci. Instr.*, vol. 33, pp. 311–314; March, 1962.) Design and performance data are given for a screened loop-type probe. The accuracy achieved at 10 Gc in measurement of the phase, amplitude and direction of surface currents is within ± 5 per cent, ± 2 per cent and ± 3 per cent, respectively.

621.317.33:537.226 2776

A New Electrode System for the Measurement of Surface Resistance—K. C. Kao. (*J. Sci. Instr.*, vol. 39, pp. 208–209; May, 1962.) The system eliminates the effect of shunt resistances caused by indirect current paths both across the remaining surface and through the bulk of the dielectric system. A uniform field is produced on the surface under consideration by avoiding fringe effects due to electrode edges.

621.317.33:537.311.33 2777

Circuit to Facilitate the Measurement by the Four-Probe Method of the Resistivity of Silicon in the Range 0.002 to 10,000 Ω cm—A. L. Barry and W. D. Edwards. (*J. Sci. Instr.*, vol. 39, pp. 119–121; March, 1962.) Measurements are made to an accuracy within ± 4 per cent using a 400-cps current which can be varied from 0.12 μa to 2.5 ma. The waveform of the signal at the voltage probes is monitored as an indication of correct operation.

621.317.33:537.311.33 2778

The Accuracy of Four-Probe Resistivity Measurements on Silicon—J. K. Hargreaves and D. Millard. (*Brit. J. Appl. Phys.*, vol. 13, pp. 231–234; May, 1962.) Causes of error in the technique are elucidated and precautions necessary for accurate results are considered.

621.317.341:621.315.221.029.6 2779

Attenuation Measurement on High-Frequency Cables and Transmission Lines—G. Bittner. (*Arch. tech. Messen*, no. 314, pp. 57–58; March, 1962.) A simple method is described which is based on the measurement of power loss and is advantageous for use at Gc frequencies.

621.317.38.029.64:538.632:537.311.33 2780

A Proposed New Method of Measuring Microwave Power and Impedance using Hall Effect in a Semiconductor—H. E. M. Barlow. (*Proc. IEE*, pt. B, vol. 109, pp. 286–289; May, 1962.) The Hall effect produced by the HF electric field in the presence of a steady applied magnetic field is first measured, then the Hall effect produced by the HF magnetic field is obtained when a steady current is applied. This gives information about the electric and magnetic vectors from which the power density and the wave impedance can be computed.

621.317.4:539.23 2781

New Anisotropy Recorder for Ferromagnetic Thin Films—R. Collette. (*Rev.. Sci. Instr.*, vol. 33, pp. 450–454; April, 1962.) The instrument gives a rapid measurement of the anisotropy over a 360° range of film rotation.

621.317.42 2782

Problems Arising in the Measurement of Small Magnetic Forces at Low Temperatures by the Faraday Method—A. N. Gerritsen and D. H. Damon. (*Rev. Sci. Instr.*, vol. 33, pp. 301–307; March, 1962.) The accuracy of measurements of magnetic susceptibility below 4.2°K may be greatly affected by the presence of the exchange gas. Some precautions are suggested.

621.317.444:550.380.8 2783

Rubidium Vapour Magnetometer—L. W. Parsons and Z. M. Wiatr. (*J. Sci. Instr.*, vol. 39 pp. 292–300; June, 1962.) The instrument measures the total geomagnetic field and records its variations continuously over the range 0–1 cps, with sensitivity 5×10^{-6} oersteds.

621.317.7:621.391.822 2784

Low-Frequency Standard Noise Source—A. C. Macpherson. (*Rev. Sci. Instr.*, vol. 33, pp. 386–387; March, 1962.) The device uses two 20-kΩ wire-wound potentiometers one in liquid nitrogen, the other at room temperature. They are coupled mechanically and electrically to give rapid independent control of source impedance and noise temperature for the frequency range 1–1000 cps.

621.317.725:621.391.822 2785

Mean-Square Voltmeter—S. M. Bozic. (*J. Sci. Instr.*, vol. 39, pp. 210–213; May, 1962.) The scale of the meter is linearly proportional to the mean-square applied voltage, 1V rms at the input giving full scale deflection. The circuit is suitable for noise voltage measurements in the range 20 cps 10 Mc.

621.317.737:621.372.413 2786

An Accurate Method for the Dynamic Measurement of the *Q*-Factor of Resonant Cavities—K. Leibrecht. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 3966–3968; June 13, 1960.) Equipment is described with which rapid measurements of cavity gain factors can be made at frequencies up to 9 Gc. Absorption and dispersion curves are displayed on a CRO.

621.317.77 2787

Electronic Device for Measuring the Phase of a Voltage Impulse with reference to the Origin of its Low-Frequency Periodic Carrier Frequency—C. Curie and J. Cava. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 692–694; January 30, 1961.)

621.317.79:537.311.33:535.215 2788

Measurement of Photomagnetic Susceptibility by Means of a Vibrating Reed Balance—J. O. Kessler and A. R. Moore. (*Rev. Sci. Instr.*, vol. 33, pp. 478–483; April, 1962.) A method for studying changes in the magnetic susceptibility of semiconductors with modulation of incident light.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.715:621.374.32 2789

Interferometry in Length Measurements—J. L. Goldberg and R. H. Brockman. (*Electronic Tech.*, vol. 39, pp. 140–144, 186–191, and 238–243; April–June, 1962.)

Part 1—Electronic Counting of Interference Fringes.

Part 2—Practical Considerations.

Part 3—Waveform Generator Suitable for Reversible Counting.

537.311.33:536.53 2790

Use of Germanium as a Second-Sound Receiver—H. A. Snyder. (*Rev. Sci. Instr.*, vol. 33, pp. 467–469; April, 1962.) Doped Ge crystals show results superior to those of other receivers for measuring temperature waves in liquid He at frequencies up to 10 kc. See *e.g.*, *Phys. Rev.*, vol. 71, pp. 600–605; May 1, 1947. (Lane, *et al.*).

621.384.62 2791

An Experimental Proton Linear Accelerator, using a Helix Structure—D. P. R. Petrie, R. Bailey, D. G. Keith-Walker, H. Longley, and D. R. Chick. (*Proc. IEE*, pt. C, vol. 108, pp. 424–432; September, 1961.) See 571 of 1958 (Chick, *et al.*).

621.385.833 2792

Experiments on the Elimination of Charges on Transmission Objects by Additional Irradiation with Slow Electrons—H. Mahl and W. Weitsch. (*Z. Naturforsch.*, vol. 17a, pp. 146–150; February, 1962.) Improvements in the images produced by small-angle diffraction of electron beams (see *e.g.*, 1283 of 1961) are illustrated.

621.385.833 2793

Energy-Selecting Electron Microscope—H. Watanabe and R. Uyeda. (*J. Phys. Soc. Japan*, vol. 17, pp. 569–570; March, 1962.) Note on a technique for obtaining an electron-microscope image formed by mono-energetic electrons of any preselected energy value.

621.385.833:621.3.032.269.1 2794

Characteristics of a Point-Cathode Triode Electron Gun—D. W. Swift and W. C. Nixon. (*Brit. J. Appl. Phys.*, vol. 13, pp. 288–293; June, 1962.)

669.046.5:537.533 2795

Technique for the Zone Melting of Insulators—L. Neumann and R. A. Huggins. (*Rev. Sci. Instr.*, vol. 33, pp. 433–434; April, 1962.) An electron-beam technique is described for the zone melting of materials with very high melting points. The build-up of a negative charge on the sample is prevented by secondary emission.

681.828:621.382.3 2796

Transistorized Organ Generators—A. Douglas. (*Electronic Eng.*, vol. 34, pp. 388–394; June, 1962.) Full circuit diagrams are included.

PROPAGATION OF WAVES

621.371.011.21 2797

The Surface Impedance Concept and the Structure of Radio Waves over Real Earth—Z. Godzinski. (*Proc. IEE*, pt. C, vol. 108, pp. 362–373; September, 1961.) The surface impedance concept is generalized to the case of an arbitrarily inhomogeneous earth; practical applications to the measurement of earth constants and geological prospecting are discussed. 77 references.

621.391.812.62.029.64 2798

Microwave Propagation Test in Mirage District—S. Ugai, Y. Kaneda, and T. Amekura. (*Rev. Elect. Comm. Lab. Japan*, vol. 9, pp. 687–717; November/December, 1961.) Tests in an area subject to mirage phenomena demonstrate that phase-controlled space-diversity reception is effective in the relief of deep interference fading. The forecasting of weather liable to cause abnormally deep fading would be useful to communication engineers.

621.391.812.63+534.2–14 2799

Ray Paths in Inhomogeneous Anisotropic Media—R. F. Mullaly. (*Aust. J. Phys.*, vol. 15, pp. 96–105; March, 1962.) Expressions are developed for the curvature of a ray path and are applied to a) radio rays in an ionized medium pervaded by a nonuniform magnetic field, and b) acoustical rays in a moving fluid.

621.391.812.63 2800

The Geometry of Radio Reflections from Field-Aligned Ionization Irregularities in the Ionosphere—E. W. Dearden. (*J. Atmos. Terr. Phys.*, vol. 24, pp. 375–384; May, 1962.) The aim is to determine the surface formed by points at which the ray paths of radio signals emanating from a common source intersect the earth's magnetic field at a given angle. Chapman's treatment (1357 of 1953) has been generalized for an arbitrary angle of incidence.

621.391.812.63 2801

High-Frequency, High-Power, Pulse Signals from an Australian Transmitter—J. F. Ward. (*Nature*, vol. 194, pp. 518–521; May 12, 1962.) Propagation anomalies were examined using a 25-Mc pulse transmission from southern Victoria, Australia. Pulse patterns were received and photographed in Australia, southeast Asia, and the United Kingdom.

621.391.812.63.029.45/.51 2802

Ionospheric Reflection Processes for Long Radio Waves: Part 1—B. S. Wescott. (*J. Atmos. terr. Phys.*, vol. 24, pp. 385–399; May, 1962.) The calculations are based on the Darwin-Hartree microscopic theory of reradiation. The formulas obtained are applied to a horizontally stratified ionosphere with an exponential electron density profile. Typical results are shown graphically.

621.391.812.63.029.45:550.385.4 2803

Very-Low-Frequency Phase Perturbations Observed during Geomagnetic Storms—C. F. Sechrist, Jr. (*J. Geophys. Res.*, vol. 67, pp. 1685–1686; April, 1962.) Phase shifts observed at 18-kc during geomagnetic storms are related to vertical movements of 200–500 m in the lower ionosphere. See 668 of February (Sechrist, and Felperin.)

RECEPTION

621.376 2804

Harmonic Analysis in Systems using Phase-Sensitive Detectors—A. M. Russell and D. A. Torchia. (*Res. Sci. Instr.*, vol. 33, pp. 442–444; April, 1962.) An analysis is made of a system of modulation and synchronous detection, the modulated signal being expressed as a Taylor expansion and as a Fourier series.

621.391.812.624 2805

The Frequency and Duration of Individual Fade-Outs in Ionospheric Scatter Links—J. Grosskopf and L. Fehlhaber. (*Nachrtech., Z.*, vol. 15, pp. 71–78; February, 1962.) From a statistical analysis of short-term fading count-carried out on scatter links operating in the fref quency range 100–2000 Mc, the occurrence os deep fading and the distribution of fading duration over long periods can be determined.

621.391.82 2806

Analysis and Prediction of Interference Factors and their Impact on the Future—R. M. Showers. (Proc. IRE, vol. 50, pp. 1316–1320; May, 1962.) The factors contributing to spectrum crowding are outlined, and methods of assessing and predicting the extent of spectrum utilization are discussed.

621.396.62:523.164 2807

The Italian Cross Radiotelescope: Part 2—Preliminary Design of the Receiver—Gelato, Rosatelli, and Sinigaglia. (See 2622.)

621.396.621 2808

The 'Stenode'—L. A. Moxon. (*Wireless*

World, vol. 68, pp. 300–304; July, 1962.) A simple circuit of 30 years ago incorporating a modern high-stability local oscillator gives an enhancement of AM carrier signal relative to the sidebands; this eliminates distortion due to selective fading. The circuit can be adjusted to provide 15–20-db rejection of one sideband.

STATIONS AND COMMUNICATION SYSTEMS

621.395.44:621.375.4 2809

The Equipment of Carrier-Current Telecommunication Links with Transistor Amplifiers—H. Krause. (*Nachrtech.*, vol. 12, pp. 54–59; February, 1962.) Discussion of operational and economic advantages of the use of transistor instead of valve amplifiers.

621.395.44:621.391.827 2810

The Problem of the Interference Effect on Balanced Carrier-Current Cables caused by Broadcast Transmitters—H. G. Prenzlow. (*Nachrtech.*, vol. 12, pp. 60–63; February, 1962.) Measurements were made on cable systems in the vicinity of medium- and long-wave transmitters operating at frequencies in the range 100–1100 kc. Means of reducing this interference are mentioned.

621.396.43:551.507.362.2 2811

Communications via Satellites—(*Electronic Eng.*, vol. 34, pp. 382–387; June, 1962.) A description of the British Post Office equipment at Goonhilly Downs and of the equipment and functions of the Telstar satellite.

621.396.43:551.507.362.2 2812

Satellite Communication Links—S. A. W. Jolliffe and W. L. Wright. (*Point to Point Telecommun.*, vol. 6, pp. 6–28; June, 1962.) A review of important aspects of communication systems using earth-satellite repeaters.

621.396.43:551.507.362.2 2813

Optimum System Engineering for Satellite Communication Links with Special Reference to the Choice of Modulation Method—W. L. Wright and S. A. W. Jolliffe. (*J. Brit. IRE*, vol. 23, pp. 381–391; May, 1962.) SSB, wide-band FM and pulse-code methods are considered. The advantages of PCM are shown and expected SNR values in practical 600-channel systems are evaluated.

621.396.65:621.391.812.624 2814

Transhorizon Radio Links—(*Rev. tech. Comp. franç. Thomson-Houston*, no. 34, pp. 1–127; June, 1961.) Seven papers are presented describing equipment, including aerials, for tropospheric scatter communication, with particular reference to recent improvements in commercial apparatus.

621.396.65.029.63 2815

4-Gc/s Radio-Link System (FM960-TV/4000) for 960 Telephony Channels and Television—(*Telefunken Ztg.*, vol. 34, pp. 285–347; December, 1961.)

Part 1—Radio-Link System—E. Willwacher. (pp. 285–294; English summary, p. 352).

Part 2—Modulation and I. F. Section.—H. Oberbeck. (pp. 295–304; English summary, pp. 352–353.)

Part 3—Microwave Section—E. Willwacher. (pp. 304–313; English summary. p. 353).

Part 4—Aerial Installations and Feeders—E. Schüttlöffel. (pp. 314–325; English summary p. 354).

Part 5—Transmission Characteristics of a Radio-Link System—H. Oberbeck and R. Steinhart. (pp. 326–340; English summary, pp. 354–355).

Part 6—Transmission of the Sound Channel—R. Heer (pp. 341–344; English summary, p. 355).

Part 7—Equipment for Diversity Reception—H. Junghans, and A. Koreis. (pp. 344–347; English summary, p. 355).

See 3150 of 1961 for 2-Gc radio-link equipment.

621.396.65.029.64 2816

7-Gc/s Radio Link Installation for 120 Telephony Channels: FM 120/7000—E. Willwacher. (*Telefunken Ztg.*, vol. 34, pp. 58–64; March, 1961. English summary, pp. 80–81.) Description of the main features of the equipment.

621.396.97+621.397.13 2817

Broadcasting Developments Now Taking Place—O. Reed, Jr. (Proc. IRE, vol. 50, pp. 837–847; May, 1962.) Review of rules and regulations and allocations covering AM and FM sound broadcasting and television systems. Developments in systems and techniques are also discussed.

621.396.97:534.76 2818

Stereophony—(See 2529.)

SUBSIDIARY APPARATUS

621.3.087.4:621.395.625.3 2819

A Theoretical Solution for the Magnetic Field in the Vicinity of a Recording Head Air Gap—E. E. Francis and T. C. Ku. (*IBM J. Res. & Dev.*, vol. 6, pp. 260–262; April, 1962.) A solution is obtained for pole pieces having an arbitrary angle θ, and is compared with Booth's solutions for $\theta=0$ and $\pi/2$ (768 of 1953). The optimum angle for the usual recording distance is $(7/18)\pi$.

621.311.69:621.383.5 2820

Preparation and Study of Silicon Solar Cells of High Efficiency—H. Valdman. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 246–248; January 9, 1961.) An efficiency of 14 per cent has been attained by diffusion of P into Si.

621.314.63:621.373.51 2821

An Audio-Frequency High-Power Generator employing Silicon Controlled Rectifiers—R. Thompson. (*Proc. IEE*, pt. B, vol. 109, pp. 249–258; May, 1962.) The design and construction of a 2-kw generator is discussed. This type of circuit has fewer components and is more reliable than conventional transistor units.

621.314.632:621.382.22 2822

Phenomenological Theory of Turnover in Point-Contact Rectifiers—T. Numata. (*J. Phys. Soc. Japan*, vol. 17, pp. 447–454; March, 1962.) Minority carriers are shown to play an important part in the turn-over behavior of semiconductor diodes, and reasonable agreement with previous experimental results is obtained.

621.314.634 2823

Capacitance Measurements on Selenium Rectifiers—U. Dolega. (*Z. Phys.*, vol. 167, pp. 46–52; February 28, 1962.) An *n*-type CdSe film is formed between the metal electrode and the *p*-type Se during the manufacture of the rectifiers. This *n-p* contact determines the rectifier characteristics. From capacity measurements the acceptor and donor concentrations in Se and CdSe, respectively, and the thickness of the CdSe film can be obtained.

621.316.721/.722:621.382.3 2824

Drifts in Silicon Transistors as a Function of Operating Time and of Temperature—G. Giralt and J. C. Polisset. (*C. R. Acad. Sci. (Paris)*, vol. 252, pp. 519–521; January 23, 1961.) The general results obtained for Ge transistors in stabilized power supplies [1782 of 1960 (Cassignol, *et al.*)] are extended to Si devices.

TELEVISION AND PHOTOTELEGRAPHY

621.397:061.3 2825

International Television Conference—(*Wireless World*, vol. 68, pp. 306–307; July, 1962.) A short review of some of the 130 papers read at the International Television Conference, London, May 31–June 7, 1962.

621.397.13+621.396.97 2826

Broadcasting Developments Now Taking Place—Reed. (See 2817.)

621.397.6.018.782.3 2827

Considerations on the Suitable Measurement and Tolerance Limits of the Transient Response of Television Equipment—H. Springer. (*Nachrtech. Z.*, vol. 15, pp. 57–62; February, 1962.) Comparison of test methods using a) a 250-kc square wave, and b) a sine-squared pulse and bar signal. A simplified pulse test method combined with a measurement of phase and amplitude distortion is proposed.

621.397.612:621.375.2 2828

The Design of a Group of Plug-In Television Studio Amplifiers—K. J. Austin. (*BBC Engrg. Div. Monographs*, no. 41, pp. 5–19; April, 1962.)

621.397.621 2829

The Relation between Noise Bandwidth and Transient Response of Flywheel Synchronization Circuits in Television Receivers—H. Grosskopf and A. Grünewald. (*Nachrtech. Z.*, vol. 15, pp. 66–69; February, 1962.) The differences in the effects of additive and modulation-type interference in flywheel synchronization systems are discussed. The noise bandwidth of the filter should not be reduced excessively if jitter is to be minimized. See also 351 of 1961 (Grosskopf & Springer).

621.397.712 2830

The B.B.C. Television Centre and its Technical Facilities—F. C. McLean, H. W. Baker, and C. H. Colborn. (*Proc. IEE*, pt. B, vol. 109, pp. 197–219; May, 1962. Discussion, pp. 219–222.)

621.397.74:621.391.833 2831

Picture Quality Assessment and Waveform Distortion Correction an Wired Television Systems—B. W. Osborne. (*J. Brit. IRE*, vol. 23, pp. 399–404; May, 1962.) The design and use of a compact, low-cost video-frequency transversal equalizer are described and an equalizer capable of correction video waveform distortion on a modulated carrier is considered.

621.397.743:621.315.212 2832

Carrier-Current Television Transmission over Coaxial Cables for Short Links—B. Precht. (*Nachrtech.*, vol. 12, pp. 64–67; February, 1962.) Description of system and equipment for use in outside-broadcast links.

TRANSMISSION

621.396.61:621.398 2833

Remote-Control Equipment for Short-Wave Transmitters—E. Baranowski. (*Nachrtech. Z.*, vol. 15, pp. 79–83; February, 1962.) Design and operational details of equipment for the overseas transmitting station Elmshorn are given.

TUBES AND THERMIONICS

621.382:539.23 2834

Schottky Emission through Thin Insulating Films—P. R. Emtage and W. Tantraporn. (*Phys. Rev. Lett.*, vol. 8, pp. 267–268; April 1, 1962.) Temperature-dependent Schottky emission predominates over tunnel emission in thick films and for high work functions. The measurements provide two ways of estimating metal-insulator work functions.

621.382:621.375.9 2835

Current Gain in Metal-Interface Amplifiers—J. M. Lavine and A. A. Iannini. (*Solid-State Electronics*, vol. 5, pp. 109–110; March/April, 1962, plates.) The large current-transfer ratio (~0.9) in a metal-interface amplifier is investigated and shown to be a result of depletion-layer transistor action.

621.382.2/.3:621.317.7 2836

Microprobing of Functioning Semi-conductor Devices for Internal Voltage and Current Distributions—D. E. Sawyer. (*Solid-State Electronics*, vol. 5, pp. 89–96; March/April, 1962.) The devices are potted and sectioned to expose the internal active region, then polished and etched. A fine-tipped tungsten carbide probe is lowered to the exposed surface and the probe voltages over a rectangular area are recorded for the bias conditions of interest, using a low-duty-cycle source.

621.382.23 2837

Characterization of Tunnel Diode Performance in Terms of Device Figure of Merit and Circuit Time Constant—L. Esaki. (*IBM J. Res. & Dev.*, vol. 6, pp. 170–178; April, 1962.) Tunnel-diode oscillation, flip-flop switching and Goto twin operation have been characterized on the basis of numerical integration of a nonlinear differential equation representing transient behavior in the simplified, lumped equivalent circuit.

621.382.23 2838

A Proposed Technique for Stabilization of Tunnel Diodes—E. G. Cristal. (*Microwave J.*, vol. 5, pp. 108–113; April, 1962.) An analysis of a method for correcting short-circuit instability of a tunnel diode.

621.382.23:621.372.44 2839

Optimum Performance of Parametric Diodes at S Band—B. J. Robinson and J. T. de Jager. (*Proc. IEE*, pt. B, vol. 109, pp. 267–276; May, 1962.) The Fourier expansion of the inverse of the barrier capacitance describes the potential noise performance better than the Fourier expansion of the capacitance itself. Experimental results give a good agreement with theoretical analysis over a wide range of pump power. The theoretical analysis suggests a definition of a simple static figure of merit which is easily measured at microwave frequencies.

621.382.3 2840

Applications of the Charge-Control Concept to Transistor Characterization—D. E. Hooper and A. R. T. Turnbull. (*Proc. IRE (Australia)*, vol. 23, pp. 132–147; March, 1962.) Stored charge is considered as a fundamental concept and is used to derive transistor operating parameters significant in large- and small-signal applications. Extensions of the basic charge-control model to more complex devices are discussed.

621.382.333 2841

Properties of Germanium Transistors obtained by Double Diffusion—R. Deschamps. (*C. R. Acad. Sci. (Paris)*, vol. 250, pp. 4307–4309; June 27, 1960.) Base resistance and breakdown voltages have been calculated and measured, and parameters important for operation at high frequency have been examined.

621.382.333:539.12.04 2842

Low-Dose Gamma Irradiation of Alloy Junction Transistors—D. E. Vaughan. (*J. Electronics Control*, vol. 12, pp. 307–312; April, 1962.) Investigations have been carried out of the effects of small (10 kilorad) doses of ^{60}Co gamma radiation on Ge *p-n-p* alloy junction transistors, with particular reference to the effect upon gain and minority carrier lifetime.

621.362.333:621.318.57 2843

Switching Transistors—J. N. Barry. (*Electronic Tech.*, vol. 39, pp. 206–211; June, 1962.) The types of transistor circuit available for switching systems are discussed and an analysis is made of transistor parameters important in the design of fast saturating circuits.

621.382.333:621.318.57 2844

Contribution to the Theory of Switching Times of Alloy-Junction Transistors—S. Neumann. (*Arch. elekt. Übertragung*, vol. 16, pp. 105–116; March, 1962.) (With reference to the work of Ebers and Moll (884 of 1955) and Beaufoy and Sparkes (308 of 1958) a simple exponential desaturation process is formulated for the current- or voltage-controlled transistor. The charge-controlled transistor can be regarded as a generalization of the current- or voltage-controlled type.

621.382.333.33.012.8 2845

Equivalent Circuits and Four-Terminal Parameters of High-Frequency Transistor—S. Kawaguchi and M. Hirai. (*Rev. Elect. Comm. Lab. Japan*, vol. 9, pp. 661–686; November/December, 1961.) Frequency-dependent characteristics of the elements of the small-signal equivalent circuits are calculated. A method of obtaining and measuring the equivalent circuit from earthed-base *h* parameters is described and data from various types of transistor are tabulated.

621.383+621.385.83 2846

Beam-Deflection and Photo Devices—K. Schlesinger and E. G. Ramberg. (PROC. IRE, vol. 50, pp. 991–1005; May, 1962.) A historical review including important recent advances. 163 references.

621.383.2 2847

Calculations of the Noise Figure and Power Gain of Photodiodes—M. J. O. Strutt. (*Arch. elekt. Übertragung*, vol. 16, pp. 158–162; March, 1962.) The photodiode is considered as a radiation receiver, *e.g.*, for the measurement of small temperature changes (*Arch. elekt. Übertragung*, vol. 15, pp. 355–358; August, 1961) and an equivalent circuit is derived which takes account of input and gain fluctuations and the output characteristics of the diode. In practice the noise figure may be 2 and the available gain 15.

621.385.032.213.23 2848

The Conductivity of Oxide Cathodes: Part 10—Spontaneous Generation of Negative Ions—G. H. Metson. (*Proc. IEE*, pt. C, vol. 108, pp. 438–449; September, 1961.) A Ba-Sr oxide matrix is found to be thermally unstable at 1020°K due to oxygen-ion generation by the Sr oxide component. The reaction is fundamental and factors determining the equilibrium concentration of free oxygen ions are described. Part 9: 2497 of July (Metson and Holmes).

621.385.032.213.23 2849

The Conductivity of Oxide Cathodes: Part 11—Thermal Stability of the Alkaline-Earth Oxides—G. H. Metson and H. Batey. (*Proc. IEE*, pt. C, vol. 108, pp. 450–454; September, 1961.) In a massive evaporation from a platinum substrate at 1200°K Ba oxide is removed in the form of unchanged molecules while the Sr and Ca oxides leave the substrate in elemental form. Part 10: 2848 above.

621.385.1 2850

Low-Voltage Valves of High Mutual Conductance—F. H. Reynolds and R. E. Hines. (*Proc. IEE*, pt. B, vol. 109, pp. 259–266; May, 1962.) Receiving type tubes have been developed suitable for operation at supply voltages in the region of 20–50 v and having workable mutual conductances up to about 20 ma/v. The requisite electrode geometry is attained by the use of a new tube component known as a double grid.

621.385.33.012 2851

The Influence of Fluctuations of the Penetration Factor on Valve Characteristics—H. Stahl. (*Nachrtech. Z.*, vol. 15, pp. 84–88; February, 1962.) The slope and distortion factor of a triode are calculated, whose penetration factor varies owing to structural imperfections. A method is given for assessing the possibilities of improving the performance of high-slope low-distortion triodes by structural changes.

621.385.4.029.63 2852

A Rugged 3-kMc/s, 40-Watt Transmitting Tetrode—J. J. Hamilton. (*J. Brit. IRE*, vol. 23, pp. 405–411; May, 1962.) Unusual design features of a planar tetrode are described, and its operating parameters and characteristics are given.

621.385.6 2853

Elektrokinetic Turbulent Energy Flow—W. Riedler. (*Arch. elekt. Übertragung*, vol. 16, pp. 129–134; March, 1962.) Theorems for dc and ac energy flow in electron beams under conditions of turbulent flow and at relativistic velocities are derived. The Manley-Rowe relations are obtained for this case. See also 2122 of June.

621.385.6 2854

Large-Signal Calculations of the Admittance of an Electron Beam Traversing a High-Frequency Gap—L. Solymar. (*J. Electronics and Control*, vol. 12, pp. 313–317; April, 1962.) Calculations of beam loading are made using formulas previously derived for the equivalent current [2095 of 1952 (Warnecke and Guénard) and for electron motion (2506 of July) in a high-frequency gap. The range of validity of the solution is given.

621.385.6.029.6 2855

Current European Developments in Microwave Tubes—A. H. W. Beck. (PROC. IRE, vol. 50, pp. 985–991; May, 1962.) Developments differing from those in the U.S.A. are noted in particular. 52 references.

621.385.623.5:621.316.726:621.376.332 2856

Microwave Discriminator for the Frequency Stabilization of a Reflex Klystron—Smith. (See 2596.)

621.385.63:621.318.2 2857

Periodic Magnetic Focusing of Solid Beams in the 10–25 kV Range—E. J. Nalos and F. K. Patton. (*Microwave J.*, vol. 5, pp. 95–100; April, 1962.) An extension of earlier work on beam focusing [*e.g.*, 1021, of 1959 (Sterrett and Hoffner)] to the case of a practical high-power traveling-wave tube.

621.385.65 2858

A Periodic Electrostatic Focusing System for Travelling-Wave Tubes—W. Henne. (*Arch. elekt. Übertragung*, vol. 16, pp. 83–92; February, 1962.) Details are given of a focusing system and the electron paths are computed on the basis of equipotential line measurements. Results of measurements on an experimental tube are in agreement with the theoretical results.

621.387.322 2859

Characteristics of Glow-Discharge Ref-

erence Tubes with Various Cathode Geometries and Surface Finishes—F. A. Benson and P. M. Chalmers. (*Proc. IEE*, pt. B, vol. 109, pp. 209–293; May, 1962.)

621.387.322.3 2860

A New Dekatron for Direct Operation of Digitrons—D. Reaney. (*Electronic Eng.*, vol. 34, pp. 372–376; June, 1962.) Performance characteristics and circuit details of the Type GGA10G and Type GSA10G number-display dekatrons are given.

MISCELLANEOUS

061.4:621.3 2861

1962 International Instruments, Electronics and Automation Exhibition— (*Electronic Tech.*, vol. 39, pp. 212–220; June, 1962.) A list of exhibitors and review of selected items shown at Olympia, London, May 28–June 2.

621.38/.39"1912/1962" 2862

Communications and Electronics, 1912–1962—(Proc. IRE, vol. 50, pp. 657–1448; May, 1962.) A group of over 100 papers arranged in collaboration with the Professional Groups of the IRE. The subject matter is divided into 28 sections. For abstracts of some of these papers see under the appropriate subject headings.

621.38/.39"2012" 2863

Communications and Electronics—2012 A.D.—(Proc. IRE, vol. 50, pp. 562–656; May, 1962.) A symposium of predictions prepared by Fellows of the IRE, covering anticipated developments in various fields during the period 1962–2012.

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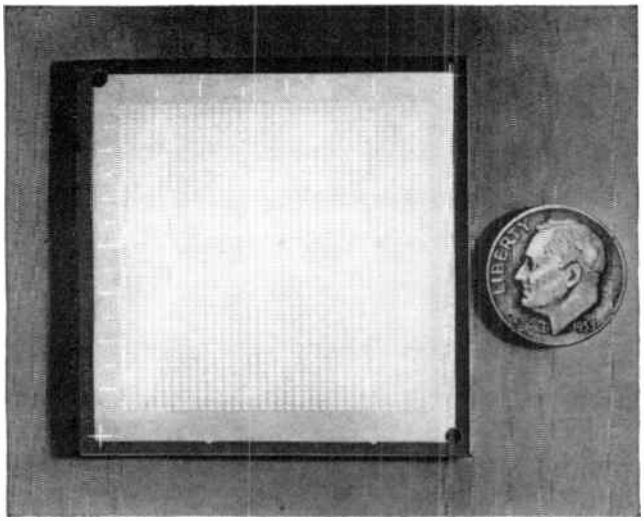
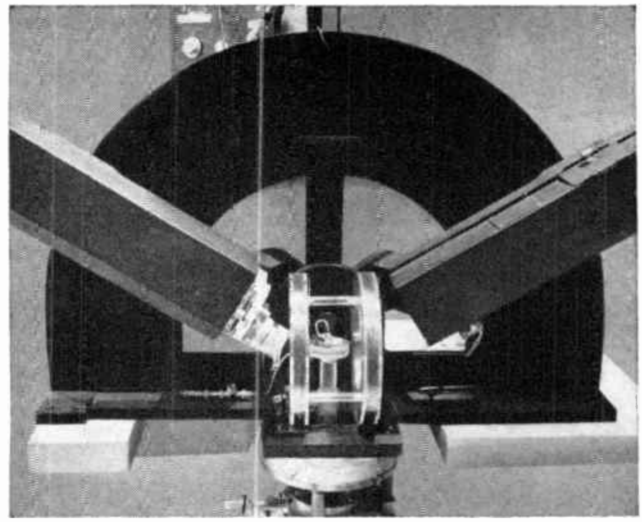


This is one of eight 2"x2" substrates in a new experimental memory containing 576 bits of information per square inch, the densest packaging ever reported for thin magnetic films.



The switching speed of this experimental thin-film device (center) is being measured by timing a polarized light beam which rotates each time direction of magnetization changes.

Computing speed is accelerating constantly. But before computers can process data, they must pass it through main storage. Unless ways are found to transfer more information in less time from main storage to central processing units, the time required to obtain stored data will limit the speed of the computer.

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More immediate gains in access time can be attained through new developments in ferrite core technology. By reducing the core size from 50 thousandths of an inch to 30 thousandths of an inch in outside diameter, IBM engineers have created a 1.2 million bit magnetic core memory with a cycle time of only 2 microseconds. A device which contributed greatly to this development, a load-sharing matrix switch, was also instrumental in the creation of a 74,000 bit "scratch pad" ferrite core memory capable of a read-write cycle time of less than 700 nanoseconds. This matrix switch makes it possible for the switching-power load to be shared by several drivers at once, thus reducing the total power requirements.

The efficiency of computing systems can be increased by improving the design of their memory structures as well as through the development of new components. IBM engineers are developing nondestructive read-out techniques which can reduce the number of machine operations required in thin-film and ferrite core memories. They have formulated addressing systems in which machine-word lengths vary according to the natural lengths of the bits of information being stored. They have devised associative memory techniques which retrieve information on the basis of related data rather than specified addresses. Out of several developments like these, which reduce machine references to memory and simplify programming, may come the memory systems of the future.

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Nash, W. B., Pomona, Calif.
Papasifakis, C. E., Havertown, Pa.
Parlanti, C. A., Natick, Mass.
Pile, D. W., Van Nuys, Calif.
Schroeder, A. H., Wappingers Falls, N. Y.
Searle, L. V., Northridge, Calif.
Wallett, R. A., Rochester, N. Y.
Williams, R. D., Palo Alto, Calif.

Transfer to Member

Blaine, W. E., Atlanta, Ga.
Carpenter, J. F., Chenango Bridge, N. Y.
Cave, J. H., Ferguson, Mo.

(Continued on page 84A)

Membership

(Continued from page 82A)

Cimera, V. J., Melrose Park, Ill.
Craigue, A., Barrington, Ill.
Cranston, E. M., Anaheim, Calif.
Diger, J. E., Richfield, Minn.
Ehrlich, W. E., Red Bank, N. J.
Frederick, C. L., Jr., Wheaton, Md.
Guth, R. U., Cleveland, Ohio
Hammond, R. L., Huntsville, Ala
Lee, W. W. S., Midwest City, Okla.
Long, A. B., Beaumont, Tex.
Martin, H. R., Syracuse, N. Y.
McWilliams, P. B., Columbus, Ohio
Mead, E. S., Hanover Center, Mass.
Mendus, W. M., Annandale, Va.
Merren, G. T., Albuquerque, N. M.
Mitchell, B. C., Baltimore, Md.
Morrisey, J. A., Washington, D. C.
Mousel, G. W., El Segundo, Calif
Murray, W. J., Bradford, Pa.
Nichols, H. F., Jr., South Hamilton, Mass.
Odom, C. H., Glendora, Calif.
Pearce, G. E., Shawinigan, Que., Canada
Piker, J. S., Texas City, Tex.
Sandefer, S. R., Garland, Tex.
Savalli, F. P., Largo, Fla.
Schiff, A., Lake Katrine, N. Y.
Shellkopf, R. E., Baltimore, Md.
Sonn, M., Norwalk, Conn.
Spence, R. C., Parkersburg, W. Va.
Spencer, R. J., Brooklyn, N. Y.
Squire, R. C., Santa Maria, Calif.
Stein, B. M., East Meadow, L. I., N. Y.
Stewart, A. G., Victoria, B. C., Canada
Symans, J. H., Reseda, Calif.
Warner, C. W., Smiths Falls, Ont., Canada

Admission to Member

Adams, V. J., Jr., Parsippany, N. J.
Agam, Y., Ramat Gan, Israel
Albers, P. H., Cedar Grove, N. J.
Ambler, L. G., Santa Maria, Calif.
Anders, E. C., South Braintree, Mass.
Arya, V. K., Pilani (Rajasthan), India
Attwood, T. V., Liverpool, England
Ausband, R. W., Clearwater, Fla.
Avera, D. S., Jr., Winston-Salem, N. C.
Bailey, H. H., Winston-Salem, N. C.
Baker, C. H., Stamford, Conn.
Barash, L. S., Greensburg, Pa.
Barber, S. E., Ridgecrest, Calif.
Behn, R. C., Columbus, Ohio
Belasco, M., Dallas, Tex.
Bell, F. A., Smithtown, L. I., N. Y.
Bilyk, Z., Long Island City, N. Y.
Blacklock, R. L., Buffalo, N. Y.
Blair, B. E., Boulder, Colo.
Blair, J. A., Dover, N. J.
Boom, D. P., Riverside, N. J.
Boyes, R. F., Tucson, Ariz.
Brackett, C. F., Lexington, Mass.
Bradley, J. D., Cambridge, Ohio
Brady "K" M., Sandy, Utah
Brockmeyer, C. H., Littleton, Colo.
Brockus, W. G., South Pasadena, Calif.
Brown, E. S., Grand Rapids, Mich.
Brown, J. R., West Asheville, N. C.
Brownlee, G. H., London, England
Brubaker, L. N., Washington, D. C.
Bruinsma, P. J. T., Ann Arbor, Mich.
Bryce, W. E., Merrisville, Pa.
Burgess, B. C., Liverpool, N. Y.
Burney, A. K., Karachi, Pakistan
Burns, W. S., Jr., Winston-Salem, N. C.
Byrd, J. D., Jr., Beltsville, Md.
Cameron, D. A., Ottawa, Ont., Canada
Campbell, D. V., Red Bank, N. J.
Carnley, I. E., Ocean Springs, Miss.
Caswell, A. R., Winthrop, Mass.
Caussin, J-P, Dinant, Belgium

(Continued on page 86A)

JFD Trimmers and LC tuners help keep Transit Satellite transmitters on exact frequencies

Transit, the Bureau of Naval Weapon's all-weather global navigation system, is scheduled for operational use in 1962. Transit will provide ships, submarines and aircraft with the most precise method ever devised for fixing their positions.

The highly critical nature of the system's measurement functions demanded highest reliability, stability and exactness in the performance of its two frequency sources. JFD VC42GW trimmer capacitors were specified for each of the two crystal-controlled oscillators to help assure frequency stability of 2 to 4 parts in 10^{10}. JFD trimmers were used also in the frequency multiplier circuit to maintain required oscillator frequency outputs.

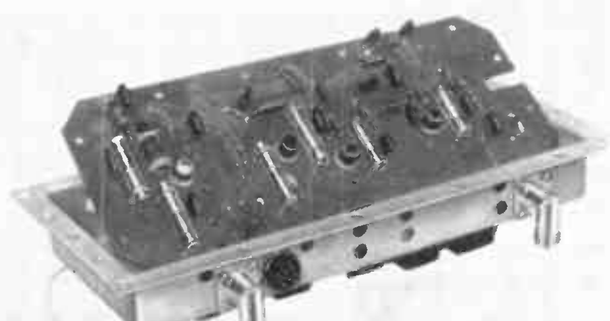
JFD LC tuners as well as trimmers were called for in both the B-system and C-system power amplifiers of the transmitter circuits and in the Transit diplexing antenna system to provide highest possible operating stability.

This is another example of how JFD precision electronic components satisfy space-challenging demands of tuning accuracy and stability under severe shock and vibration. Fewer parts, precise tolerances, patented telescoping anti-backlash adjustment are a few of the reasons why more engineers specify JFD

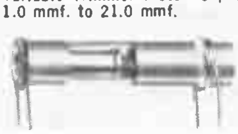
For complete information, contact your local JFD Field office or your local JFD franchised Industrial Distributor.

Applied Physics Laboratory of the Johns Hopkins University specified JFD Trimmer Capacitors and Tuners in the Transit 2-A Satellite.

JFD LC Tuners and Trimmers in Transit frequency multipliers and power multiplier amplifier circuits provide maximum tuning range in minimum space . . . high reliability and ruggedness.



JFD VC42GW actual size
Variable Trimmer Piston Capacitor
1.0 mmf. to 21.0 mmf.



JFD LC326 LC Tuner actual size
200 to 450 mc.
self-resonating frequency

JFD

JFD ELECTRONICS CORPORATION

Components Division • 6101 16th Avenue, Brooklyn, New York • Phone DEwey 1-1000 • TWX-NY25040

JFD WESTERN	JFD MIDWESTERN	JFD NORTHEASTERN	JFD CANADA
P. O. Box 3416	P. O. B. 588	Ruth Drive, P. O. Box 228	51 McCormack Street
Van Nuys, Calif.	Skokie, Illinois	Marlboro, Mass.	Toronto, Ontario, Canada
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FIXED AND VARIABLE. DISTRIBUTED AND LUMPED CONSTANT DELAY LINES • PULSE FORMING NETWORKS

Membership

(Continued from page 81A)

Chavez, R., Long Beach, Calif.
Clark, R. R., Salt Lake City, Utah
Coates, V. R., Mount Holly, N. J.
Cochran, A. L., Richardson, Tex.
Cohen, A. H., Edison, N. J.
Cohen, B. E., Canoga Park, Calif.
Coley, H. J., Atlanta, Ga.
Colson, K. R., Glendale, Calif.
Como, P. E., Youngwood, Pa.
Conway, F. J., Mobile, Ala.
Corp, A. F., Rochester, N. Y.
Cosgrove, M. J., Hamilton, Ont., Canada
Craver, J. S., North Wales, Pa.
Court-Hampton, D., Riverton, N. J.
Curtis, J. A., Lynchburg, Va.
D'Agostini, M. M., Rome, Italy
Das, S., Albuquerque, N. M.
Davidson, M. J., Palisades, N. Y.
Davis, S. W., North Vancouver, B. C., Canada
DeGreene, K. B., Santa Monica, Calif.
DeLoriol, F., Bern, Switzerland
DeVito, M. J., Freehold, N. J.
Dodd, J. A., Jr., Haddonfield, N. J.
Dodd, J. C., Westmont, N. J.
Dolin, J. D., San Francisco, Calif.
Doyle, E. F., Mount Arlington, N. J.
Easterwood, M. J., Beaumont, Tex.
Ebner, L. J., San Diego, Calif.
Eden, W. D., Los Alamitos, Calif.
Ellis, H., Jr., San Diego, Calif.
Evans, W. M., Bristol, England
Finan, R. K., Fords, N. J.
Fioretti, E. E., Chicago, Ill.

(Continued on page 122A)

MICRO-MESH SIEVES DETERMINE PARTICLE SIZE DISTRIBUTION-Accurately

Set of Buckbee Mears Co. Micro-mesh sieves, standard sizes from 20 to 90 microns. Smaller sizes to order. Background: Enlarged evaporation mask for Mesa transistor.

Buckbee Mears research in subminiature mesh has solved many problems for the industry. A few current items are—

EVAPORATION MASKS—Mesa Transistors

SHADOW MASKS—Color TV

STORAGE MESH—Radar Tubes

CALIBRATED DIALS—Test Equipment

Anything that can be drawn can be reproduced. Send us your problems

SETS FROM 20 TO 90 MICRONS

The accurate measurement of particle size distribution of cracking catalyst has become practical. Buckbee Mears has developed screen sets which have been thoroughly tested as reported on by Ketjen N. V. Amsterdam*, and ASTME*. One comment "micro-mesh sieve method produces data with great accuracy and reproducibility." Another "the sieves made by Buckbee Mears Company appear especially attractive."

According to these papers, previous methods had disadvantages and results by different methods varied considerably. Now BMC sieves are recommended as a primary standard for testing size distribution of dry materials (cracking catalyst).

**Copies of these papers will be sent on request also complete sieve information.*

buckbee mears

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Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

DEVELOPMENT AND DESIGN ENGINEER

Development and design engineer, for advanced work in electronically regulated power supplies. Must have thorough working knowledge of solid state circuits and feedback systems. Salary to $16,000 according to qualifications and experience. Position offers a career opportunity in a rapidly expanding organization. Apply to Chief Engineer, Kepco, Inc., Flushing 52, N.Y.

COMMUNICATIONS FOREMAN

U.S. Oil Company, operation Middle East, seeks foreman with FCC or International Radio Telephone and Radio Telegraph First Class to supervise complete communications system with diversified equipment (AM, FM, HF, VHF, Teletype Printer, 100 Line PABX, Multichanneling, etc.). Candidate must have thorough background Electronic Theory with ten years practical experience (2 to 3 years Supervisory Experience). Approximately $11,600/year which includes $204 month Field Living Allowance; Bachelor Status; 30 day annual vacation with transportation paid; Savings Plan. No U.S. or Foreign Income Tax. Send Resume to Box 2074.

SENIOR PROJECT ENGINEERS

Immediate openings available for electrical engineers with solid state circuit design experience. Must be capable of assuming project administrative responsibilities. Will handle technical liaison activities with outside contractors. Be capable of directing in-house design effort, component evaluation, and subsystems tests. Masters Degree in Physics or Electrical Engineering desirable, but will accept Bachelors Degree with record of solid project achievement. An Equal Opportunity Employer. Send complete resume to Jet Propulsion Laboratory, 4804 Oak Grove Drive, Pasadena, California.

TRANSFORMER DESIGN ENGINEER

Transformer design engineer for electronic power supply components and to assist in pioneering development of magnetic type power control devices. Applicants must have a thorough knowledge of magnetic materials and component design. Salary to $16,000 according to qualification and experience. Position offers a career opportunity in an expanding organization. Apply to Chief Engineer, Kepco, Inc., Flushing 52, New York.

ELECTRONIC ENGINEER

A unique position is available for an Electronic Engineer to work in productizing advanced chemical and physical laboratory instrumentation in the field of nuclear magnetic resonance spectroscopy. Successful candidate should have a background including at least 5 years experience in electronic circuit design. Position offers outstanding growth potential and individual should have project engineer capabilities. Outstanding benefit program includes Cash Profit Sharing and Stock Purchase Plans. Submit resume in confidence to: Varian Associates, 611 Hansen Way, Palo Alto, California.

(Continued on page 92A)

RESEARCH ENGINEERS; GO FUNDAMENTAL— GO ALLISON!

● Engineers looking for opportunities to convert laboratory experiments to initial stages of development will find exciting new challenges at Allison.

Current contracts have generated long-term programs in new energy conversion projects such as Military Compact Reactor (MCR), Liquid Metal Regenerative Fuel Cell, Boundary Layer Control (BLC), and others.

We have immediate requirements for men with experience and advanced degrees:

MECHANICAL ENGINEERS—Ph.D. or Sc.D. mechanical or electrical engineering, 5 years experience in solid state devices. Must be capable of directing design and construction of prototype models of solid state energy conversion concepts previously worked out in research laboratory.

Also, M.S. with 5 years actual experience in design field, including working knowledge in processing, fabrication, materials, and metal joining techniques. Must have good understanding of fluid mechanics, thermodynamics, heat transfer and mechanics of materials.

INSTRUMENTATION SPECIALISTS—M.S. or equivalent with background in thermodynamics, fluid mechanics, electronics and systems control. Requires working experience in electronics and controls associated with (1)—energy conversion devices operating in the plasma temperature ranges and utilizing electrical discharge methods and or (2)—nuclear power systems.

HEAT TRANSFER SPECIALISTS—Ph.D. in ChE or ME, or equivalent. Requires a high level of analytical ability and broad experience in research, design, and application of liquid metals as heat transfer fluids. Will be responsible for major research programs involving liquid metals.

SYSTEMS ENGINEERS—B.S. or M.S. Requires broad background in areas related to design, test, installation and operation of nuclear power systems. Duties include planning future requirements, applications, and programs for these systems.

DIRECT CONVERSION DEVICE SPECIALISTS—Sc.D. or equivalent in ChE. Background in electrochemical engineering, liquid metal technology or related fields. For development of liquid metal cell devices and handling of liquid metals in the environment of an electrochemical conversion system.

Also, ME with some experience in design and fabrication of thermionic converters or related devices (microwave tubes, small ceramic-metal tubes, etc.). Requires knowledge of thermionic materials, refractory metals, high temperature ceramics, thermionic emitters, and mechanical design problems associated with high temperature electron devices.

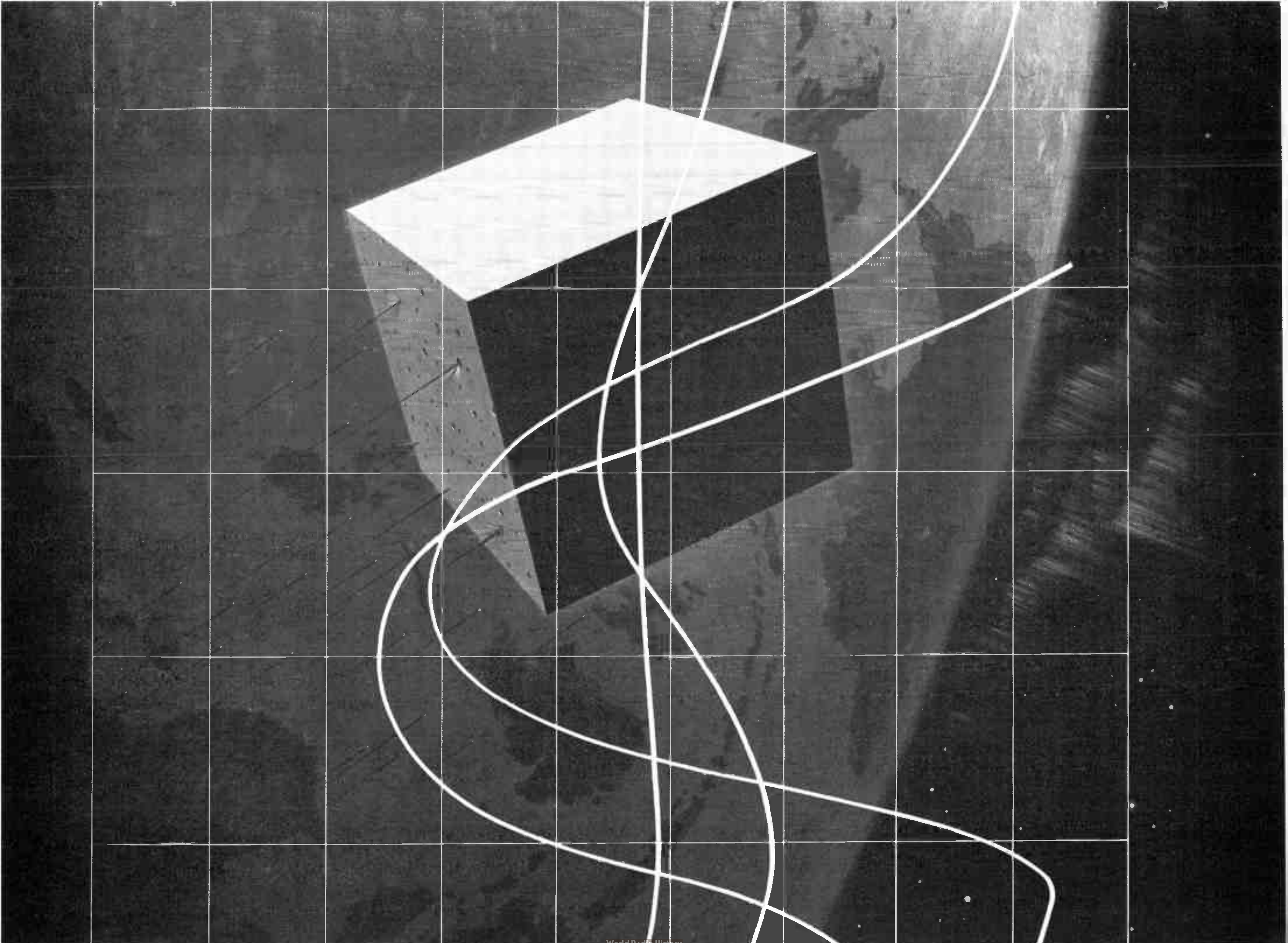
Other challenging openings exist for scientists and engineers interested in advancing the various forms of energy conversion systems.

Openings available NOW. A promising future for those who qualify is available in the creative environment at Allison—plus all the opportunities and advantages offered through an organization with the character of General Motors. Interested? Let us hear from you. Send your resume or write to: Mr. V. A. Rhodes, Professional and Scientific Placement, Dept. 611, Allison Division, General Motors Corporation, Indianapolis 6, Indiana.

An equal opportunity employer

Energy Conversion *Is Our Business*

ALLISON DIVISION, GENERAL MOTORS CORPORATION



Positions Open

(Continued from page 88A)

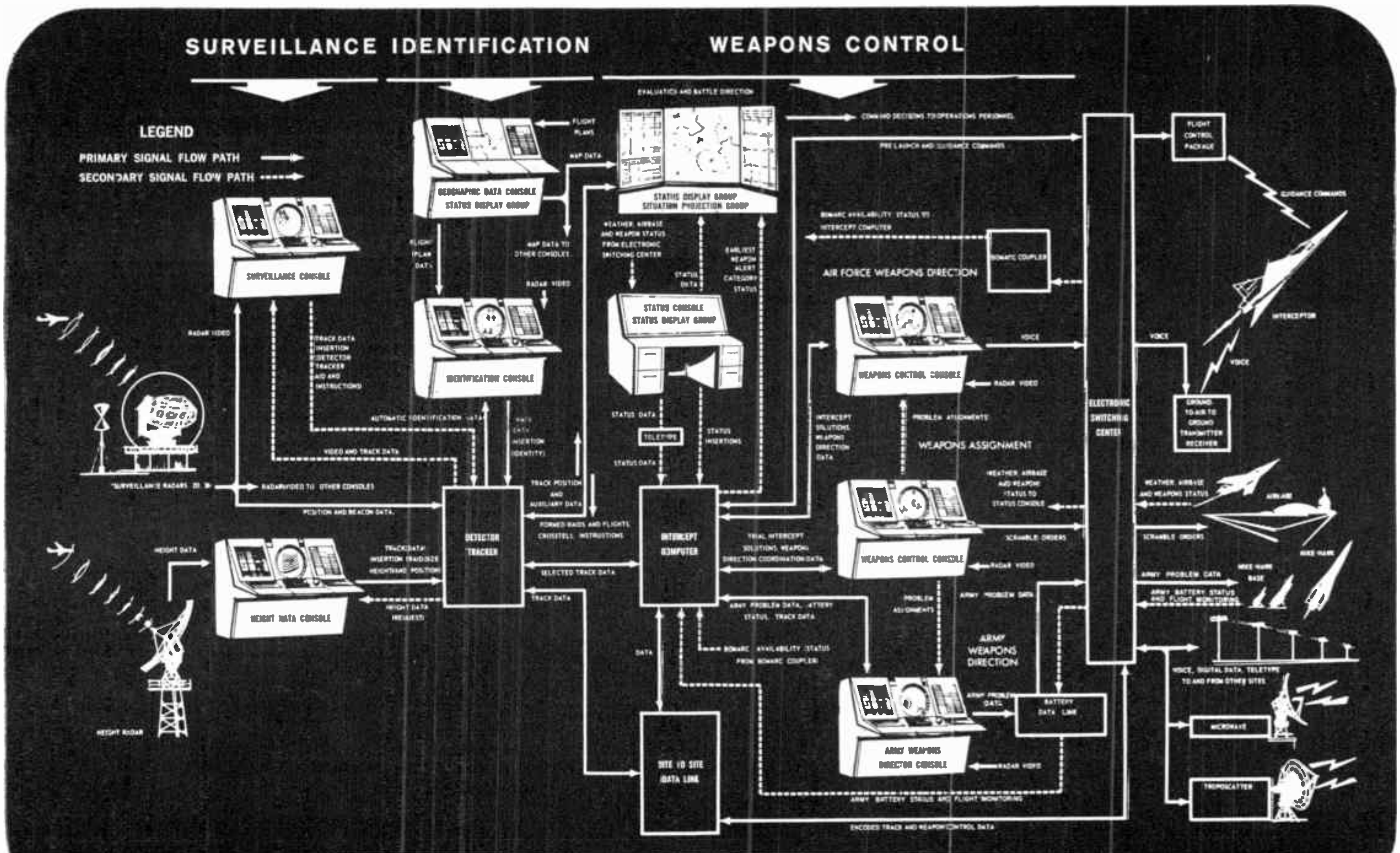
SENIOR ELECTRONIC ENGINEER

Excellent position involves working in the experimental and initial development phases of advanced types of commercial scientific instrumentation. Must be willing to work with unusual circuitry and components which often require extending the state of the art. Some background in tube and transistor circuits, radio frequencies, microwave, stabilized amplifiers and other servoproblems is desirable. Work will be in close contact with both physics and chemistry. A BS or MS in EE or Physics plus several years experience. Outstanding benefit program includes cash profit sharing and stock purchase plans. Submit resume in confidence to: Varian Associates, 611 Hansen Way, Palo Alto, California.

INSTRUMENTAL SCIENCES FACULTY POSITION

All graduate instruction and research. Young Ph.D. Physicist or Electrical Engineer with strong interest in the multidiscipline field or instrumentation and control. Develop courses and research program for M.S. degree and for expansion into Ph.D. program. Twelve month employment. Reply with resume to: Head, Electronics and Instrumentation, University of Arkansas, Graduate Institute of Technology, P.O. Box 3017, Little Rock, Arkansas.

(Continued on page 96A)



A big loop to close...

Your next job? It depends on what you're seeking ... a compatible increment in salary ... or a no-holds-barred chance to wrestle with problems of an entirely new magnitude. An example of the latter lies between the lines of the diagram above. ☐ Superficially, the schematic traces the data flow at a hypothetical Control and Reporting Center of the Air Weapons Control System 412-L. In the over-all mosaic of a fully integrated 412-L System, such a Control and Reporting Center is a key element. ☐ We are managing the integration and implementation of the 412-L program under the direction of the U.S.A.F. Electronic Systems Division, from system concept and development, through the installation and checkout of complete operating equipments. If you'd like more information about this program, and the calibre of opportunities it offers, send us an outline of your education and experience. Your personal reply will include a technical brochure describing the entire system. There's no obligation.

These are the general areas of immediate opportunity: Communication Systems / Applied Mathematics / Materials Engineering / Microminiature Electronic Packaging / Semiconductor Circuits Design / Computer Systems Applications / Electronic Liaison-Production Engineering / Microminiature Mechanical Design / Operations Analysis / Systems Equipment Analysis / Telecommunication Systems Design / Project Integration Engineering / Equipment Evaluation.

Address your inquiry in confidence to Mr. P. W. Christos, Div. 53-MI, Defense Systems Department, General Electric Company, Northern Lights Office Building, Syracuse, New York.

DSD DEFENSE SYSTEMS DEPARTMENT

A Department of the Defense Electronics Division

GENERAL ELECTRIC

Northern Lights Office Building — Syracuse, New York

An Equal Opportunity Employer

Positions Open

(Continued from page 92A)

SALESMAN

To sell semi-technical instrument in eastern Pennsylvania and Delaware for manufacturers' representative on straight commission (no draw, no expenses; we handle the product on that basis). Degree not needed. Age is immaterial so long as you possess health and integrity. Write Box 2075.

Ph.D. MICROWAVES AND/OR SOLID STATE

Excellent opportunity for Ph.D. with background in microwaves and/or solid state. Some teaching and some research required. Private consulting encouraged. Department has Ph.D. program. Good salary and rank arrangements will be made for exceptional man from either the industrial or academic fields. Address replies to Dr. Fred Schumann, Chairman, Electrical Engineering Department, Vanderbilt University, Nashville, Tennessee.

ASSOCIATE PROFESSOR ELECTRICAL ENGINEERING

Ph.D. degree required. Salary $7500 to $8000 per academic year. Opportunity to develop research activity in area of interest. Address replies to: Head, Electrical Engineering Department, South Dakota School of Mines and Technology, Rapid City, South Dakota.

(Continued on page 98A)

Positions Open

(Continued from page 96A)

Ph.D. ELECTRICAL ENGINEERING

Unusual opportunity in U.S. Civil Service to recent Ph.D. in Electrical Engineering to continue his research or teaching while administering Army-sponsored basic research projects at U.S. universities, colleges and institutes. Air conditioned offices located on the Duke University campus immediately adjacent to the Engineering College and Physics-Mathematics Departments. Associates on the Engineering Sciences Staff active in individual research projects at Duke. University campus very attractive in a progressive area of the South. Good family environment. Civil Service levels begin at GS-11 ($7560) to GS-14 ($12,210), depending on background qualifications. Write details of background or submit a Form 57 (obtainable at any post office) to J. J. Murray, Army Research Office (Durham), Box CM, Duke Station, Durham, North Carolina.

ELECTRONIC ENGINEERS

Electronic Engineers for permanent positions with Federal Communications Commission, Washington, D.C., GS-12, $8055 to GS-13 $10,635. Must be graduate engineers with good knowledge of problems involved in the area of space communications. Government will pay expenses for transportation of employee, his family, and household goods to Washington. Attractive fringe benefits include retirement, life and health insurance, automatic pay increases. Good opportunity to enter government career service. Reply to Personnel Officer, Federal Communications Commission, Washington 25, D.C.

LIFE SCIENCE ENGINEERS

The human factor as a design detail is as old as engineering; as a studied design factor it is relatively new. Engineers with training or experience in the biological or behavioral sciences who are interested in analysis and experimentation in the areas of terrain-following and other radar displays, photointerpretation, advance vehicle handling qualities and heuristic programming techniques are invited to reply to: F. P. Rentschler, Cornell Aeronautical Laboratory, Inc., Buffalo 21, New York.

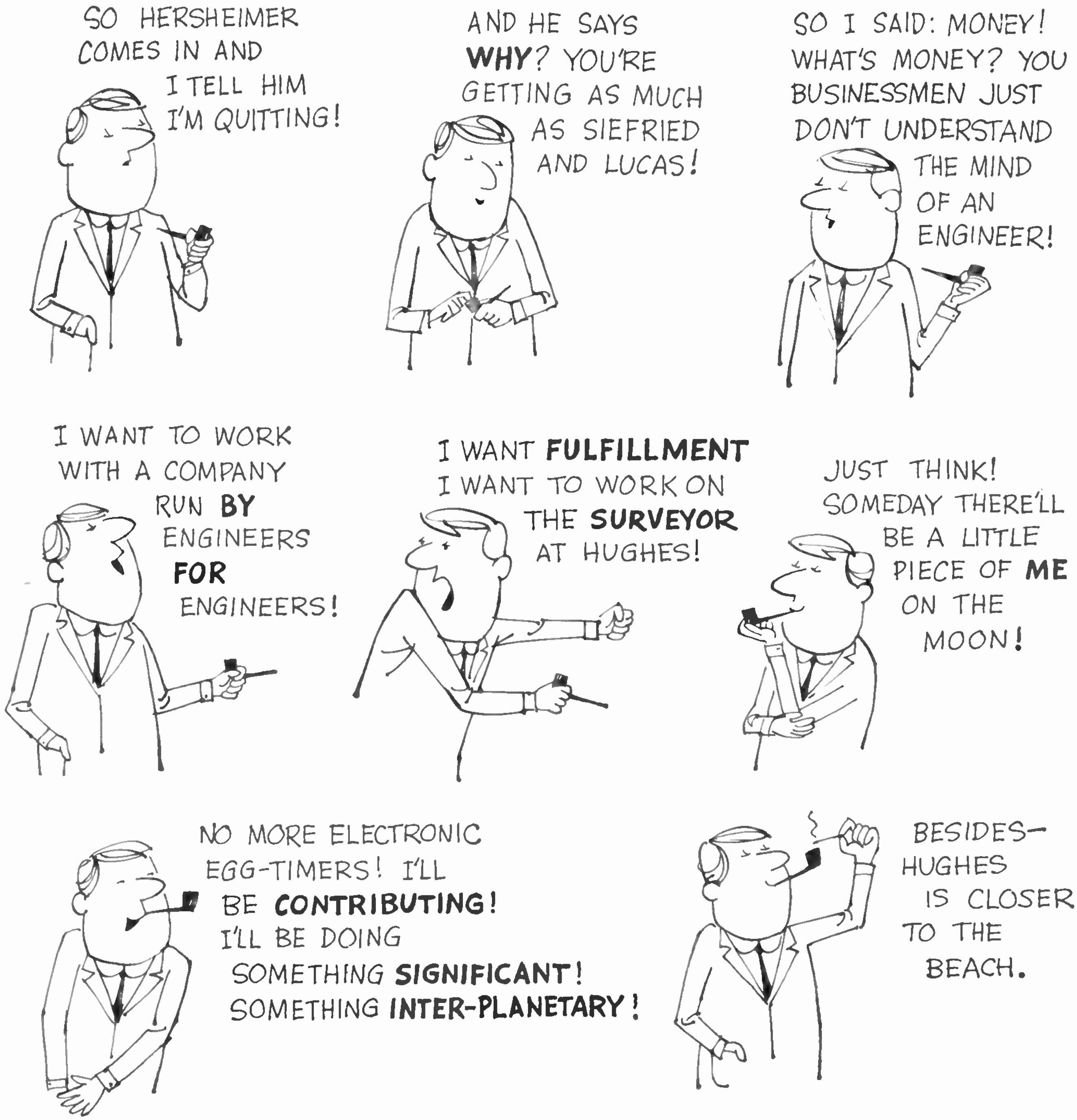
RESEARCH ASSOCIATE or ASSISTANT PROFESSOR

Immediate opening for professional-level applicant as Research Associate or Assistant Professor to supervise designing, fabricating, and maintaining specialized electronic equipment for research in science and engineering. A properly qualified person will, if he desires, have the opportunity to participate in electronic research or teach. A staff member on full-time appointment may enroll for one course each semester. Salary in range of $7500 to $8500. Annual appointment with one month vacation. Application or further inquiry should be made to: Dean Virgil W. Adkisson, Research Coordinator, University of Arkansas, Fayetteville, Arkansas.

ELECTRONIC ENGINEER

Electronic Engineer, B.S., recent graduate with a few years of practical experience in electronics or communications, for design and field engineering work. Projects will involve instrumentation for studies of the auroral ionosphere, radio-wave propagation, and geomagnetism. Salary in the $900-$1000 per month range. Write: Geophysical Institute, University of Alaska, College, Alaska.

(Continued on page 100A)



Hughes is hiring! Numerous opportunities now exist in a variety of advanced projects and studies. Examples include: The SURVEYOR—which will soft land an instrumented payload on the moon, SYNCOM—synchronous-orbit communications satellite, VATE—automatic test equipment for ballistic missiles, anti-ballistic missile defense systems—boost-intercept, mid-course and terminal, and many others. Positions are open at all levels for specialists with degrees from accredited universities.

Please airmail your resume to:

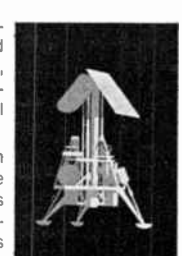
Mr. Robert A. Martin
Head of Employment
Hughes Aerospace Divisions
11940 W. Jefferson Blvd.
Culver City 37, California

CONTROLS ENGINEERS. Concerns airborne computers and other controls related areas for: missiles and space vehicles, satellites, radar tracking, control circuitry, control systems, control techniques, transistorized equalization networks and control servomechanisms.

CIRCUIT DESIGNERS. Involves analysis and synthesis of systems for: telemetering and command circuits for space vehicles, high efficiency power supplies for airborne and space electronic systems, space command, space television, guidance and control systems, and many others.

INFRARED SPECIALISTS. To perform systems analysis and preliminary design in infrared activities for satellite detection and identification, air-to-air missiles AICBM, infrared range measurement, air-to-air detection search sets, optical systems, detection cryogenics and others.

SYSTEMS ANALYSTS. To consider such basic problems as: requirements of manned space flight; automatic target recognition requirements for unmanned satellites or high speed strike reconnaissance systems; IR systems requirements for ballistic missile defense.



Creating a new world with Electronics

HUGHES

HUGHES AIRCRAFT COMPANY
AEROSPACE DIVISIONS

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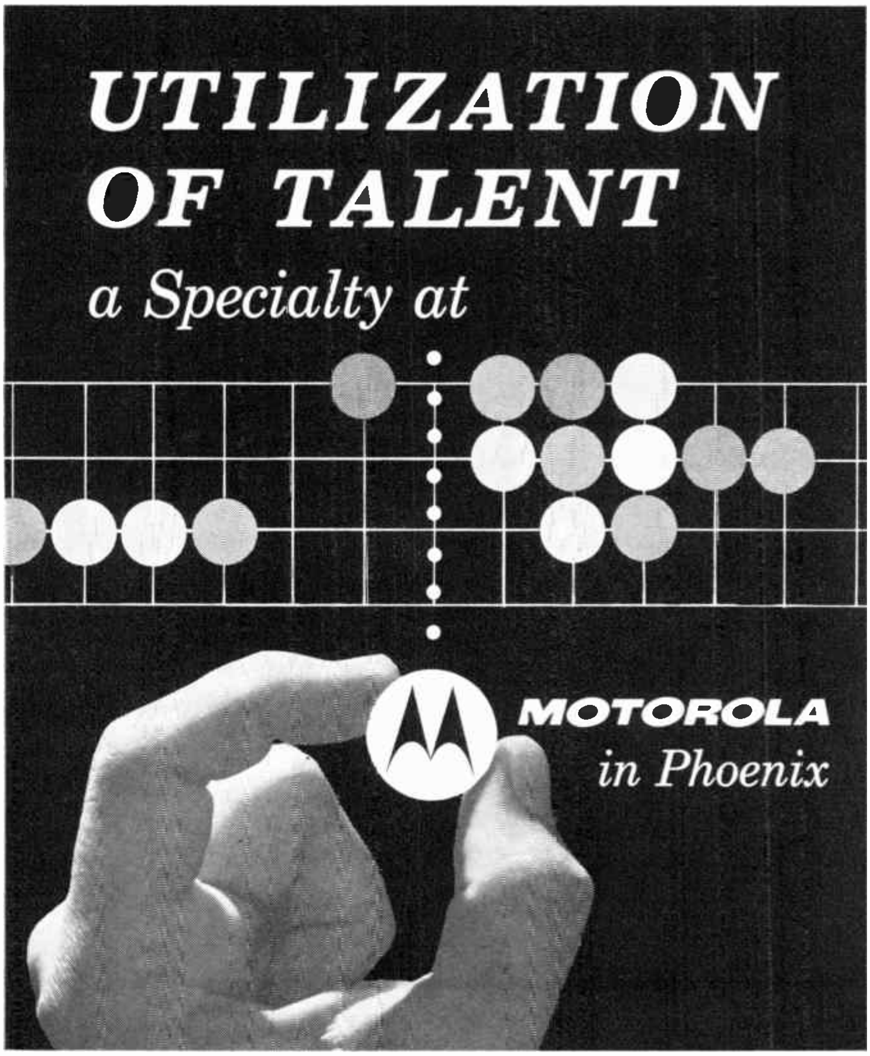

It takes more than knowledge, dedication, and imagination to achieve your career objectives. You've also got to pick the right company to work for — one that has a variety of state-of-the-art projects. One that's going places. One that knows how to make optimum use of good engineering abilities. Motorola is that kind of company — a pace-setter that excels in the selective placement of engineers and scientists. If *you're* looking for the opportunity to fully utilize your talents on high priority programs, write to Phil Nienstedt, Dept. 619.

SPECIFIC OPPORTUNITIES ARE:

Antennas and Propagation
Command and Control
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Integrated Circuitry
Reliability Analysis
Reliability Program Coordination
Parts Reliability
Data Acquisition, Processing and Display
Radar and Radar Transponders
Guidance and Navigation
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Telemetry
Instrumentation and Display
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Military Electronics Division
WESTERN CENTER · P.O. BOX 1417, SCOTTSDALE, ARIZONA

Motorola also offers opportunities at Chicago, Illinois, and at Culver City and Riverside, California

Positions Open

(Continued from page 98A)

ELECTRICAL DESIGN ENGINEERS

Experienced in video-amplifiers and pulse circuitry design. Familiarity with solid-state devices and transistorized circuitry would be desirable. Send resume to: George T. Rimbach, Supervisor of Personnel, HRB-Singer Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

MECHANICAL DESIGN ENGINEERS

Experienced in military electromechanical assemblies and high-speed rotational devices. Send resume to: George H. Rimbach, Supervisor of Personnel, HRB-Singer Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

RELIABILITY ENGINEERS

Knowledge of reliability theory and experience in military specifications, prediction techniques, and evaluation and testing methods. Send resume to: George H. Rimbach, Supervisor of Personnel, HRB-Singer Inc., Science Park, P.O. Box 60, State College, Pa. Qualified applicants will hold at least a four-year degree in science or engineering with three to five years related experience.

SUPERVISORY TECHNICAL PUBLICATIONS EDITOR

Incumbent will be the Assistant Technical Publications Officer for the Naval Aviation Engineering Service Unit. He will supervise the publication of the *Digest of U. S. Naval Aviation Electronics* and other technical publications. All manuscript copy will be submitted to him for review and editing. He will supervise a group of engineers engaged as writers and a clerical force and instruct a continuing changing group of Naval officers in the art of technical writing. GS-12—$8955 per annum. Send resume to: Industrial Relations Department, Naval Air Material Center, Philadelphia 12, Pa.

TEACHING POSITION

Ph.D. in Electrical Engineering needed for fall quarter. Undergraduate and master's programs. Nine months contract with summer employment readily available in area. Salary good and depends on rank and experience. New facilities. Electronics or communications fields preferable. Address replies to: Chairman, Department of Electrical Engineering, Seattle University, Seattle 22, Washington.

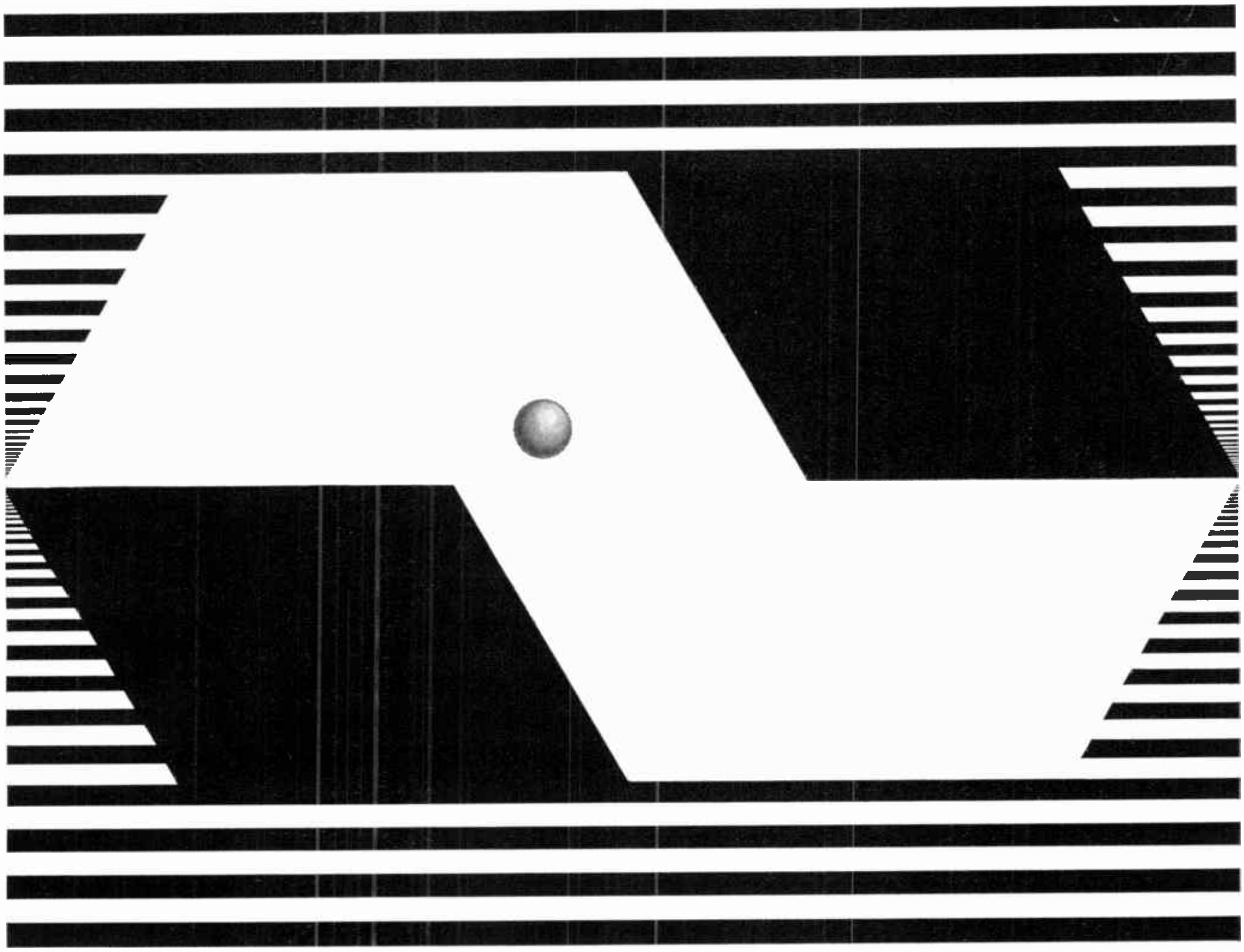
ELECTRONICS OR RADIO ENGINEER

Rapidly growing Manufacturing Company, employing 160, located near Philadelphia, requires individual with minimum of 3 years experience in microwave design and measurements to head research and development in the application of coaxial cable to high frequency equipment: B.Sc. in physics or E.E. minimum requirement, Masters desirable. Send resume and salary requirements to Box #2076.

CRITICAL VACANCIES AT GRIFFISS AIR FORCE BASE, NEW YORK

Electrical Engineer	GS-5	$5335 pa
Electronic Engineer	GS-5	$5335 pa

(Continued on page 104A)



Explorers in the shape of things to come

An idea in the mind of man...that's where every achievement in the world begins. Peer into the minds of Lockheed Scientists and Engineers. There you see ideas in the making—ideas that some day will take on form and substance. Not all, of course. Some are too "far out." But, no matter how visionary, all ideas win serious attention.

As a result, this freedom of imagination has led to many distinguished accomplishments at Lockheed. And the future holds still more. For, among Lockheed's ever-expanding programs are: Spacecraft; Satellites; Man-in-Space Studies; Hypersonic Manned Aircraft; Advanced Helicopter Design; Sophisticated ASW and Ocean Systems.

Scientists and Engineers who thrive in an atmosphere of freedom; whose creative processes flourish through exchange of ideas; who relish exploring the unexplored—to such men we say: Lockheed has a place for you. For example: In Human Factors; Electronics Research; Thermodynamics; Guidance and Control; Stress; Servosystems; Reliability; Dynamics; Manufacturing Engineering; Astrophysics; Astrodynamics; Advanced Systems Planning; RF Equipment Engineering; Bioastronautics and Space Medicine; Weapons Effects; Aerophysics; Digital Communications; Antennas and Propagation Engineering; Tracking, Telemetry and Command Engineering; Communications Analysis. Send résumé to: Mr. E. W. Des Lauriers, Manager Professional Placement Staff, Dept. 1809, 2402 N. Hollywood Way, Burbank, California. An equal opportunity employer.

LOCKHEED
CALIFORNIA COMPANY
A DIVISION OF LOCKHEED AIRCRAFT CORPORATION

Positions Open

(Continued from page 100A)

Electronic Engineer (Electro-Magnetics	GS-11	$8340 pa
Medical Officer (Occupational Health & Medicine)	GS-12	$8955 pa
*Radar Repairer	W-8	$2.41 ph
*Ground Radio Installer	W-8	$2.41 ph
*Radar Repairer	W-11	$2.66 ph

*—Positions require 75% travel along the Eastern Seaboard.

POTENTIAL CHIEF ENGINEER

Minimum 5 year experience, semi-conductor and vacuum tubes, circuit design, degree required, test equipment experience desirable. Opportunity to become Chief Engineer of a small growing test equipment manufacturer. Send resume of education and experience to: Dynatran Electronics Corp., 178 Herricks Rd., Mineola, Long Island, New York.

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.

FLORIDA SALES REPRESENTATIVE

An aggressive engineer with M B A and established O.E.M. contacts is available to sell your electronic instruments or components. Age 30, married. Ultimate goal is to establish a representative organization. Box 3994 W.

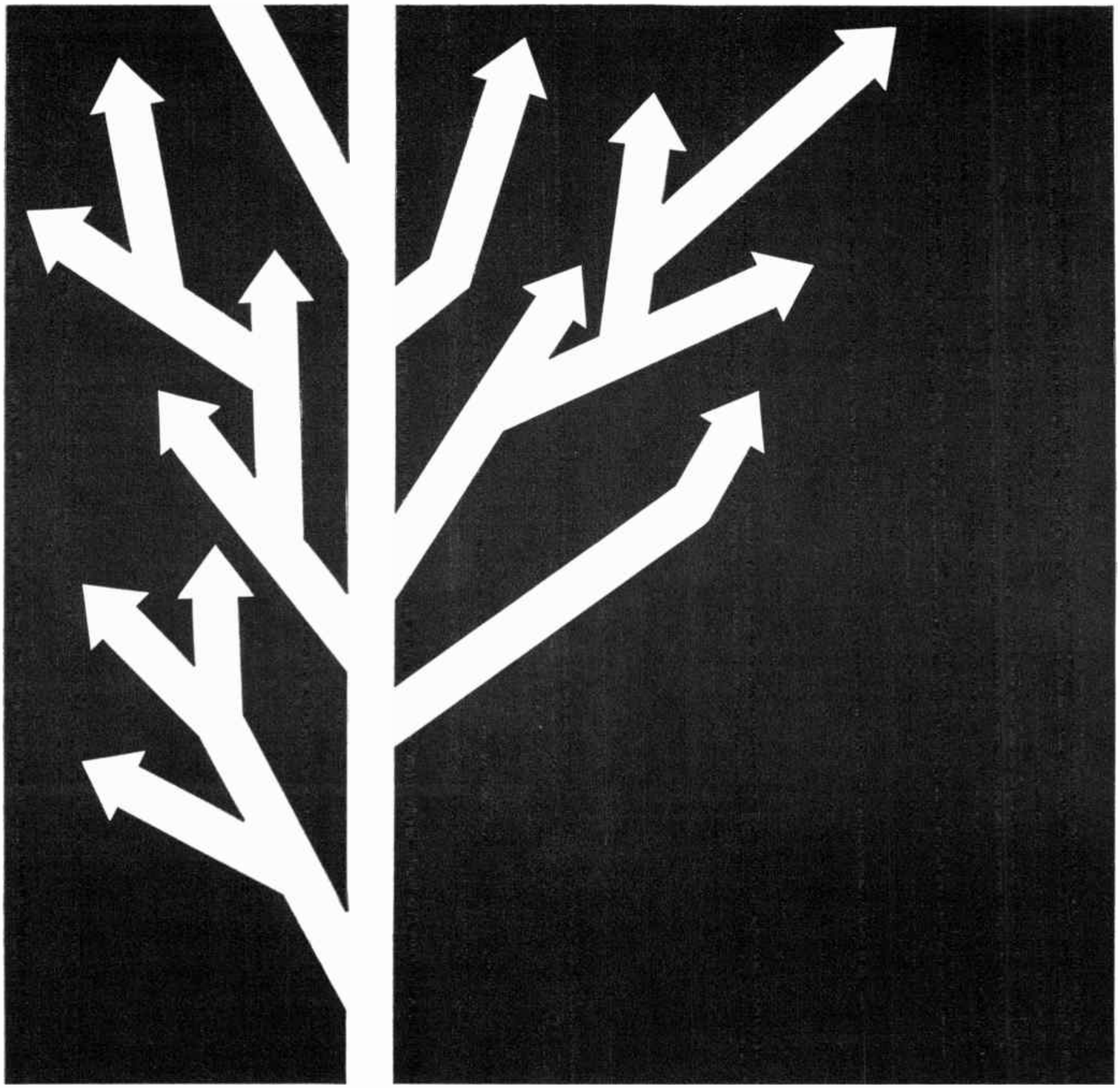
PUBLICATIONS ENGINEER, SENIOR

Experience: Supervisor, computer manuals group; 18 years experience in electronics and writing. Graduate of Capitol Radio Eng. Inst., B.A. in English at Univ. of Md. 30 sem. units graduate study in English plus misc. training courses in electronics. Box 3995 W.

MICROMINIATURE COMPONENT ENGINEER

Graduate engineer, age 35, with ten years' experience in design, development and production of semiconductors, resistors, capacitors and microminiature circuits, seeks responsible position with manufacturer or user. Experience includes semiconductor device fabrication, thin films, photolithography, ceramic and glass-to-metal seals, ultramicrominiaturization techniques. Box 3996 W.

(Continued on page 107A)



NOWHERE TO GO... BUT UP. Growth is the parallel story of Philco Western Development Laboratories and the career engineers and scientists who have joined WDL and have played roles in communications and control systems development for major U. S. space age projects. From a small nucleus, Philco WDL has grown to a staff of more than 3000, and the sprawling Palo Alto WDL complex is continually being expanded with modern laboratories to accommodate the growing staff. Growth means that there's work to be done. Growth means a record of achievement. To you, at Philco WDL, growth means a real career, opportunity for advancement, opportunity for achievement at the frontier of the nation's space age effort.

Write in confidence for information on how you can find your career at Philco WDL, with the additional rewards of ideal living on the San Francisco Peninsula and professional and monetary advancement commensurate with your own ability. Requirements include B.S. or advanced degree (electronics, mathematics, physics) and U. S. Citizenship or currently transferable D.O.D. clearance. Address Mr. Patrick Manning, Department R-9.

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Positions Wanted

By Armed Forces Veterans

(Continued from page 104A)

ELECTRONICS TECHNICIAN

Retiring from military service April 1963 with twenty years of communications-electronics experience. Graduate civilian and military schools. Resume upon request. Box 3997 W.

COMMUNICATIONS ENGINEER

Currently recalled as Captain in U.S. Army Security Agency. Industrial R & D experience in radio DF, communications, satellites, and military communications systems. B.S. and M.S. in EE; PE; Ph.D. course work; languages; age 31; highest clearances. Desires long term assignment in Europe in communication systems engineering or management. Box 3998 W.

ELECTRICAL ENGINEER

Elect. Engr. Tech; RCA T3 Grad; Married; 28; Evening student at CCNY. 2½ yrs. with switching systems engineering dept. of R & D lab. Desires challenging position with room for advancement in N.Y.C. Box 3999 W.

MAINTAINABILITY ENGINEERING

BSEE: MS Experimental Psychology; 8 years experience as electronics technician in military and civilian electronics desires position in maintainability engineering. Anywhere in the world except S. California, U.S. Citizen, formerly held Crypto clearance. Married, available approximately September 10, 1962. Write Box 4008 W.

(Continued on page 108A)

Positions Wanted

By Armed Forces Veterans

(Continued from page 107A)

PATENT/ENGINEER

Examiner, Patent Office, Electro-Mech. Art. Some experience in E.E., Veteran, Navy Electronics, B.S. Physics. Desires position in metropolitan New York City. Write Box 4009 W.

AUDIO EQUIPMENT DESIGN

BSEE 1958 (Electronics Option), Georgia Tech., age 26, single. Have been on duty in the Air Force since January 6, 1959. Desire ultimate employment in audio equipment design, research & development. Available in June or July 1962. Write Box 4010 W.

ELECTRICAL ENGINEER

BE & MS in EE. Registered PE, California. 5 years supervisory experience in Navy (LT USNR), including 2 years shipboard and 3 years instructor duty in U.S. Naval Academy in EE. 2 years associated with conduct and administration of industrial technical training on advanced equipment for major firm. Family. Fluent in Spanish. U.S. Citizen. Desire position in engineering in Central America because of family ties. Write Box 4014 W.

ELECTRICAL ENGINEER

USNR Officer, 25, BS '59. Three years supervisory and instructor experience in digital computation training program including computer mathematics, programming, transistor and logic circuitry. Previous industrial electronic experience. Desire opportunity in systems or computer design. Available September 1962. Write Box 4015 W.

EASTERN SALES REPRESENTATIVE

Will provide your company with profitable sales coverage. Base operations in New York City. Near-future expansion to three man organization. Comprehensive engineering and marketing background in systems, components and capabilities: electronics optics, electro-optics and opto-mechanics: military and commercial. Write Box 4016 W.

Section Meetings

AKRON

"Proposed Unification of IRE & AIEE," A. Bereskin, University of Cincinnati; Student Paper Competition; Joint with AIEE; 3/20/62

"Information Storage Density of Magnetic Recording & Other Systems," M. Camros, Armor Research Foundation; Joint with Cleveland IRE Section; 4/12/62.

"Terminal Air Traffic," R. Meuleman, AVCO; Joint with PGANE; 5/15/62.

Tour of Akron-Canton Airport Facilities; 6/19/62.

ALAMOGORDO-HOLLOMAN

"SYNCOM" (Synchronous Communications Satellite), S. Klien, Hughes Aircraft Co.; 6/19/62.

ALBUQUERQUE-LOS ALAMOS

Annual Picnic; 6/16/62.

(Continued on page 110A)

Section Meetings

(*Continued from page 108A*)

ATLANTA

"Communications, Present and Future," J. W. Travis, Southern Bell Telephone Co.; Election of officers; 6/22/62.

BAY OF QUINTE

"Flying Saucers," W. L. Smith, Dept. of Transport; 2/21/62.

"The RCAF Intergrated Data Processing System for Inventory Control," G. C. Mansell, Data Processing Air Material Command; 3/21/62.

"Nits, Bits, and Dits," R. N. E. Haughton, Bell Telphone Co.; Report on the 1962 International IRE Convention; K. V. Burkett; 4/19/62.

Annual Dinner; Election of officers; 5/23/62.

BINGHAMTON

"A Method of Defining PPI Resolution," D. W. Deno, G. E. Co.; "An Electrical-Optical Analogy" F. Saltz, IBM; "Predicing Transistor Turn-on Delay Time in the Common Emitter Configuration," R. J. Wilfinger, IBM; "Channel Capacity of a Realizable Passive Two-Port," T. B. Horgan, IBM; 6/18/62.

CEDAR RAPIDS

"The History of Avionics," R. Bruland, Collins Radio Co.; 5/16/62.

Annual Picnic; 6/16/62.

CENTRAL PENNSYLVANIA

"Radiation Experiments with the Tiros Weather Satellite," W. Nordberg, NASA, Garddard Space Flight Center; Joint Banquet Meeting with AIEE; 4/16/62.

"The IRE-AIEE Merger," A. B. Bereskin, Univ. of Cincinnati; Joint with AIEE and IRE Williamsport and Emporium Sections, 5/8/62.

CHINA LAKE

"Electronics in Everyday Life," R. G. S. Sewell, U. S. Naval Ordnance Test Station; 6/21/62.

CONNECTICUT

"Interaction of Light Waves in Non-Linear Dielectrics," N. Bloembergen, Harvard University; 4/19/62.

Discussion on "Engineering Education," W. A. Shaw, So. New Eng. Tel. Co.; W. P. Berggren, College of Engrg. Univ. of Bridgeport; W. R. Bush, Hamilton Std. Div.; F. Zweig, Yale School of Engrg.; R. E. Evans, Remmington Arms Co.; A. B. Bronwell, Univ. of Conn.; Joint with Fairfield County Subsection; 5/17/62.

"A Family Plan for Radiation Survival," P. P. F. Bujalski & H. H. Gatch, Jr., Windsor Locks & Self, Mystic; Election of officers; Joint with PGNS; 6/7/62.

DENVER

"New Trends in Electrical Energy Generation," L. F. Epstein, G. E. Co.; 3/9/62.

"Professionalism in Engineering," L. M. Robertson, Public Service Company of Colo.; 4/20/62.

"Teaching Machines," R. deKeiffer, Univ. of Colorado; Election of officers; 5/25/62.

DETROIT

"What Every Engineer's Wife Should Know," W. Kock, Bendix Systems Div.; Election of officers; 6/5/62.

EL PASO

"Ultra-Reliable Potentiometers," K. Doty, Trimpot Corp.; 4/26/62.

(*Continued on page 112A*)

Section Meetings

(Continued from page 110A)

Tour of Precision Measurements Lab. at Biggs AFB; 5/31/62.
Election of officers; 6/28/62.

EMPORIUM

"The Proposed IRE/AIEE Merger," A. Bereskin, Univ. of Cincinnati; 5/15/62.
"Why We Must Go Into Space," W. R. Dorenberger, Bell Aerosystems Co.; 6/12/62.

EVANSVILLE-OWENSBORO

Annual Picnic; 6/10/62.

FLORIDA WEST COAST

"Future of Electronics in the Home," C. A. Sadlow, Westinghouse; 6/20/62

FORT HUACHUCA

"The Myth of Electronic Warfare," W. Bryant, TPED; Election of Officers; 6/26/62.

FORT WAYNE

"How to Succeed in Business in Spite of Being an Engineer," E. A. White, Bowmar Instrument Corp.; Election of officers; 6/7/62.

GENEVA

"The International Astronautical Federation," L. P. Shepherd, IAF; Films—USA Space Flight—Glenn,; "USSR Space Flight—Again to the Stars"; 5/29/62.
Films—"USA Space Flight," "Friendship 7"; "USSR Space Flight," "Gagarin Flight,"; 6/6/62.

HAWAII

Election of Officers; 6/23/62.

ISRAEL

Lecture Series—"Sensory Information and the Electrical Activity of the Brain," W. A. Rosenblith, MIT; 4/3-5-9/62.

LAS VEGAS

"Proposed IRE/AIEE Merger," D. Reynolds, Univ. of Washington; 6/25/62.

LITTLE ROCK

"Voices from the Sky," R. Cook, Southwestern Bell Telephone Co.; 5/28/62.

LUBBOCK

"Project Mercury," C. Boaz, Southwestern Bell Telephone Co.; 6/19/62.

MIAMI

"A Decade of Growth in Electronics," D. R. Hull, USN, retired; 4/19/62.
Student Papers Contest—"Photo Multipliers," R. Buckley; "Electronically Measured Water Ski Jumps," R. Couch; "Automatically Controlled Power Supplies," R. Craig; "Mirror Theory in Plasmas," S. Jacobson; "Cryogenics," J. Weber; Univ. of Miami; 5/23/62.
"Birth of a Section," P. H. Craig, Airpax Co.; "The AIEE and Unity," R. H. Stevens, Florida Power and Light Co.; "The Role of a Student in a Profession," F. B. Lucas, Univ. of Miami; "What is a Region," M. R. Briggs, Westinghouse Electric Corp.; "Electronics, Men and Machines," P. E. Haggerty, President of IRE; 6/16/62.

MONTREAL

"Reliability—The Prospect Before Us," I. Kirkpatrick, RCA Victor Co.; Election of Officers; 6/13/62.

(Continued on page 116A)

What opportunities does Honeywell-Aero offer you? Read the stories of these

6 engineers...and why they moved to Honeywell

Don Walker, Materials Engineering

(Formerly of California)

"I was first attracted to Honeywell by the top level importance of the work I'd be doing, along with the creative freedom I'd have. The potential challenge offered by Honeywell's broad programs in the space and missile field were of prime importance to me. After I arrived here I quickly became aware of other advantages, such as the number of lakes and variety of summer activities available."

Gordon Handberg, Aerospace Development

(Formerly of Fort Worth, Texas)

"The Honeywell-Aero Division knows where it's going and what it's doing. And the supervisors know what they're after. Yet there's a tremendous amount of freedom here. The supervisor tells the engineer what needs to be done but doesn't necessarily specify the mechanism for getting it done; that's left up to the engineer. Besides, the engineer has a knowledge of his project from beginning to end."

Don Drung, Marine Systems

(Recently from Ohio)

"The nature of my work is challenging—I'm working with top level people on the most important engineering projects that exist today. And I have a good deal of working freedom. There's also a sufficient volume and variety of work to mean security for the individual engineer. And in many fields Honeywell-Aero is leading the industry. Besides this, the locale offers good fishing and The Minnesota Twins."

A personal invitation from Mr. W. T. Noll, Vice President and General Manager of Honeywell's Aeronautical Division in Minneapolis

"The men you see on these pages are representative of the hundreds of scientists and engineers who are finding rewarding work and careers at Honeywell.

At Honeywell we believe in and provide these things: freedom to create; a highly personal atmosphere; time to pursue individual concepts; a chance to work on the nation's most important aerospace projects; pleasant, unregimented working conditions; company-supported educational opportunities.

If you would like to know what Honeywell can offer you, fill out and mail this inquiry card. It will be held in strictest confidence. You will be personally contacted by a member of our professional staff in your career field."

W. T. Noll

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ATTENTION: Mr. Roy Richardson

Dale Gunderson, Computer Systems Research

(Recently from New Jersey)

"The Aeronautical Division of Honeywell offers me a lot of opportunity for advancement in my field of interest . . . computers. Incidentally, Minneapolis is one of the leading computer centers in the country. Besides, we enjoy the advantages of this area: Uncrowded conditions (I live just 10 minutes from work in a really nice suburb), first rate schools, outdoor life."

Sheldon Sorensen, Data Proc. and Comp.

(Native Minnesotan)

"First, as a mathematician I felt Honeywell offered me challenging and financially rewarding work. Second, here the individual is important. Third, the area has all the cultural and social advantages of the largest cities in the country (big-league sports, symphony, art centers, top level schools, etc.) yet manages to maintain a warm, friendly atmosphere."

Norman Schamber, Evaluation Engineer

(Recently from Utah)

"I knew that at Honeywell I'd be working on challenging engineering problems. I also knew that if I could help solve them, this is the kind of company that would give me recognition and advancement. In addition, this is a fine section of the country to live in, especially from the standpoint of education facilities and the caliber and stability of the people."

Right now we have openings for scientists and engineers to work on projects like Apollo, Mercury, Gemini and X-15

Instrumentation and Displays Engineer • Circuit Engineer • Packaging Engineer • Microelectronics Engineer • Airborne Display & Control Engineers • Engineering Project Administrators • Semiconductor Applications Engineers • Reliability Engineers • Applied Mathematicians • Vehicle Flight Controls Systems Analysts • Flight Cont. Systems Engineer • Flight Ref. & Nav. Systems Dev. Engineers

If you are interested in knowing more about a professional career with Honeywell in Minneapolis, fill in and mail the inquiry card attached to this ad, or write Mr. Roy Richardson, Director of Technical Service, Minneapolis-Honeywell, Aeronautical Division, 2660 Ridgway Road, Minneapolis 40, Minn.

To explore professional opportunities in other Honeywell locations, coast-to-coast, send your application in confidence, to Mr. H. G. Eckstrom, Minneapolis-Honeywell, Minneapolis 8, Minnesota.

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EDUCATION:

DEGREE: BS YR ____ MS YR ____ PhD YR ____

MAJOR SUBJECT:
☐ Aeronautical Engineering ☐ Mechanical Engineering
☐ Electrical Engineering ☐ Mathematics
☐ Chemical Engineering ☐ Physics

OTHER ____

FOLD AND STAPLE

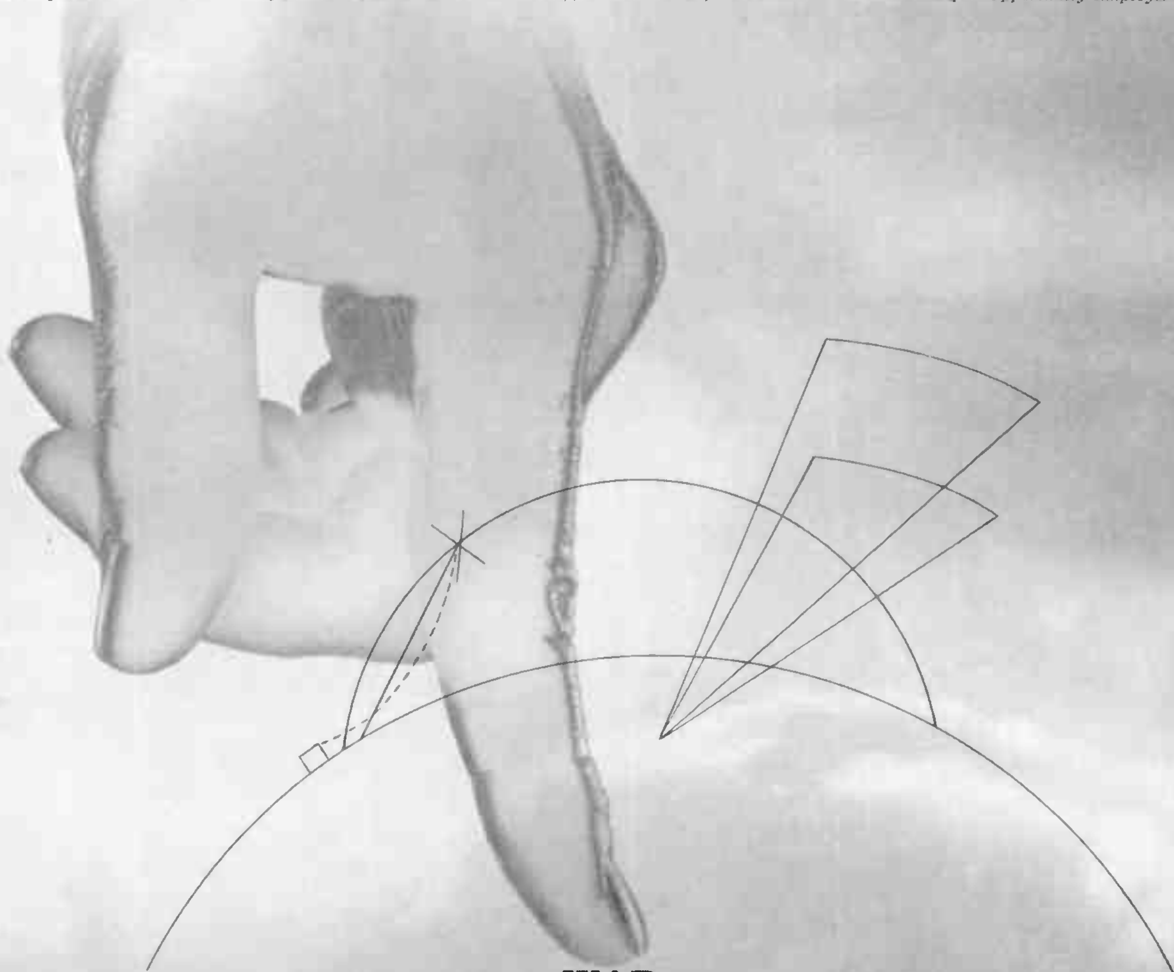
EXPERIENCE:

PRIMARY SPECIALTY ____ YRS ____

SECONDARY SPECIALTY ____ YRS ____

WHAT ARE YOUR CAREER AIMS OR GOALS?
(Comment, if you care to. It will help us place you in a position that will be personally rewarding to you.)

To mail this card, simply detach, fold and seal or staple. We pay the postage. Your background will be reviewed by a technical specialist, and your inquiry will be personally and privately acknowledged.



Section Meetings

(Continued from page 112A)

NORTHERN NEW JERSEY

Trip to IBM Research Center; 6/13/62.

NORTHWEST FLORIDA

"Naval Mine Warfare and Countermeasures Techniques," W. C. Bennett, USN; 5/24/62.

"Electronics in Archaeology," W. C. Lazarus, Eglin Air Force Base; 6/21/62.

ORLANDO

"The IRE," P. E. Haggerty, President of IRE; "How to Succeed as an Engineer," E. Fallon, The Martin Co.; 6/14/62.

PITTSBURGH

Panel Discussion—"Why Two Institutes?" A. B. Bereskin, Univ. of Cincinnati; W. F. Denkhaus, Bell Telephone Co.; A. A. Johnson, Westinghouse Electric Corp.; 4/4/62.

"New Method of Characterizing Nonlinear Devices," L. J. Giacoletto, Michigan State University; 5/7/62.

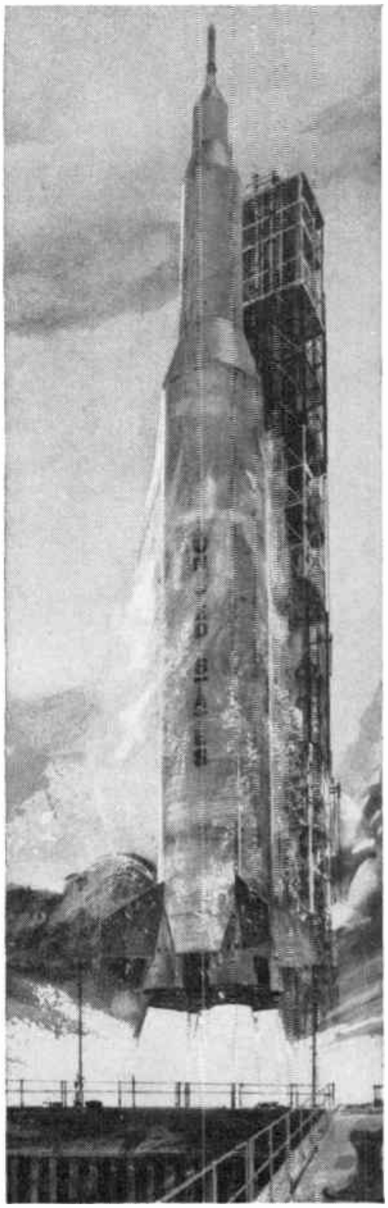
Boat Cruise along the Monongahela, Ohio & Allegheny Rivers; 6/16/62.

QUEBEC

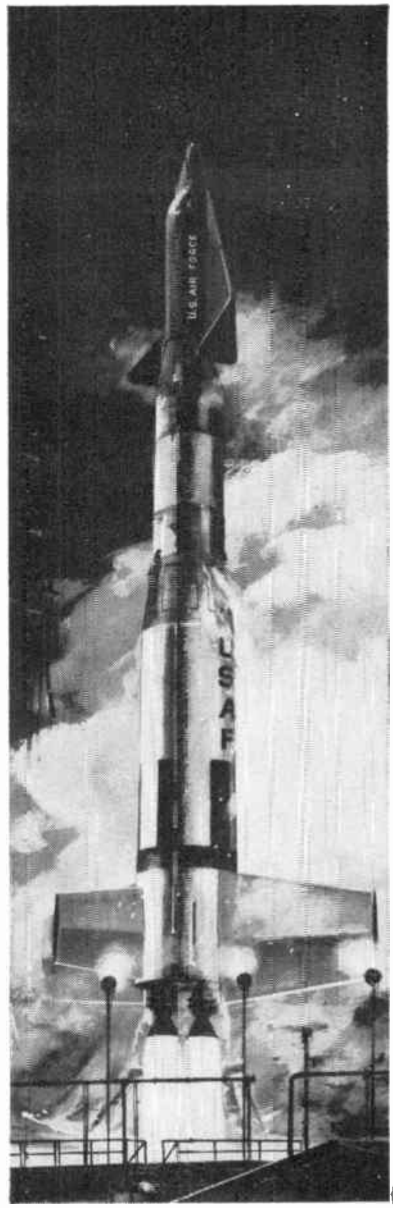
"Upper Atmospheric Composition Studies Using Microwaves," J. Beaulieu, Canadian Armament & Development Est.; Election of officers; 5/29/62.

"Structural and Thermal Design of the Topside Sounder Satellite," J. Mar, Defence Research Telecommunications Est.; 6/19/62.

(Continued on page 120A)



Saturn Launch Vehicle

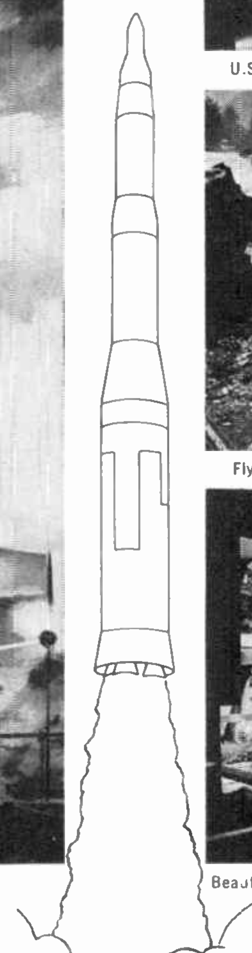


Dyna-Soar Space Glider

U.S. Science Pavilion at Seattle World's Fair

Fly-casting on Skykomish River near Seattle

Beautiful homes line Seattle's 200 miles of shore



Minuteman ICBM

Boeing openings for Electronic/Electrical Engineers

Boeing's Aero-Space Division, holder of major contracts on such advanced space programs as SATURN, DYNA-SOAR and MINUTEMAN, has a number of immediate, long-range openings offering professional challenge, and in many instances, unique ground-floor opportunities in newly-formed divisional organizations.

ELECTRONIC/ELECTRICAL engineering assignments are available in many areas, including the design of electronic equipment, the development of improved design techniques, electronic parts, electronic packaging techniques and the design and development of improved electronic systems.

Salaries are commensurate with all levels of education and experience. Minimum requirements are a B.S. degree in any applicable scientific discipline. Most of the assignments are in the uncongested Pacific Northwest, offering outstanding family living and recreational advantages.

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Section Meetings

(Continued from page 116A)

RIO DE JANEIRO

"The Participation of Brazil in the Experimental Program of Communications via Artificial Satellites," J. L. Emerick, RADIONAL; 4/11/62.

"Inadequacies of Existing Laws Covering Telecommunications & their Effects on the Country's Economy, Administration & Security," J. da Costa Vallim, Brazilian Air Force; 5/9/62.

SAN ANTONIO-AUSTIN

"LORAN C Long Range Navigation System," N. C. Dickerson, Collins Radio Co.; 4/26/62.

SAN FRANCISCO

Forum—IRE/AIEE Consolidation; 4/26/62.

Description of the pseudo-scientific arguments that have been employed to bolster or win the weaker side of a question, R. Weller, Lockheed Missiles & Space Co.; 5/11/62.

"Listening in on the Universe," C. L. Seeger, Stanford University; 6/12/62.

SCHENECTADY

"Airborne Particles," T. A. Rich, G. E. Co.; 6/12/62.

"Optical Masers," K. Tomiyasu, G. E. Co.; 5/8/62.

SHREVEPORT

"Air Force DATACOM (Phase I—COMIOGNET) System," F. W. Schultz, USAF; 5/15/62.

"Conquest in Communications," J. Z. Millar, Western Union Co.; 6/5/62.

SOUTH CAROLINA

Social Meeting; Joint with AIEE; 6/13/62.

TUCSON

"Lightning," W. Evans, Univ. of Arizona; 2/15/62.

"Solar Energy Laboratory," C. Hodges, Univ. of Arizona; 3/21/62.

"Spectrum Conservation and Transmitter Mutual Interference," M. Winkler, RCA; 4/26/62.

"Fundamental Systems for Producing Pulse Code Modulation," G. E. Clark, University of Ariz.; "A Voltage Variable Bridge Suitable for Automatic Gain Control," G. F. Ingle, Univ. of Arizona; 5/16/62.

VIRGINIA

Executive Committee Meeting; 6/8/62.

WESTERN MICHIGAN

"Transistors and Some of their Applications," L. J. Giacoletto, Mich. State Univ.; 4/11/62.

"FM Multiplexing," D. Wolters, Wood TV; 5/9/62.

Tour of WKZO-TV, F. Morse & Engineering Personnel; 6/13/62.

WASHINGTON

Debate on "Should the IRE & AIEE Merge?" W. Pike, R. Voight, S. Bailey, W. Swift; 5/7/62.

Social Evening; 6/15/62.

SUBSECTIONS

CAMBRIDGE

"Electronic Encephalography," V. H. Fischer, Battelle Memorial Inst.; Election of Officers; 6/26/62.

CATSKILL

Annual Dinner Dance; 6/23/62.

(Continued on page 142A)

Membership

(Continued from page 86A)

Fitzgerald, C. S., Arlington, Mass.
Fondeur, R. H., Levittown, N. J.
Fowler, A. M., Jr., Anaheim, Calif.
Fox, J. O., North Highlands, Calif.
Frank, R., Palo Alto, Calif.
Freeman, G. G., Compton, Calif.
Frost, P. A., Rockville, Conn.
Funk, R. R., Framingham, Mass.
Gallup, B. H., Rome, N. Y.
Garside, E. L., Blackwood, N. J.
Geckos, G. J., Chicago, Ill.
Gelinas, R. B., Redwood City, Calif.
Gianaris, G. N., Orlando, Fla.
Giordano, J. A., Glen Burnie, Md.
Green, M. A., Brunswick, Me.
Grieshaber, G. H., Camillus, N. Y.
Gross, J. L., Baltimore, Md.
Hahn, G. M., Monterey, Calif.
Haich, R. D., Rochester, N. Y.
Hall, M. E., Visalia, Calif.
Hall, R. T., Forest Hills, L. I., N. Y.
Harlan, W. T., Richardson, Tex.
Harris, D. L., Dallas, Tex.
Hastings, K. E., Rochester, Minn.
Haymaker, W. G., Jr., Granada Hills, Calif.
Henmueller, F., Chicago, Ill.

(Continued on page 124A)

Membership

(Continued from page 122A)

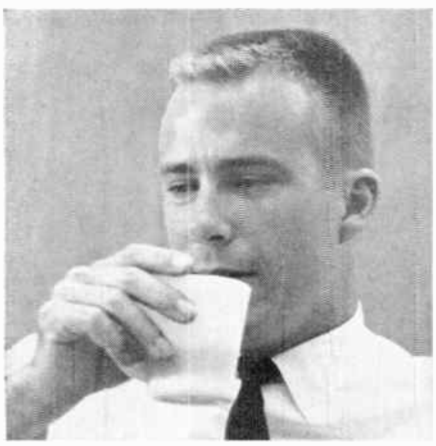
Herge, R. R., Hyde Park, N. Y.
Herman, L. M., Columbus, Ohio
Herrin, B. G., Santa Maria, Calif.
Hersey, M. P., Jr., Orinda, Calif.
Hill, J. W., Walton-on-Thames, Surrey, England
Hiller, J., Sydney, Australia
Hinshaw, M. L., APO, New York, N. Y.
Hirschman, M., Utica, N. Y.
Hornby, B. R. L., Walton-on-Thames, Surrey, England
Horowitz, D. J., Warminster, Pa.
Hunter, W. P., Oxnard, Calif.
Iden, C. C., Cambridge, Ohio
Ikard, W. L., Tulsa, Okla.
Isaksson, H., Kobenhavn, Valby, Danmark
Jacobs, H., East Northport, L. I., N. Y.
Jensen, L. V., Torrance, Calif.
Jester, C. E., Jackson, Miss.
Johnston, W. T., Whittier, Calif.
Judd, M. A., Loveland, Colo.
Judd, O. P., Los Angeles, Calif.
Kalos, S., West Hollywood, Fla.
Kazerman, S., Malden, Mass.
Kessen, R. F., Columbia, S. C.
Kessler, I., Rockville, Md.
Knuth, L. L., Seattle, Wash.
Kress, J. H., Eau Gallie, Fla.
Lampos, N. T., West Orange, N. J.
Latorre, V. R., Bellevue, Wash.
Lapanne, L. L., Orlando, Fla.
Lea, A. T., Bragknell, Berks, England
Leach, R. F., Lynnfield, Mass.
LeBlanc, E. A., Huntington, L. I., N. Y.

(Continued on page 126A)

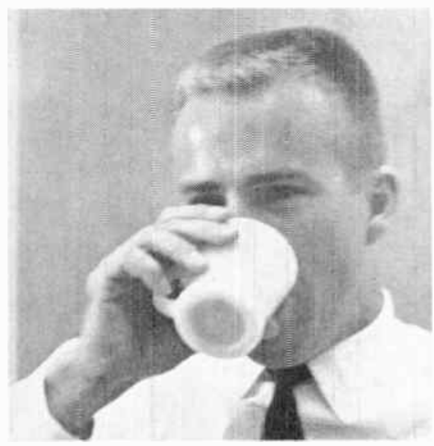
We've got 1980 on the drawing boards.

Solar probes. Landings on Mars. Round trip trajectories. That's my field. Working on trajectories of future spacecraft with advanced propulsion systems. Ion propulsion, for example.

The normal chemical rocket has a high initial thrust. Everything happens all at once, when it leaves the Earth. Then it coasts the rest of the way with a small maneuver to trim things up.

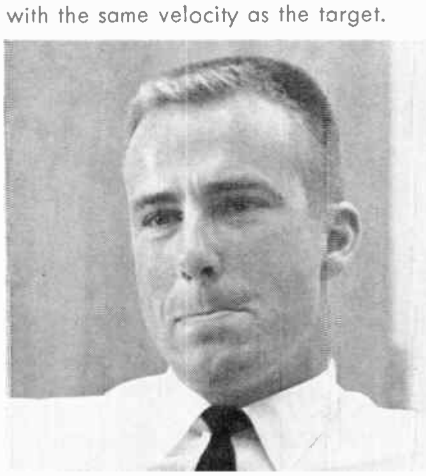


Advanced propulsion-type trajectories are more efficient but have low thrust, which causes the rocket to operate over most of the trajectory. A typical mission, for example, is one where the spacecraft approaches its destination target with the same velocity as the target.



We've been working on advanced propulsion trajectories, applying optimization theory to come out with the best estimate of payload capabilities.

We managed to come up with trajectory techniques that don't involve the design parameters of existing propulsion systems. It's all in a paper I gave at an ARS meeting in San Francisco last March.



From this information we've been compiling, you can find out what payload capability an interplanetary mission will have at any date from now to 1980, how much gas it's going to cost you, and how long it takes to hit target.

The people I work with are pretty hard to beat anywhere. As far as being in the midst of things, you'd have to go some to find a place like this. All the facilities you need. And I'm only a 15-minute bike-ride to work.

You've just been talking with Dr. William Melbourne of Jet Propulsion Laboratory. He's been at JPL for six years. And he's here to stay. Do you feel that way about where you are now? If not, why not write to JPL today.

"Research and development for America's lunar, planetary and interplanetary exploration programs."

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4804 Oak Grove Drive, Pasadena, California
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"An equal opportunity employer"

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(Continued from page 124A)

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Lee, J. R., Alexandria, Va.
Leifer, M., Bronx, N. Y.
Lemay, R. V., Cranston, R. I.
Leslie, E. B., Wanamassa, N. J.
Lewis, J. G., Lantana, Fla.
Lindfors, P. O., Edwards, Calif.
Liveley, S. P., Laval Des Rapides, Que., Canada
Lochinger, R. B., Cranbury, N. J.
Lodge, W. D., Brunswick, Me.
Lott, L. M., Bethany, Okla.
Luck, J. O., Murray Hill, N. J.
Ludford, J. F., Rome, N. Y.
Lutz, R. W., Williamstown, Mass.
MacCormick, A. J. A., Woomera, South Australia
Malagari, P. E., Windsor, N. Y.
Maravilla, M. M., Tokyo, Japan
Martin, W. L., Canoga Park, Calif.
Massa, J. A., Gardena, Calif.
May, C. B., Ridgecrest, Calif.
McAhron, M. C., Las Vegas, Nev.
McBain, R., Chicago, Ill.
McDonald, J. W., Arlington, Va.
McDowell, A. N., Amherst, N. H.
McIntyre, G. W., Kearns, Utah
McLeod, I. N., Akron, Ohio
Meeks, A. E., Merrick, L. I., N. Y.
Meli, J. J., Dayton, Ohio
Middlestead, R. W., Drexel Hill, Pa.
Millar, G., Newtownabbey, North Ireland
Miller, J. E., Baltimore, Md.
Miller, R. W., Canton, N. Y.
Mitchell, C. H., San Luis Obispo, Calif.
Mollen, J. C., Norwich, N. Y.
Moore, R. B., Washington, D. C.
Moran, J. K., North Babylon, L. I., N. Y.
Morton, K. C., Mobile, Ala.
Moxley, J. E., Pacoima, Calif.
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Mueller, K. M., Burlington, Mass.
Muller, F. A., Amsterdam, Netherlands
Murray, T. K., Jr., Millbrook, N. Y.
Neilson, R. A., Manhattan Beach, Calif.
Newson, H. R., Syracuse, N. Y.
Nonokuchi, K., Tokyo, Japan
Olson, F. J., Costa Mesa, Calif.
Ostinett, R. J., Fayetteville, N. Y.
Packard, W. B., Oklahoma City, Okla
Paine, R. B., Fair Haven, N. J.
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Pohlman, R. T., Bountiful, Utah
Popp, D. J., Chicago, Ill.
Rayan, A. A., Amman, Jordan
Raxter, N. E., Goleta, Calif.
Reynolds, W. P., Berkeley, Calif.
Ricketts, T., Dorchester, Mass.
Riddle, J. E., Woodland Hills, Calif.

(Continued on page 128A)

Membership

(Continued from page 126A)

Robison, P. E., Long Branch, N. J.
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Rote, J. C., Roseland, N. J.
Roveti, D. K., Washington, D. C.
Salgarkar, S. N., Bombay, India
Sasaki, I., Kawasaki, Japan
Satterfield, M. V. G., Palos Verdes Estates, Calif.
Scanlan, S. O., Croydon, Surrey, England
Scarbrough, H. H., Warrington, Fla.
Schleicher, C. L., Hicksville, L. I., N. Y.
Schwartz, M. B., Fair Lawn, N. J.
Seelhoff, H. L., Plymouth, Mich.
Shajenko, P., Warren, Mich.
Sharma, P. D., Panjab, India
Shipman, R. W., Detroit, Mich.
Shults, R. D., Atlanta, Ga.
Siegel, M. L., Scottsdale, Ariz.
Simons, E. H., Albuquerque, N. M.
Slatten, L. L., Manhattan Beach, Calif.
Smith, C. A., London, England
Soderholm, R. V., Alexandria, Va.
Stamler, A. D., Fairfax, Va.
Stirling, N. C., Playa del Rey, Calif.
Subbarao, M. K., Poona, India
Sudlow, W. R., Houston, Tex.
Thomas, G. L., Kokomo, Ind.
Thomas, W. N., Philadelphia, Pa.
Toche, C. A., Woodland Hills, Calif.
Vail, T. E., Pasadena, Calif.
Van Weelden, R. K., Evanston, Ill.
Vercillo, A. J., Chicago, Ill.
Waaben, S. G., Princeton, N. J.
Watson, T. A., Stamford, Conn.
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Weinstein, S., Woodbury, L. I., N. Y.
Wham, R. D., Pfafftown, N. C.
Whittall, W. J., Aurora, Ont., Canada
Williamson, W. J., Bayswater, Western Australia
Wojtasinski, R. J., Jupiter, Fla.
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Woodward, J. C., II, APO, San Francisco, Calif.
Wuerstl, J. D., Los Angeles, Calif.
Yamada, E. Y., Torrance, Calif.
Yerry, D. F., Maple Shade, N. J.
Yim, H. C., Gardena, Calif.
Yoshida, Y., Tokyo, Japan

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Asada, A., Osaka-shi, Japan
Atwood, C. B., Cincinnati, Ohio
Austin, H. M., Washington, D. C.
Beltran, S. F., Mexico, D. F., Mexico
Beresford, J. C., Montreal, Que., Canada
Beverstock, C. A., APO, New York, N. Y.
Bigelow, H. W., Washington, D. C.
Blanchard, F. H., Houston, Tex.
Brancati, T. A., Commack, L. I., N. Y.
Brown, D. S., Syracuse, N. Y.
Buss, A., St. Louis, Mo.
Capparelli, M. J., Jr., Utica, N. Y.
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Clark, R. C., Houston, Tex.
Derby, B. R., Kent, Ohio
Donahue, F. X., Revere, Mass.
Dorsey, J. E., Jr., Houston, Tex.
Dry, J. S., Jr., Orlando, Fla.
Durand, E. E., Jr., Pewee Valley, Ky.
Eiland, J. L., Houston, Tex.
Erhart, R. C., Valley Station, Ky.
Fong, T. M., Hong Kong
Forti, J. S., Washington, D. C.
Fox, M. A., Las Vegas, Nev.
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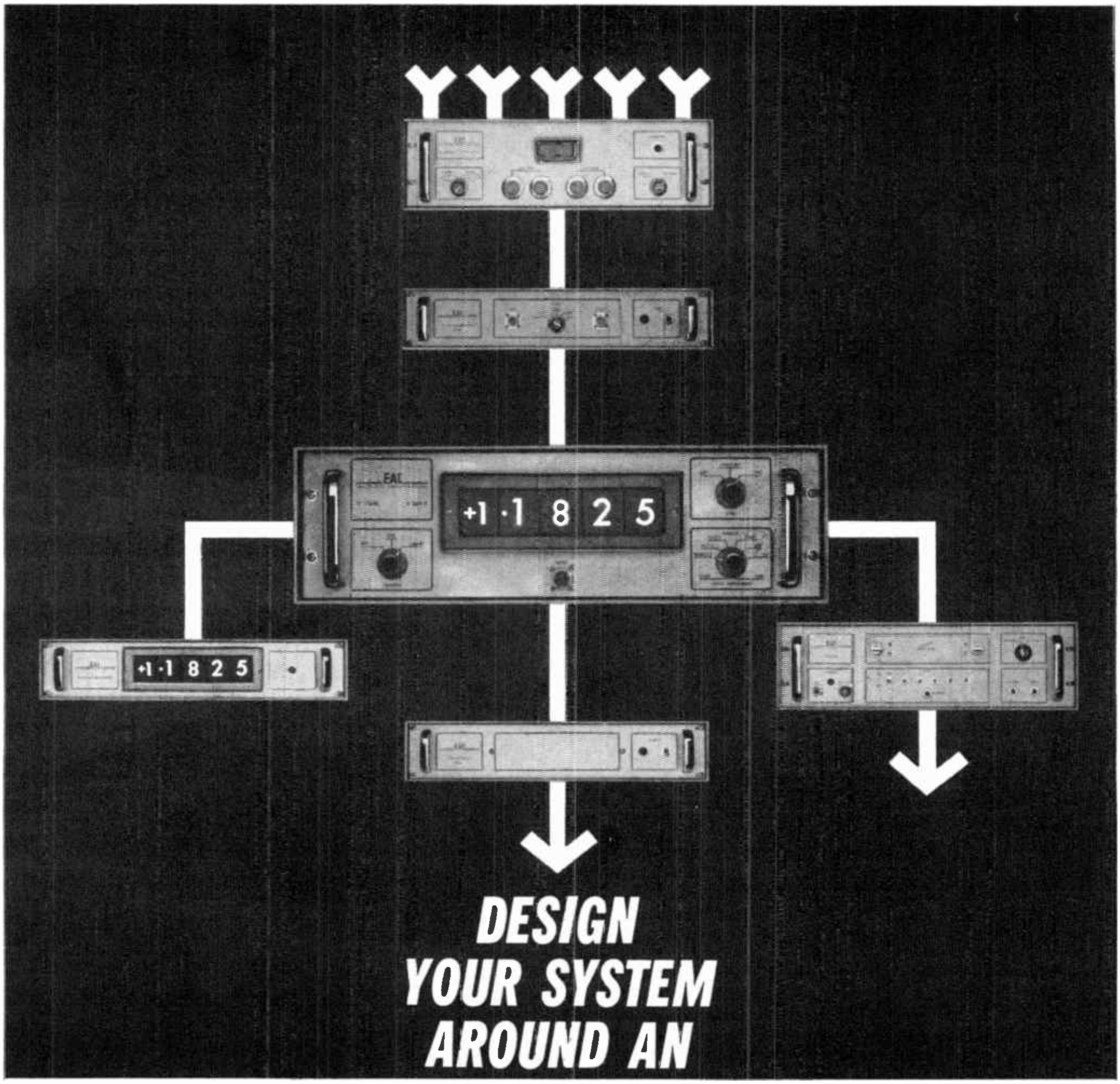
Goldfarb, M. S., South Euclid, Ohio
Graham, J. W., Lubbock, Tex.
Griffiths, J. E., Ilford, Essex, England
Grinnell, B. J., Owego, N. Y.
Haberman, H., Woodside, L. I., N. Y.
Hagan, N. A., September Iles, Que., Canada
Haxton, D. W., Goleta, Calif.
Hayes, E. J., Jr., Takoma Park, Md.
Haylock, D. G., San Mateo, Calif.
Holbrook, P. W., West Swanzey, N. H.
Horton, R. A., Bellaire, Tex.
Howard, F. G., Freeport, L. I., N. Y.
Howe, L. D., Jr., Apalachin, N. Y.
Jansen, J. C., Oswego, Ore.
Jesperson, C. P., Chicago, Ill.
Kaufmann, L. B., Houston, Tex.
Krampe, A., Danbury, Conn.
Lawres, N. I., Glen Cove, L. I., N. Y.
Lawshe, C. A., Jr., Gulfport, Miss.
Lawson, A. S., Toronto, Ont., Canada
Levoy, D. G., El Segundo, Calif.
Lippman, R. I., Philadelphia, Pa.
MacKeraghan, L. R., Chittenango, N. Y.
Mackey, A. C., FPO New York, N. Y
Mandakis, J. B., Winston-Salem, N. C.
Martynowski, R. F., Philadelphia, Pa.
McKinley, H. F., Houston, Tex.
Mehnert, R. C., APO New York, N. Y.
Merken, A. L., East Northport, L. I., N. Y.
Mill, C. B., Burlington, N. C.
Molnar, J., Wyckoff, N. J.
Neville, J. J., Honolulu, Hawaii
Olson, F. W., APO New York, N. Y.
O'Mathuna, P. G., Montreal, Que., Canada
Owens, A., Jr., Paramus, N. J.
Palmer, J. R., Sanford, Me.
Pipe, G. E. P., Los Altos, Calif.
Plate, P. J., Mount Kisco, N. Y.
Powell, C. J., Jr., Fort Bliss, Tex.
Ranshous, W. E., Conneaut, Ohio
Ray, E. F., Los Angeles, Calif.
Reisner, R., Wilmington, Calif.
Robinson, R. C., St. Petersburg, Fla.
Rose, J. R., Jr., Inglewood, Calif.
Rose, S. M., Akron, Ohio
Schwartz, P. B., Philadelphia, Pa.
Seigel, F. G., Brooklyn, N. Y.
Skolnick, B., Brooklyn, N. Y.
Skolnick, D., Chicago, Ill.
Sorensen, N. S., Provo, Utah
Spungin, N., Lexington, Mass.
Staggs, J. B., Bensenville, Ill.
Stemberg, G. D., Redwood, Calif.
Stowe, J. P., Newton, Conn.
Taylor, J. C., Don Mills, Ont., Canada
Tubbs, T. P., Patrick AFB, Fla.
Tucker, J. M. C., Indianapolis, Ind.
Tuger, J. A., North Palm Beach, Fla.
Turner, J. H., Burlington, N. C.
Viola, P. J., Houston, Tex.
Vleming, D. R., Poughkeepsie, N. Y.
Williams, F. D., APO New York, N. Y.
Wynn, E. H., Arnold, Md.
Zeik, J. E., Bridgeport, Ohio



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New instrumentation system capabilities are available to systems designers and laboratory engineers in EAI's fully transistorized digital voltmeters and system accessories. Solid-state reliability, six-month guaranteed stability, high speed and complete system provisions make EAI Series 5000 and 5001 DVM's ideal for the most critical applications. ☐ *EAI Series 5000 and 5001 Digital Voltmeters are completely solid-state with 0.01% plus 1 digit absolute accuracy, six months stability, 10 milliseconds maximum reading time, buffer storage with BCD and decimal outputs and full-time input impedance to 1000 megohms.* ☐ *Series 5500 AC-DC Converter is a completely solid-state operational amplifier type providing 0.05% accuracy with 150 milliseconds maximum settling time.* ☐ *Series 5700 Input Scanner utilizes highly reliable cross-bar switching for scanning up to 30 points per second.* ***EAI output devices include:*** ☐ *Series 5600 Remote Visual Readout* ☐ *Series 5610 Printer Control for compatibility with parallel printer* ☐ *Series 5010 Serializer for compatibility with serial recorder* ☐ *Series 5620 High Speed Printer for recording up to 220 lines per minute* ☐ *Complete information on EAI Series 5000 and 5001 transistorized digital voltmeters and system accessories are available in Bulletin IC 6077. Write for your copy today.*

EAI

ELECTRONIC ASSOCIATES, INC. *Long Branch, New Jersey*

NEWS New Products

(Continued from page 48A)

Audio Engineering Society Announces Fourteenth Annual Convention Program and Professional Products Exhibits

More than one hundred papers covering almost every aspect of sound and sound reproduction will be presented at the Fourteenth Annual Audio Engineering Society Convention to be held at the Hotel Barbizon-Plaza, October 15th through 19th.

Because of the growing importance of communications, emphasis will be placed on new aspects in this field—FM stereo broadcasting and modern telephony. According to Convention Chairman H. E. Roys, the fifteen sessions this fall will interest not only engineers, but also audio personnel and technicians in specialized branches who regularly attend the Society's Conventions.

Recording Techniques

People concerned with disc recording and reproduction will have a field day—three sessions devoted to these topics alone. Recording techniques in Europe will receive considerable attention also, during which time papers written by research engineers at Telefunken-Decca, Germany, Philips Phonographic Industries, the Netherlands, Ortofon Industry A/S of Denmark, along with others, will be presented.

At the session sponsored by IRE-PGA, several papers on Stereophonics will be read, in addition to a panel discussion of the topic, "What Hath Stereo Wrought."

Guest speaker at the Annual Banquet to be held in the Barbizon Room on Thursday, October 18, at 7:00 P.M. will be George R. Marek, vice president and general manager, RCA Victor Record Division, who will speak on "Sound—History and Future." Presentation of awards and fellowships by the Society will take place at this time.

Exhibits

Due to the success of previous Professional Products Exhibits, the Society will hold its fifth exhibit this year.

The program this year has been organized by H. E. Roys, chairman, R. C. Moyer, vice chairman, Benjamin B. Bauer, Harold Bodel, Warren L. Braun, Robert W. Carr, Gilbert F. Dutton, J. Donald Harris, F. K. Harvey, H. Philip Iehle, Irving Joel, Ronald D. Klumpen, C. J. LeBel, A. H. Lind, Rein Narma, D. J. Plunkett, E. H. Uecke, J. E. Volkmann, D. R. von Recklinghausen, and P. B. Williams.

Sessions

The complete list of technical sessions follows:

Monday, October 15
- 9:00 A.M.—Annual Business Meeting
- 9:30 A.M.—Microphones and Earphones
- 1:30 P.M.—Audio Electronics
- 7:30 P.M.—Loudspeakers

Tuesday, October 16
- 9:30 A.M.—Disc Recording and Reproduction I
- 1:30 P.M.—Disc Recording and Reproduction II
- 7:30 P.M.—Recording Techniques in Europe
- 7:30 P.M.—Music and Electronics

Wednesday, October 17
- 9:30 A.M.—Magnetic Recording
- 1:30 P.M.—Requisites of Modern Telephony
- 7:30 P.M.—Stereophonics—(Sponsored by IRE-PGA)

Thursday, October 18
- 9:30 A.M.—Sound Reinforcement and Acoustics
- 1:30 P.M.—FM Stereo Broadcasting I
- 7:00 P.M.—Annual Banquet—Presentation of Awards

Friday, October 19
- 9:30 A.M.—FM Stereo Broadcasting II
- 1:30 P.M.—Broadcast Audio/Studio Equipment
- 7:30 P.M.—Psychoacoustics

Exhibit Hours—Audio Engineers Show

Tuesday through Friday October 16–19, 1962 — Noon to 6:45 P.M., except Thursday and Friday to 5:00 P.M.

I WISH SOMEBODY WOULD COME UP WITH A GOOD **LOW COST TWT** FOR APPLICATIONS WHERE NOISE LEVEL ISN'T CRITICAL.

HAVEN'T YOU HEARD ABOUT GE'S...

?

THEIR WHAT? DON'T TALK IN PICTURES. IT ISN'T NATURAL.

I WONDER IF TALKING LIKE A CATALOG IS NATURAL?

THEN IF IT'S **FACTS** YOU WANT, HERE THEY ARE... GENERAL ELECTRIC'S NEW, LOW COST ZM-3105 TWT IS a rugged, dependable, "high-performance-in-critical-temperatures" TWT. General Electric is now offering the ZM-3105 at about half the usual price for such a product.

The ZM-3105 operates in ambient temperatures of −65 to +100 degrees C. Helps solve your hot and cold installation problems. G.E. has made the ZM-3105 tough, too. During endurance tests it holds up under 50-G shock and 10-G vibration from 55 to 2000 CPS. Looks like it'll lead to a whole new line of TWT's for application where noise factor is not a major consideration.

ZM-3105 characteristics include . . .

Frequency	7000 to 11,000 Megacycles
Heater	
Voltage	6.3 Volts
Current, nominal	0.4 Ampere
Focusing Method—Periodic Permanent Magnet	
Noise Figure, maximum	25 Decibels
Small Signal Gain, minimum	30 Decibels
Saturation Power Output, minimum	20 Milliwatts
Collector Dissipation	1.0 Watt
Impedance, Coaxial	
Input, VSWR	Less than 2.5 to 1
Output, VSWR	Less than 2.5 to 1

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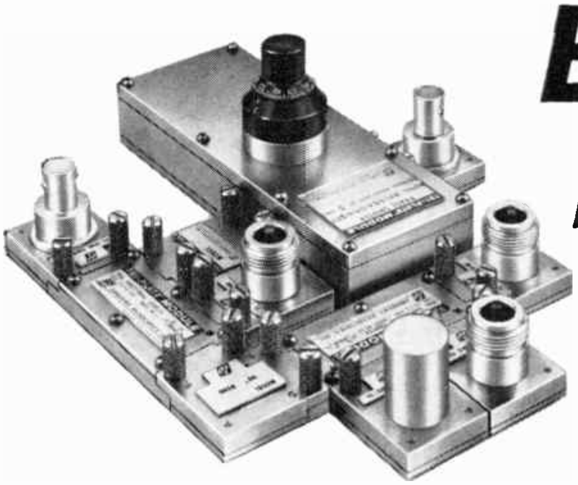
Microwave, semiconductor and fast switching circuits with high density packaging, that heretofore have been thought impractical or impossible to build because of their bulk or complexity, can be successfully produced in TRI-PLATE Strip Transmission Line. □ In breadboarding, packaging and quantity production, what the concept of strip transmission promised, TRI-PLATE techniques deliver — they've made the concept a practical reality! □ New circuit ideas — no matter how different or daring — are tested quickly, easily and economically with TRI-PLATE Strip Transmission Line Modules. They let you go from paper schematics to functioning circuits in just minutes to evaluate new design concepts. □ A new approach to the design of a Dual Directional Coupler-Variable Phase Shifter, for example, was recently conceived by Bendix engineers. They needed a package that weighed only 8 ounces — and they needed it in just



Evaluate new design concepts with Tri-Plate® modules

30 days! By breadboarding with TRI-PLATE Modules the circuit was proved practical, and Bendix gave Sanders the go-ahead to produce the design in quantity as Integrated TRI-PLATE Packages. □ Production models were delivered on schedule and weighed only 6 ounces! This is but one illustration of the new directions in electronics made possible by TRI-PLATE Products. □ There are more than 600 TRI-PLATE Modules — including over 150 TRI-PLATE Mounts for standard and advanced semiconductor devices — available to help you speed the time from design to production. And systems designed in Modules can be produced in quantity as Integrated TRI-PLATE Packages, with performance equal to or better than the modular prototype, and with great savings in size and weight. □ For further information about TRI-PLATE Products — including specifications and prices — or for consultation regarding your specific requirements, write to Sanders Associates, Inc., Microwave Products Department, Nashua, New Hampshire.

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CREATING NEW DIRECTIONS IN ELECTRONICS
SANDERS TRI-PLATE® STRIP TRANSMISSION LINE

Professional Group Meetings

Aerospace and Navigational Electronics

Philadelphia—December 14

"Dynasoar Communications and Tracking Subsystems," J. Byron Garrett, Jr., RCA, Camden, N. J.

Philadelphia—February 22

"The Computer Simulation of Space Vehicle Characteristics," Harold G. Tremblay, US Naval Air Development Center, Johnsville, Pa.

Antennas and Propagation Microwave Theory and Techniques

Philadelphia—November 15

"Microwave Optic Frontiers," L. B. Lambert, Columbia University.

Philadelphia—May 16

"This Microwave Business," T. N. Anderson, Budd-Stanley, Syosset, N. Y. "Electronically Steerable Antenna Systems," R. M. Scudder, RCA, Moorestown, N. J.

San Diego—May 8

"Tunnel Diode Circuits at Microwave Frequencies," C. L. Cuccia, RCA, Los Angeles, Calif.

Bio-Medical Electronics

Los Angeles—June 14

"Automated Data Processing for a Modern Hospital—A Simulated Study," H. H. Wilson, Systems Development Corporation, Santa Monica, Calif.

Memphis—June 14

"Neurosurgical Aspects of Medicine," C. D. Ray, M.D., University of Tennessee.

Rochester—June 11

"Flow Transducers—Electrical Problems in Long Term Implanted Transducers for Study of Blood Flow," Dr. Frederick Olmsted, Western Reserve University, Cleveland, Ohio.

San Francisco—March 20

"DADTA (Discrimination Apparatus for Discrete Trial Analysis)," Drs. Karl Pribram and Dan Kimball, Stanford Medical School; Gerald Pressman, Stanford Research Institute.

Broadcasting

Vancouver—January 15

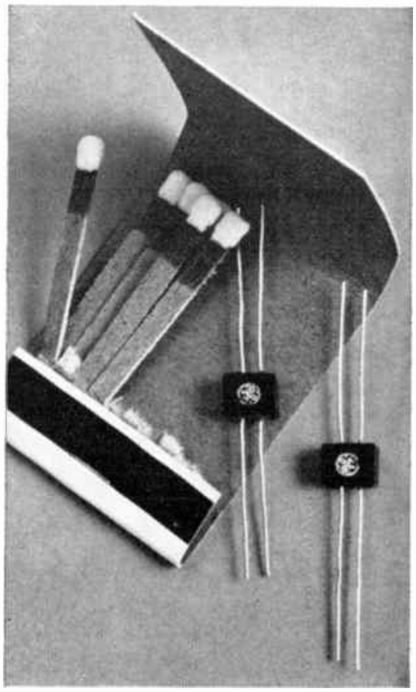
"Field trip to studios of CHAN-TV," E. G. Rose, CHAN-TV, Burnaby, B. C.

(Continued on page 134A)

ELECTRONICS progress in semiconductors

Matching Peak to Peak

. . . has at least one thing in common with dancing cheek to cheek . . . you can't get much closer. The matching we mean occurs in the new TDP-1 through TDP-3 germanium tunnel diode matched pairs which have peak point currents of 1.0, 2.2, and 4.7 ma. Each pair is matched for peak point current to within 1% and for peak point voltage to within 5 mv over the temperature range of 15°C to 55°C. Each pair is also matched for dynamic peak point current to within 1% at a frequency of 10 megacycles. You just don't get 'em any closer than that, on a dance floor or in a majority logic computer circuit.



They're characterized for use as high speed voltage and current comparators, as sense amplifiers and in pulse modulation and sampling systems. As for performance, a 1 ma TD pair will sense a 10 microampere current change in about 10 nanoseconds. Incidentally, inside that small epoxy case we use the G-E developed subminiature axial lead package which features very high junction mechanical strength capability and low case capacitance (which is one of the reasons these tunnel diodes matched pairs are so efficient).

Did you know that if you drop a transistor only 4½" onto a hardwood bench you can cause up to 5,000 g shock? Avoid common semiconductor handling abuses. They can't do you or the transistor any good.

The Offer Still Goes

A while ago in these columns we suggested that you get a group together at lunchtime and watch a free movie in color. "Departure for Perfection" runs 25 minutes, tells the why and how of semiconductor manufacturing, including semiconductor theory, construction techniques, and reliability criteria. Good coverage of semiconductor fundamentals. and enjoyable too. Favorable response prompted us to renew the offer. All we ask is that you write to us on your company letterhead for a copy of the print, and that you return the print when you're through with it. Fair enough? Write to Section 231135.

Reliability based on 10 years of experience is pretty hard to beat. We've been making and testing germanium alloy and rate grown transistors for that long. The 2N43 series, for example, has life test charts covering over 50,000 hours. Next time you need the economy and reliability of germanium transistors in the complete frequency spectrum from 1 to 20 mc, call your G-E District Sales Manager.

Any questions? Write us at Section 231135, General Electric Company, Semiconductor Products Department, Electronics Park, Syracuse, New York. In Canada: Canadian General Electric, 189 Dufferin St., Toronto, Ont. Export: International General Electric, 159 Madison Ave., New York 16, New York.

GENERAL ELECTRIC

Professional Group Meetings

(Continued from page 133A)

Communications Systems Information Theory

Los Angeles—January 30

"Solid and Liquid Lasers," Dr. T. Maiman, Quantatron Inc., Santa Monica, Calif.

"Operational Maser Systems for Space Communications," Dr. W. Higa, Jet Propulsion Lab., Pasadena, Calif.

Electron Devices

Albuquerque-Los Alamos—June 18

"Electron Devices in Europe," G. C. Dacey, Sandia Corporation, Sandia Base, Albuquerque, N. M.

Los Angeles—June 12

"Ionic Propulsion For Space Vehicles," J. R. Anderson, Hughes Research Laboratories, Malibu, Calif.

San Francisco—May 30

"Helium-Neon Gas Phase Optical Maser," William Bell, Spectra-Physics Inc., Mountain View, Calif.

Schenectady—June 12

"Airborne Particles," T. A. Rich, General Electric Co., Schenectady, N. Y.

Electronic Computers

Omaha-Lincoln—January 25

"Medical Applications of Computers," Barney Watson, V. A. Hospital, Omaha, Neb.

Philadelphia—May 8

"Report on Cobol."

Moderator: A. J. Allott, U. S. Ordnance Management Office.

Panel: H. S. Bright, Philco; A. T. Hill, Philadelphia Marine Quartermaster Depot; A. S. Kransley, RCA; and R. H. Regiani, Ordnance Tank Automotive Command.

Pittsburgh—December 11

"Machine Data Plotting," J. E. Fedako, R. H. Delgado, P. G. Rankin, Gulf Research and Development Labs., Hammarville, Pa.

Pittsburgh—February 28

"Dynamic Programming" (tutorial), Dr. Irving Lefkowitz, Case Institute of Technology, Cleveland, Ohio.

Pittsburgh—May 14

"MAC Hybrid and Computer Systems," L. R. Abkemeier, MacDonnell Aircraft.

(Continued on page 136A)

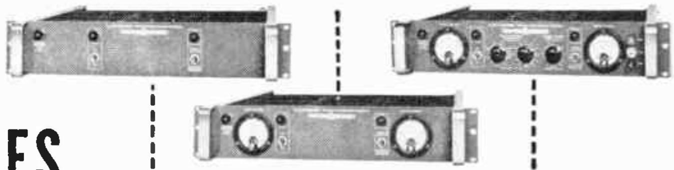
Only Lambda offers:

Environment-engineered power supplies mass-produced at Competitive Prices

NEW LE SERIES

WITHOUT METERS
(Controls in Rear)
LE 101

WITH METERS
(Front Panel Control)
LE 101-FM



WITH METERS
(Controls in Rear)
LE 101-M

DC OUTPUT
(Regulated for line and load.)

0-36 VDC	5 Amp	LE 101	$420	LE 101-M	$460	LE 101-FM	$470
0-36 VDC	10 Amp	LE 102	525	LE 102-M	565	LE 102-FM	575
0-36 VDC	15 Amp	LE 103	595	LE 103-M	635	LE 103-FM	645
0-36 VDC	25 Amp	LE 104	775	LE 104-M	815	LE 104-FM	825
0-18 VDC	8 Amp	LE 105	425	LE 105-M	465	LE 105-FM	475
0- 9 VDC	10 Amp	LE 109	430	LE 109-M	470	LE 109-FM	480

Plus THESE FEATURES:

- All solid state
- Completely protected
- Adjustable automatic current limiting
- Continuously variable
- Remotely programmable over entire range
- Constant voltage/constant current
- Wide input frequency and voltage range
- Convection cooled—no blowers
- 50°C ambient
- Guaranteed 5 years

CONDENSED TENTATIVE DATA

REGULATED VOLTAGE:

Regulation

(line and load) ...... Less than .05 per cent or 8 millivolts (whichever is greater). For input variations from 105-135 VAC and for load variations from 0 to full load.

Transient Response

(line) .... Output voltage is constant within regulation specifications for any 15 volt line voltage change within 105-135 VAC.

(load) .... Output voltage is constant within 25 MV for load change from 0 to full load or full load to 0 within 50 microseconds of application.

Remote Programming .... 50 ohms/volt constant over entire voltage range.

Ripple and Noise .... Less than 0.5 millivolt rms.

Temperature Coefficient .. Less than 0.015%/°C.

AC INPUT: .... 105-135 VAC; 45-66 CPS and 320-480 CPS in two bands selected by switch.

OVERLOAD PROTECTION:

Thermal .... Thermostat, reset by power switch, thermal overload indicator light front panel.

Electrical:

External Overload

Protection .... Adjustable, automatic electronic current limiting.

METERS: .... Ruggedized voltmeter and ammeter to Mil-M-10304B specifications on metered models.

PHYSICAL DATA:

Mounting .... Standard 19" rack mounting.

Size .... LE 101, LE 105, LE 109 3½" H x 19" W x 16" D
LE 102 .... 5¼" H x 19" W x 16" D
LE 103 .... 7" H x 19" W x 16½" D
LE 104 .... 10½" H x 19" W x 16½" D

SEND FOR COMPLETE DATA ON LE SERIES

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Southeastern Region: W. A. Brown & Associates, Inc., Engineering Representatives
Orlando, Fla. • Fort Lauderdale, Fla. • Huntsville, Ala. • Alexandria, Va. • Winston-Salem, N. C.

LA133

Professional Group Meetings

(Continued from page 134A)

"Hybrid Simulation of a Missile," R. Gellman, General Electric Co.

"Progress on CADDA," M. Gisser, National Bureau of Standards.

"Hybrid Computer Operation," R. Herbst, Westinghouse Electric Corp.

"Study of the L-D Steel Making Process Using a Hybrid Computer," R. T. P. Putman, Westinghouse.

"Combined Analog Digital System Versus Integrated Hybrid System," T. D. Truitt, Electronic Associates Inc.

Pittsburgh—May 21

"Communication Between Computers," Neil Clark, Control Data Corp., Minneapolis, Minn.

San Francisco—April 24

"List Processing and Practical Problems," J. Weizenbaum.

Engineering Management

Chicago—June 14

"NEC Policies and Plans," James Kogan, General Precision Equipment, Chicago, Ill.

Philadelphia—April 24

"Project Management," S. Sadin, G-E, MSVD, Philadelphia, Pa.

Engineering Writing and Speech

Northern New Jersey—June 14

"Motivation in Writing," Dr. E. J. Piel, West Essex Regional H. S., North Caldwell, N. J.

Washington, D. C.—February 21

"Government Evaluation of Your Technical Proposals," C. B. Palmer, NASA.

Human Factors in Electronics Reliability and Quality Control

Los Angeles—April 17

"The Meaning of .999 Reliability," A. L. Gichtin, Aerospace Corp.

"The Effects of Humans on System Reliability," A. M. Freed, Aerojet General.

"System Design on the Mercury Project," D. R. White, Space Technology Labs.

Information Theory

Los Angeles—April 9

"Receivers for Randomly Varying Channels," Dr. T. Kailath, Jet Propulsion Lab., Pasadena, Calif.

"Machine Learning in Automatic Pattern Recognition," Dr. D. Braverman, CIT, Pasadena, Calif.

(Continued on page 138A)

Professional Group Meetings

(Continued from page 136A)

Instrumentation

Long Island—April 17

"Dockside Testing of Instrumentation on Mobile Atlantic Range Stations," E. L. Roel, Sperry Gyroscope Co., Great Neck, N. Y.

Long Island—September 26

"Seminar on Phase and Time Measurement," Dr. Paul Yu, Ad-Yu Electronics Lab., Inc., Passaic, N. J.

Long Island—November 26

"Tunnel Diodes—Their Characterization and Use in Control Systems and Instrumentation," V. A. Vulcan, General Instrument Semiconductor Div., Hicksville, N. Y.

Microwave Theory and Techniques

Long Island—February 20

"Interaction of Microwaves and Plasmas," Dr. Nathan Marcuvite, Polytechnic Institute of Brooklyn.

Long Island—April 24

"This Microwave Business," T. N. Anderson, Budd-Stanley Co., Syosset, N. Y.

Los Angeles—June 7

"Recent Developments in Parametric Electron Beam Devices," Dr. Robert Adler, Zenith.

Orlando—June 6

"Microwave Facility Tour," E. C. Roberts, Southern Bell Telephone and Telegraph Co.

Schenectady—May 8

"Optical Masers,. Dr. Kiyo Tomiyasu, General Electric Co., Schenectady, N. Y.

Microwave Theory and Techniques/Antennas and Propagation

Albuquerque-Los Alamos—May 28

"Dry Calorametric Microwave Power Meters and Measurements," William Sebastin, P.R.D. Electronics, Los Angeles, Calif.

Albuquerque-Los Alamos—June 13

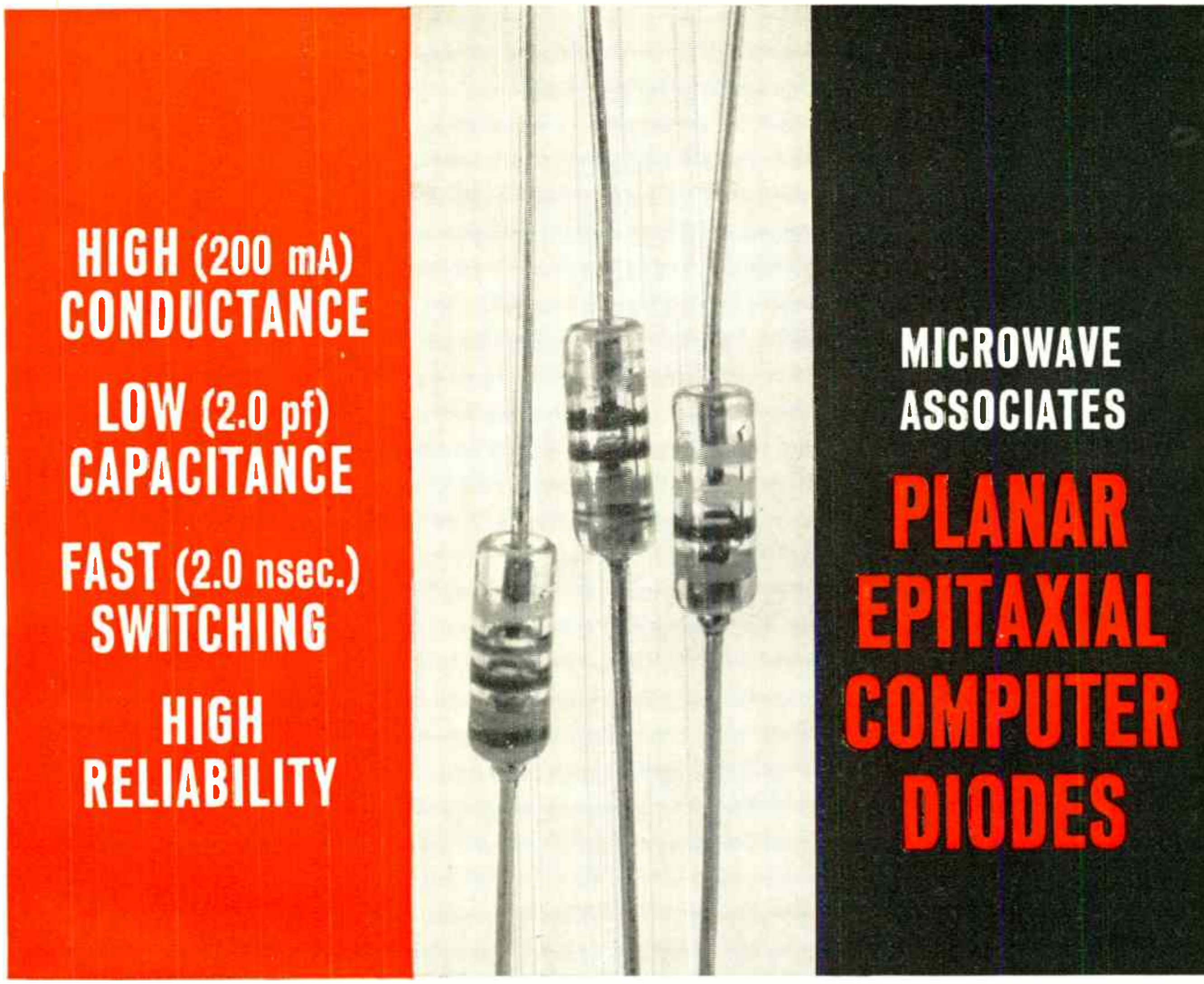
Films: "Sandia Corporation Story," "Big Bounce" "Project Echo.," "Coaxial and Microwave Miracles," "Electrical Safety."

Military Electronics

Long Island—March 13

"Space Surveillance and the Space Track System," Dr. E. W. Wahl, Air Force Systems Command.

(Continued on page 140A)



Designed to incorporate all the parameters to yield the ultimate in a computer diode and made by the planar epitaxial process for greater reliability, Microwave Associates new line of silicon computer diodes are now available for ultra-fast switching, low capacitance applications. ■ The units outlined below are available with 2 nanosecond switching speeds. Maximum capacitance is 2 picofarads with forward conductance levels from 10 to 200 mA. ■ The available range of forward conductances of these diodes up to 500 mA @ 1.0 V, makes them suitable for applications from low current logic to high current memory core driver circuitry. ■ The unique double hermetic seal construction yields the ultimate in moisture resistance and the rugged internal assembly completely eliminates front contact problems. For greater heat dissipation the units have a .040″ Dumet heat sink on the cathode end. They will meet the most stringent electrical and mechanical specifications for aircraft, missile or satellite computers. ■ *Microwave Associates planar epitaxial computer diodes can provide the answer to some of your problems. For complete information call or write:*

TYPE MA-	FORWARD CURRENT 1V DC (mA)	PEAK REVERSE VOLTAGE @ 5μA	CAPACITANCE MAX (μμf)	REVERSE CURRENT (μA) @−50 V		REVERSE RECOVERY (n Sec)	POWER DISSIPATION (mW)
				@ 25°C	@ 125°C		
4441	10	75	2 @ 0V	.05	50	2*	250
4442	20	75	2 @ 0V	.05	50	2*	250
4443	50	75	2 @ 0V	.05	50	2*	250
4444	100	75	2 @ 0V	.05	50	2*	250
4445	200	75	2 @ 0V	.05	50	2*	250
4446	500	75	8 @ 0V	.1	100	150**	500

*Recovery time to 1 ma when switched from 10 ma to −6V through 100 ohm loop impedance
**Recovery time to 5 ma when switched from 500 ma to −50V through 1000 ohm loop impedance.

MICROWAVE ASSOCIATES, INC.

United Kingdom Sales: Microwave and Semiconductor Devices, Ltd. Skimpot Trading Estate, Luton, Bedfordshire, England
Other Export Sales: Microwave International Corp. 36 W. 44th Street, N.Y.C., N.Y., U.S.A. Cable: Microken

BURLINGTON, MASSACHUSETTS • WESTERN UNION FAX
TWX: BURLINGTON, MASS. 942 • BROWNING 2-3000

Professional Group Meetings

(*Continued from page 138A*)

San Francisco—June 19

"Mariner R Telemetering System," C. C. Kirsten, Jet Propulsion Laboratory, Los Angeles, Calif.

Reliability and Quality Control

Los Angeles—May 21

"Reliability Approach for Midas," R. A. Orr, Aerojet General Corp., Azusa.

"Parts Application and Procurement Specification," R. L. Krider, Aerojet General Corp., Azusa.

"Reliability Program Implementation and Results," H. J. Walther, Aerojet General Corp., Azusa.

Philadelphia—May 22

Tour of RCA Space Environment Center, Hightstown, N.J.

Reliability and Quality Control/Component Parts

Los Angeles—June 11

"Solid TA Capacitors—Minuteman Rel. Prog.," Bernard Hecht, Sprague Electric, North Adams, Mass.

Space Electronics and Telemetry

Los Angeles—April 17

"Mautel—Microminiaturized Autonetics Telemetry," T. R. Denigan, Autonetics, Anaheim, Calif.

"The General Purpose Digital Computer—PCM Ground Station," C. A. Walker, United Electro Dynamics, Oxnard, Calif.

Los Angeles—May 15

"Tanlock—Correlation Detector," L. M. Robinson, North American Aviation, Downey, Calif.

"Efficient Time Sharing of Sampled Data as Applied to Spacecraft Telemetry Systems," John Meyer, Jet Propulsion Lab., Pasadena, Calif.

San Francisco—May 15

"Command and Communication Systems for Future Space Vehicles," R. E. Klokow and R. G. Davis, LMSC Research and Engineering.

Washington, D.C.—May 22

Panel Discussion: "Telemetry in the Sixties, Its Problems and Promises."

Panel Members: Benn Martin, Jet Propulsion Lab.; James Rorex, Marshall Space Flight Center; F. B. Smith, Langley Research Center; R. W. Rochelle, Joseph Purcell and Herman LaGow, Goddard Space Flight Center.

(*Continued on page 142A*)

Professional Group Meetings

(Continued from page 140A)

Vehicular Communications

Chicago—April 13

"Mobile System Field Strength Survey," G. A. Olive, RCA.

Chicago—June 12

"Skyphone," J. E. Cain, General Motors Corp., Delco Radio Div.

Los Angeles—January 18

"Open Forum on Mobile Radio Maintenance," A Panel of Experts with Frank Ashby as Moderator.

Los Angeles—February 16

Field Trip through the Communication Headquarters facilities—Southern California Edison Co., Alhambra, Calif.

Section Meetings

(Continued from page 120A)

Lancaster

Tour of Armstrong Cork Co.; 4/10/62.

"Operational Characteristics of the Comlognet System," J. A. Kalz, RCA; Election of Officers; 5/22/62.

New Hampshire

"Stereo—FM—Multiplex," D. R. VonRecklinghausen, H. H. Scott Inc.; 4/4/62.

"Project Westford," R. M. Lerner, MIT Lincoln Labs.; 6/7/62.

"Electronic Organs," W. R. Reneker and R. Campbell, Kinsman Corp.; Election of Officers; 6/19/62.

Pasadena

"Radio Astronomy & Expansion of the Universe," A. R. Sandage, Staff Astronomer; Mt. Wilson & Palomar Observatories; 2/21/62.

"Synchronous Communications Satellites," H. Rosen, Hughes Aircraft Co.; 4/18/62.

"Recent Advances in the Production of Intense Magnetic Fields," A. F. Hildebrant, J. P. L.—N.A.S.A.; 5/16/62.

"Recent Developments in Microelectronics," M. B. Prince, Electro Optical Systems; 6/20/62.

Southern

Plant Tour—Nortronics; Election of Officers; 6/20/62.

Southern Alberta

"Application of Modern Electronics to Seismic Explorations," H. H. Heffering, Imperial Oil Co.; Election of Officers; 5/17/62.

Southwestern Ontario

Discussion of Proposed IRE/AIEE Merger—D. V. Stocker, Wayne State Univ.; T. W. W. Stewart, University of Western Ontario; Election of Officers; 5/15/62.

Victoria

"Television Antenna Distribution System for Apartment Blocks," R. V. Mielen, Tele-Tech Service Ltd.; 4/25/62.

"Forest Values and Wood Uses," W. Reith, B. C. Forest Service Dept.; 5/23/62.

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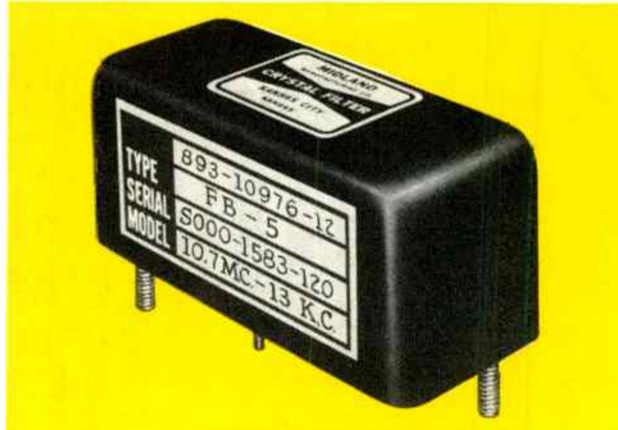
SOLUTION:

1 Midland filters with guaranteed ultimate discrimination of more than 100 db with 60db/3db BWR < 1.8 **2** A low cost stock filter with virtually no insertion loss

FACT Midland crystal filters are the result of exact design methods and real production knowhow.

Facts are facts and filters are Midland's business. Their filter and crystal engineering skills and facilities assure the user of top reliability and performance. * This is Midland's Type FB-5 crystal filter produced by the tens of thousands — the only sure proof of production ability. It is an 8 pole — 6 zero precision network that incorporates no added dissipative elements in inband ripple control. Result: Superior selectivity with essentially no midband insertion loss. A quality production component with immediate delivery. Engineering Bulletin NBS-103 is available detailing complete technical information. Prices on request.

** Write for Midland's capabilities and facilities brochure, "Midland — in microspect".*



SPECIFICATIONS

Center Freq: 10.7 MC ± 375 CPS
Bandwidth @ 6 db.: 13.0 KC Min. — 13.8 KC Max.
60 db/6 db BWR: 1.8 Max.
100 db/6 db BWR: 2.2 Max.
Ultimate Attenuation: 105 db. Min., 8 MC to 14 MC
Midband Insertion Loss: 0.5 db. Nominal, 1 db. Max.
Inband Ripple: 0.5 db. Nominal, 0.8 db. Max.
Operating Temp. Range: —55° C to+90° C
Zin/Zout Req: 1100 OHMS ± 5% in parallel with adjustable capacitor 0-5 picofarads.
Dimensions: 2 3/8" L x 1" W x 1 1/32" H

Midland MANUFACTURING COMPANY
Division Pacific Industries, Inc.
3155 Fiberglas Road Kansas City 15, Kansas

PROBING THE "TRANSPROBE" WAVEGUIDE SWITCH

The only moving parts of the "Transprobe" are solenoid plunger and a light weight, dielectrically supported metal probe. Result . . . greater flexibility, longer life, ability to switch under full waveguide power. The design can be applied to any waveguide size. Unit shown is a single-pole-double-throw, X-band switch. The design is equally successful for adaptation to SPDT, SP3T, transfer switch or special configurations. Typical specifications: Frequency . . . 8.2-12.4 KMC, VSWR . . . 1.20, Insertion Loss . . . 0.2 db, Crosstalk . . . 35 db, Life . . . 2,000,000 operations. To probe more thoroughly the unusual advantages in this new approach to waveguide switching write Transco Products, Inc., 12210 Nebraska Avenue, Los Angeles 25, California.

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Measure - - - - - - - - - MILLIVOLTS - - - - - - - - - NANOAMPERES accurately at low cost with DYNATRAN's - - -

Model 1829

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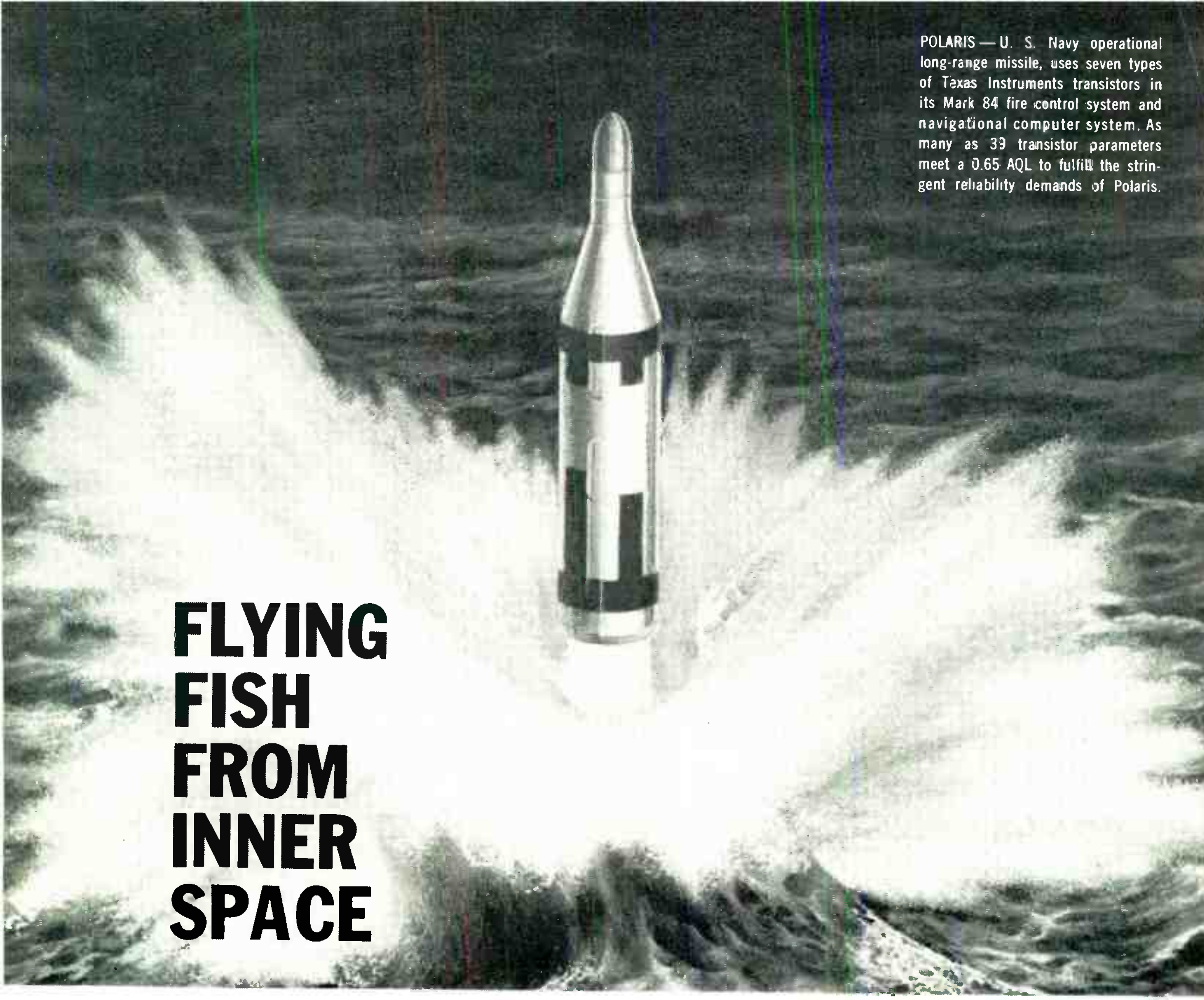

Rely On TI Advanced Transistor Technology

As Many As 39 Parameters Meet 0.65 AQL . . .

Now you can design this same Polaris transistor performance into your circuits, products or systems. TI's Circuit Development-Applications department would welcome the opportunity to consult with you on your military or industrial design requirements. ■ The following TI device types are the counterparts of the Polaris units delivered to a 0.65 AQL. They are produced by the same technology, equipment and personnel that produce "Polaris quality" transistors . . . and made possible by TI's capabilities in precise process control. **2N1047 series, 2N1893/2N696 series, 2N1132 series, 2N743 series, 2N490 series, 2N1302 series, 2N428 series.** When you add TI reliability to your products, you add it at low cost because many years of high-volume production have allowed TI to develop extensive mechanization techniques. ■ More than 800 separate automatic production and testing facilities provide the manufacturing flexibility to meet your special product requirements . . . again at low cost. ■ You will find the transistors you need among the over 400 standard TI device types already in stock for your convenience. ■ Put TI transistors and TI's creative engineering assistance to work for you to provide better reliability for your products at lower cost. Call your TI sales representative, or write directly to TI's Circuit Development-Applications department about your specific requirements. You'll find out why so many already specify TI.

TRANSISTOR PRODUCTS DIVISION

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a New Concept

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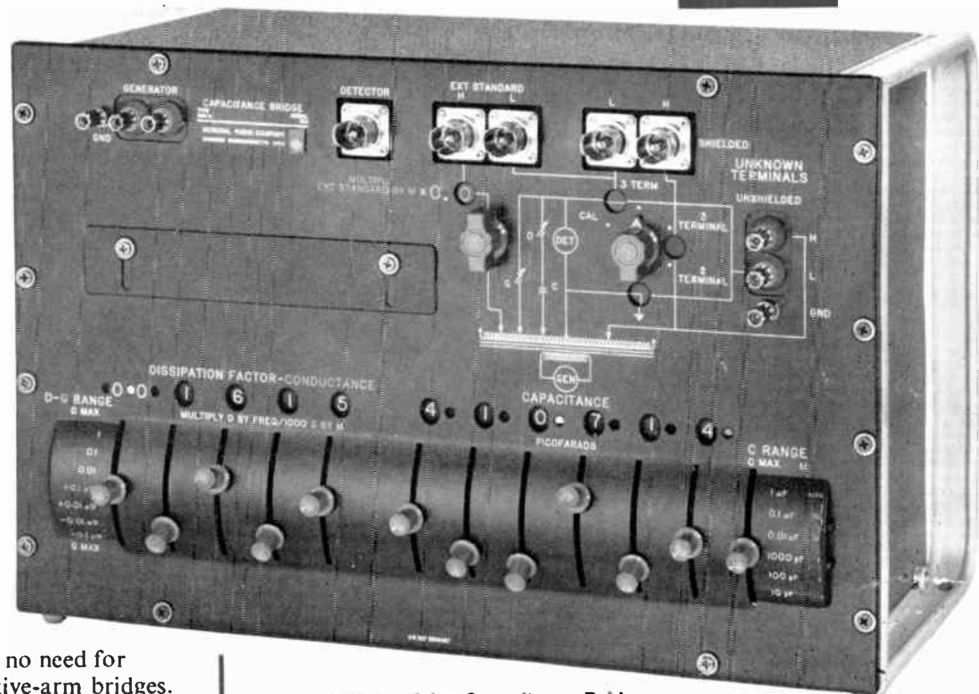
with lever balancing controls, digital readout, automatic decimal point location and unit indication greatly simplifies balancing.



RANGE, 10 μpf to 1μf

DIRECT READING ACCURACY, ±0.01%

RESOLUTION, 1 ppm

- ★ A transformer ratio-arm capacitance bridge having the capability of one-step 3-terminal capacitance measurement — no need for secondary balances as required by resistive-arm bridges.
- ★ Measures 3-terminal and grounded or ungrounded 2-terminal capacitors. A ground capacitance of 1μf produces an error of only 0.01% in the measurement of a 1000-pf capacitor.
- ★ Highly stable and accurate internal capacitance standards . . . all standards are made from Invar alloy, and the six largest are hermetically sealed in dry nitrogen, resulting in a stability of better than 5 ppm/°C.
- ★ All internal standards can be quickly checked against each other for consistency. Only a single external standard is required to establish the absolute calibration of the entire set.
- ★ Fast one-step intercomparisons of 3-terminal capacitors differing in value by 10,000 to 1 can easily be made.
- ★ When loss of unknown capacitor is less than that of standard, indication is given directly in terms of G.
- ★ Bridge circuit on panel automatically indicates proper bridge connections for each measuring situation.

Type 1615-A Precision Capacitance Bridge

Capacitance Range (6 ranges): 10^{-17} to 10^{-6} farads (10 μpf to 1μf), direct reading; 6-figure resolution, smallest division 10^{-17} farads. Ranges can be extended by use of external standards.

Dissipation-Factor Range (3 ranges): 0.000001 to 1 at 1 kc, direct reading. Directly proportional to frequency at other frequencies.

Conductance Range (2 + ranges; 2 − ranges): 10^{-6} μmho to 100 μmho; independent of frequency; 4-figure resolution, smallest division 10^{-6} μmho.

Accuracy: *Capacitance,* ±0.01%; direct reading, with internal standards; at high frequencies, error is 0.002% $C_{\mu f}\left(\frac{f}{1000}\right)^2$
Capacitance, approximately 1 ppm when comparing against external standards.
Dissipation factor, ±(0.1% + 7 ppm) of measured value.
Conductance, ±1% + 0.0001 μmho.

Frequency Range: Approximately 100 cycles to 10 kc.

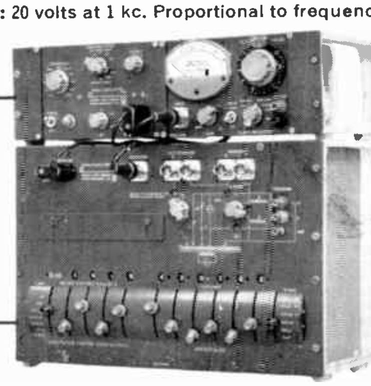
Temperature Coefficients of Internal Standards: About 5 ppm/°C.

Maximum Voltage: 20 volts at 1 kc. Proportional to frequency.

Price: $1475

Complete Capacitance Measuring Assembly.

Type 1620-A . . . includes the Type 1615-A Bridge; Type 1232-A Tuned Amplifier and Null Detector, a low-noise high-gain instrument with a 20-c to 20-kc range and a full scale sensitivity of 1μv; and the new Type 1311-A Bridge Oscillator, with 11 fixed frequencies from 50c to 10 kc. Price for the complete assembly is $2080.



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