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Proceedings of the IRE

in this issue

CAN THE SOCIAL SCIENCES BE MADE EXACT

PANEL TYPE DISPLAY DEVICES

IMPROVED FILM CRYOTRON

GALLIUM ARSENIDE TUNNEL DIODES

REFLECTED-BEAM KINESCOPE

TUNNEL DIODE AMPLIFIER NOISE

NONRECIPROCAL PARAMETRIC DEVICE

SINGLE-SIDEBAND COMPATIBILITY

MAGNETORESISTIVE FLIP-FLOPS

STANDARDS ON NUCLEAR TERMS

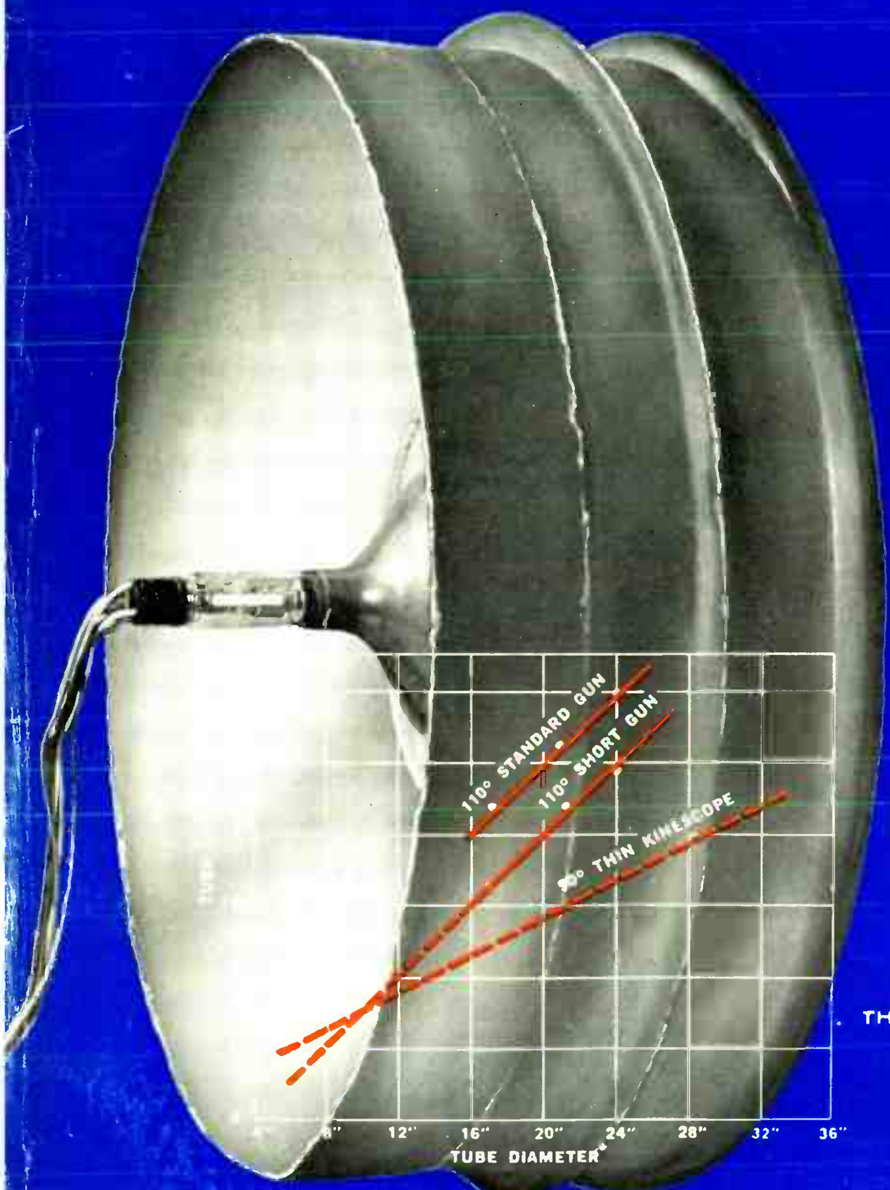
NOISE IN OSCILLATORS

BACKGROUND NOISE IN OSCILLATORS

NOISE IN REGENERATIVE OSCILLATORS

TRANSACTIONS ABSTRACTS

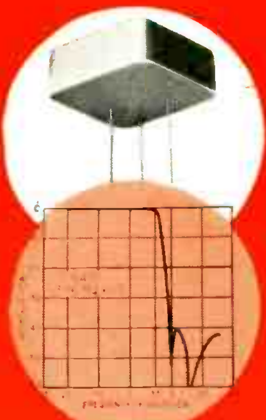
ABSTRACTS AND REFERENCES



THIN KINESCOPE: PAGE 1409

SPECIAL FILTERS

TO YOUR REQUIREMENTS



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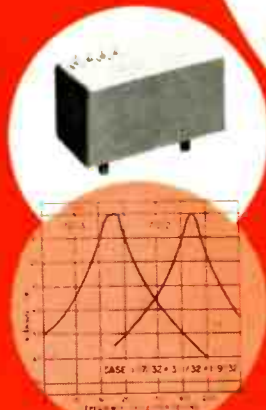

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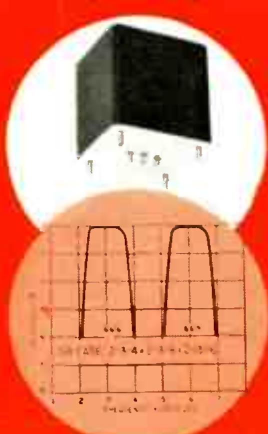
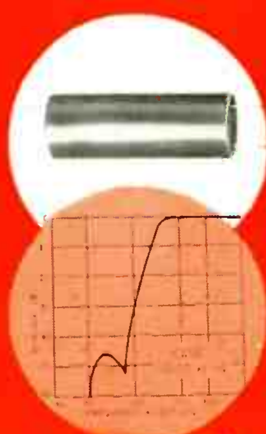


Three KC and 6 KC flat top band pass filters. 600 ohms to 20K ohms. MIL-T-27A each filter 1.7 lbs.



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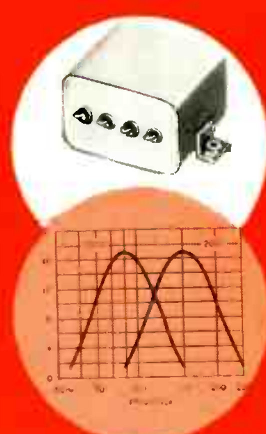

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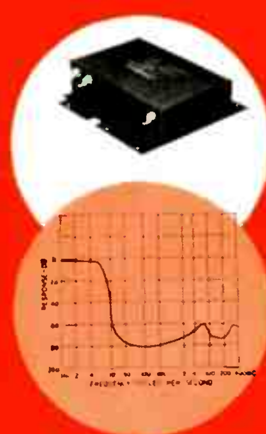

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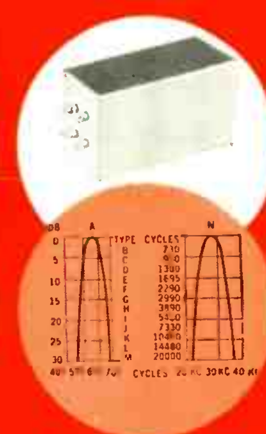

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COVER

A kinescope has been designed at RCA Laboratories in which the electron beam is reflected in a manner which gives the tube an effective deflection angle of nearly 180°. This novel design, reported on page 1409, results in a substantially shorter picture tube.

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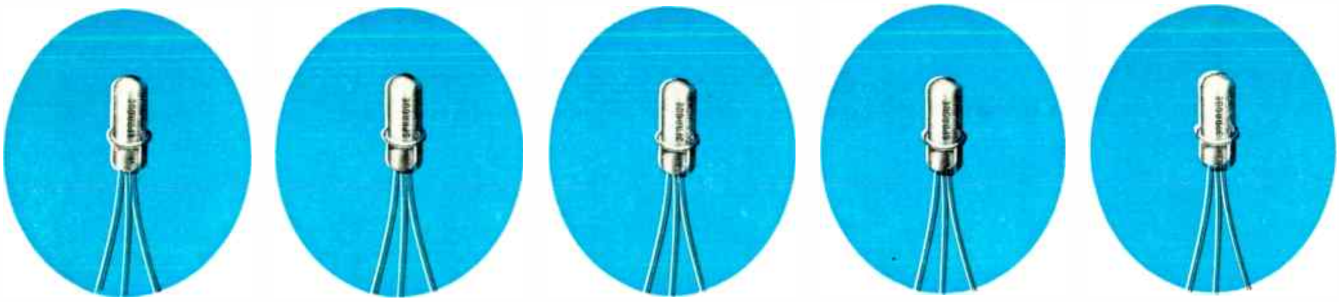

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Frank Arams, Marty Grace, and Sy Okwit of AIL's Department of Applied Electronics have recently been investigating some of the potential uses of ferrimagnetic materials at microwave frequencies. One interesting and useful application arises in the field of low-level limiters. A brief description of this work is given below.

Passive Microwave Limiters*

We have developed S-band and L-band ferrite limiters which have exceedingly low threshold power levels. A photograph of such a limiter with adjustable permanent magnet is shown in Figure 1.

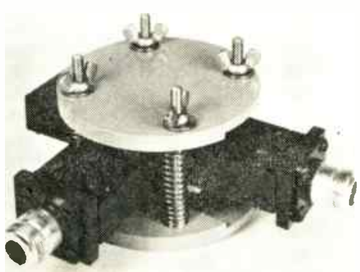

Figure 1.

The circuit arrangement follows the work of DeGrasse (Ref. 1). It consists of two crossed strip-transmission-line half-wavelength resonators between common ground planes which are oriented at right angles to one another, in order to obtain maximum decoupling between them. The resonators are open-circuited at both ends, so that there is an r-f magnetic field maximum at their centers where the resonators cross. A .020″ diameter sphere of single-crystal yttrium-iron garnet (YIG) is placed between the strips. When it is biased to ferrimagnetic resonance by means of the permanent magnet, it serves as the gyro-magnetic coupling element between the resonators.

The driving r-f magnetic field in the input resonator will cause the electron spins in the YIG to precess, thereby inducing an r-f magnetic field in the output resonator. Thus, almost all the r-f power is coupled to the output, via the tightly-coupled YIG sphere, which, behaves as a miniature microwave resonator having a very high unloaded Q. For example, for a YIG ferrimagnetic resonance linewidth of 1 Mc, we calculate an unloaded Q of 10,000 at X-band, and 3000 at S-band. Due to the tight coupling, the bandwidth of the device is substantially greater than the linewidth of the YIG.

* This work was supported by the Department of Defense.

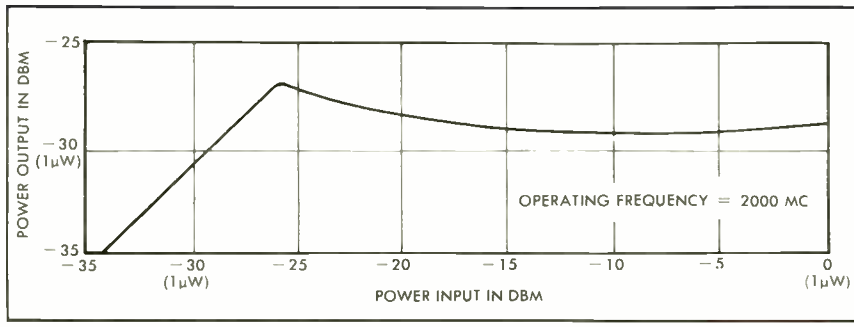

Figure 2.

Thus, the device is a triple-tuned pre-selector having limiting properties due to non-linear effects within the ferrite, which are enhanced by proper choice of frequency, geometry and ferrite parameters (Ref. 2). When the r-f power level is raised, non-linear coupling to (and hence parametric excitation of) lower-frequency spin-waves propagating within the ferrite will occur (Ref. 3). The excitation of the spin-waves prevents the increase of the precession angle of the electron spins beyond a sharply defined threshold and hence limits the power-transfer capability of the YIG sphere.

We have built such limiters at S and L-bands. The transmission characteristic of the S-band limiter is shown in Figure 2. We see that the limiting threshold is quite sharp and occurs near 3 MICRO-watts and that the power output has a plateau that extends over more than a 25 db range. Thus, such a device could be used to protect sensitive microwave receivers against overloads.

Insertion loss of the limiter is 0.6 db at low signal levels, and the 3 db bandwidth is 60 Mc. Off-band rejection is 20 db at 40 Mc from center frequency. Several units can be cascaded, if desired. We have obtained limiting action in similar units from 1000 to 3400 Mc. The device can be tuned by varying the magnetic field and tuning the resonators. A tuning range of 200 Mc has been obtained.

The power handling capability was not measured, but should be high because the limiter operates by REFLECTING the main portion of the incident signal. The VSWR rises continuously from 1.1 below the limiting threshold to 15 and higher for a signal input level 15 db above the limiting threshold. The device exhibits a leading-edge spike of approximately 100-millimicrosecond duration.

We have also constructed a crossed-strip-transmission-line limiter which does not employ resonators, and hence is tunable electronically by varying the applied magnetic field. An electronic tuning range from 2000 Mc to 3400 Mc was obtained with a limiting threshold of −21 dbm.

These are but two examples of novel microwave devices that can be constructed using narrow-linewidth YIG crystals.

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1. R. W. DeGrasse, Jour. Appl. Phys., vol. 30, p 155S, April 1949.
2. R. LeCraw, E. Spencer and C. Porter, Phys. Rev., vol. 110, p 1311, 1958.
3. H. Suhl, Proc. IRE, vol. 44, p 1270, 1956.

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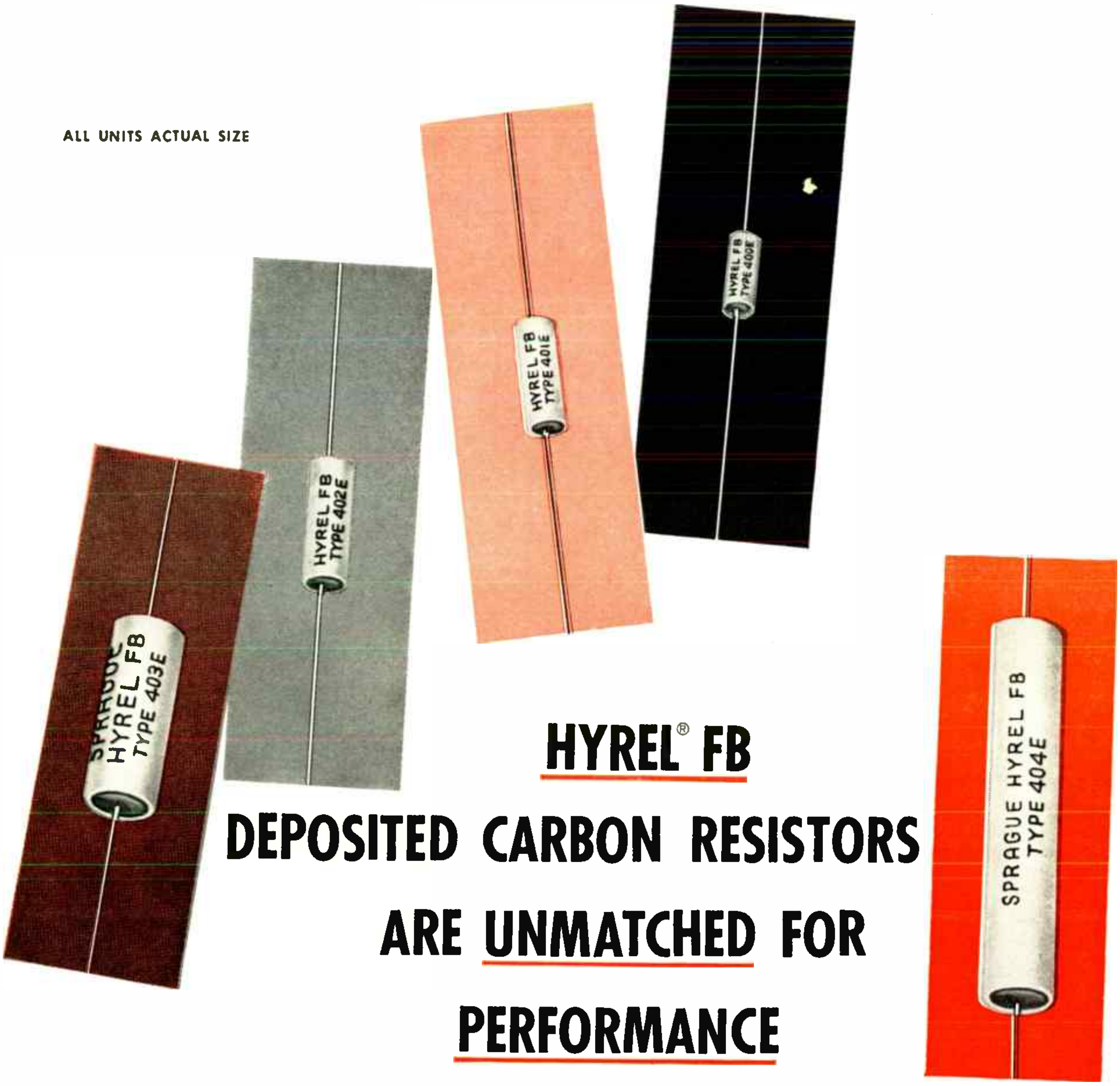

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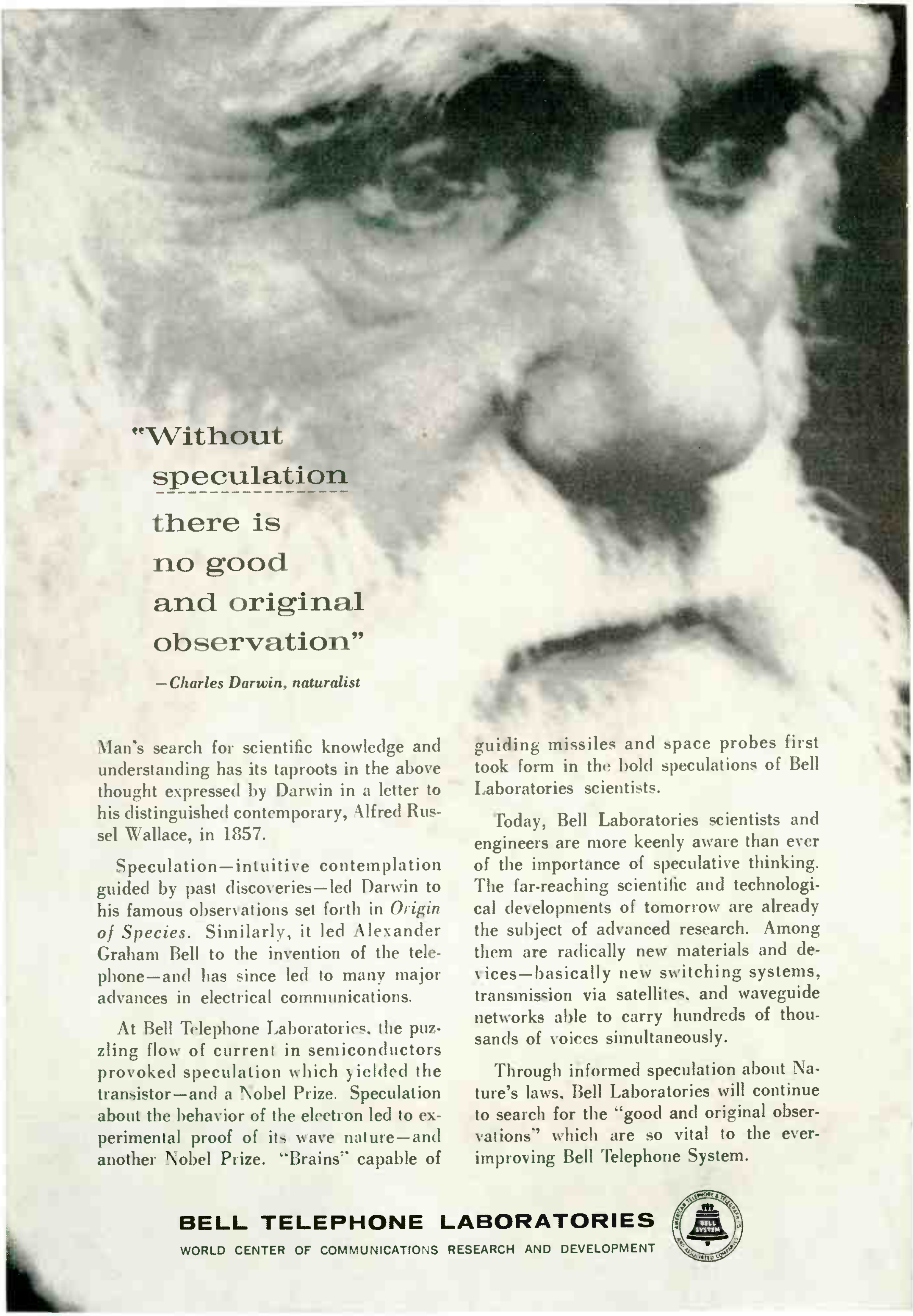

"Without speculation there is no good and original observation"

—Charles Darwin, naturalist

Man's search for scientific knowledge and understanding has its taproots in the above thought expressed by Darwin in a letter to his distinguished contemporary, Alfred Russel Wallace, in 1857.

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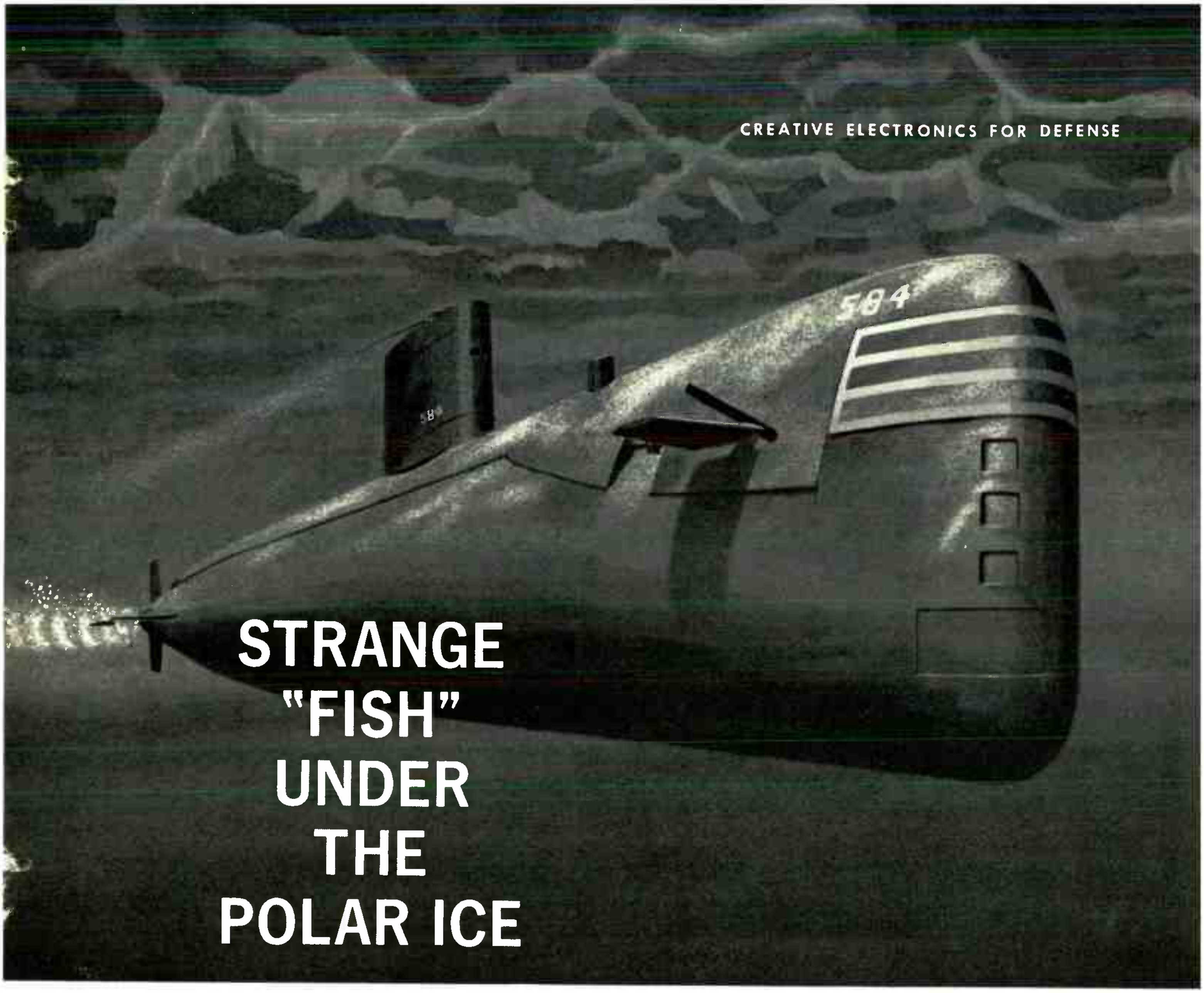

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Meetings with Exhibits

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.

August 23-26, 1960

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

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

September 9-10, 1960

Conference on Communications, Tomorrow's Techniques—A Survey. Roosevelt Hotel, Cedar Rapids, Iowa.

Exhibits: Mr. Ron Gjertson, Cedar Rapids, Iowa.

September 19-22, 1960

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

Exhibits: Mr. Leon King, Jansky and Bailey, Inc., 1339 Wisconsin Ave., N.W., Washington, D.C.

October 3-5, 1960

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

Exhibits: Mr. R. E. Bischoff, 19 Westminster Road, Utica, N.Y.

October 10-12, 1960

National Electronics Conference, Hotel Sherman, Chicago, Ill.

Exhibits: National Electronics Conference, Inc., 228 North La Salle St., Chicago 1, Ill.

October 19-21, 1960

Symposium on Space Navigation, Deshler-Hilton Hotel & Civic Center, Columbus, Ohio

Exhibits: Mr. William P. McNally, 35 Laurel St., Floral Park, L.I., N.Y.

October 24-26, 1960

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

Exhibits: Dr. Harold Schutz, Westinghouse Electric Corp., Air Arm Div., P.O. Box 746, Baltimore, Md.

October 26-28, 1960

Fifth Annual Conference on Non-Linear Magnetics and Magnetic Amplifiers, Bellevue-Stratford Hotel, Philadelphia, Pa.

Exhibits: J. L. Whitlock Associates, 6044 Ninth St., North, Arlington 5, Va.

Oct. 31-Nov. 2, 1960

13th Annual Conference on Electrical Techniques in Medicine & Biology, Sheraton-Park Hotel, Washington, D.C.

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

(Continued on page 10A)

KAY *Precision* RANDOM NOISE GENERATORS

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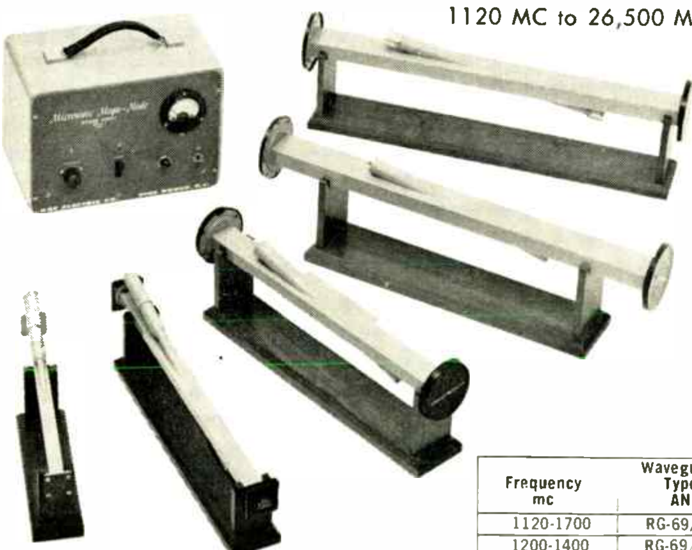

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1700-2600	RG-104/U	UG-435A U	**	870-A	**	$495.00
2200-3300	RG-112/U	UG-553/U	**	880-A	**	$495.00
2600-3950	**RG-48/U**	**UG-214/U**	**261-A**	**260-A**	**262-A**	**$175.00††**
3950-5850	**RG-49/U**	**UG-149A/U**	**271-A**	**270-A**	**272-A**	**$175.00††**
5850-8200	**RG-50/U**	**UG-344/U**	**281-A**	**280-A**	**282-A**	**$175.00††**
7050-10,000	**RG-51/U**	**UG-51/U**	**291-A**	**290-A**	**292-A**	**$175.00††**
8200-12,400	**RG-52/U**	**UG-39/U**	**301-A**	**300-A**	**302-A**	**$175.00††**
12,400-18,000	RG-91/U	UG-419/U	521-A	**	522-A	$250.00
18,000-26,500	RG-53/U	UG-425/U	531-A	**	532-A	$250.00

†† Any three plus power supply: $595.00. Any in excess of three: $167.00 ea.
** None available.
* All prices f.o.b. factory.

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IMPULSE

A DIGEST OF NEW DEVELOPMENTS IN ELECTRONICS AND AUTOMATION

PUBLISHED BY ROME CABLE DIV. OF ALCOA, ROME, N. Y.
PIONEERS IN INSTRUMENTATION CABLE ENGINEERING

DEEP PROBLEMS OF THE BRINY DEEP. While outer space and missiles continue to grab the headlines, the Navy has sounded a big blast and challenge to the electronics system designer. Urgently needed are solutions to a seemingly endless series of problems encountered in the area of Anti-Sub Warfare. There are indications that ASW represents such a pressing need that it may soon reach the proportions of the ICBM program. Of the total of nearly $2 billion earmarked for ASW, some $25 million will go into electronic equipment, most of it sonar devices. The problem basically breaks down into four major areas: surveillance, localization, classification, and killing. One expert singles out the first as the most perplexing of all, but the Navy is ready for full-steam-ahead on all phases. Good starting point for interested parties is BuShips Form 550.5. So, anchors aweigh! Because ASW is a deep problem and the Navy needs you to cope with it.

INFORMATION SQUEEZE. Do you have time to read all the technical journals and periodicals that bear on your work? Equally important, have you a reliable system for classifying, filing, and retrieving articles you want? No simple problem this, it is being investigated jointly by the Special Libraries Association and the American Library Association. Of course, individuals in many firms also have come up with suggestions. One we saw did a very good job of breaking down the entire electrical field into manageable divisions. Another more complex approach is being tried by the chemical people. They are working with a large computer firm on a system that arranges titles by a method called "Key Word in Context." The computer does the work of evaluating titles of articles and then classifying according to "key words." Printout provides groupings of key words and a detailed bibliography.

DUTCH TREAT. Revival of interest in microwave cooking may be given a boost by an electronic oven developed in Holland. This one is designed to cook prefrozen meals at a rate that could run to 150 an hour. Meals come in one end at minus 25°C and come out at serving temperature of plus 80°C. A total microwave capacity of 10 KW is provided by five magnetron elements. Equalization and isolation zones between each unit prevent the magnetrons from affecting one another. Actual heating time for a meal is one minute, but a total of eight minutes in the oven is needed, the remainder being spent in the equalization and isolation zones.

LAND OF THE NEVER-SETTING SUN. A solar battery with 648 silicon cells is being used to power the warning light on top of at least one Japanese lighthouse. It's been in operation since November and has been so satisfactory that plans call for building six more. The flashes can be seen for nine miles.

CABLEMAN'S CORNER. The subject of cable testing is an important one. This is the phase of production that determines whether or not the cable you are purchasing is in accordance with your standards and requirements. In the field of electronics and automation, cables are required to suit various stringent electrical, mechanical, and/or chemical environments. Many years of study and testing have gone into the design of test equipment to be used for these critical tests. It is not enough to know that a cable has been tested in a manner that is "essentially" the same as the required standard. Slight variations in equipment design or methods of tests can mean the difference between conformance and non-conformance. Make sure the test data you receive gives a true picture of the performance of your cable. When you need cable, call on a cable specialist. Phone Rome 3000, or write: Rome Cable Division of Alcoa, Dept. 1280, Rome, New York.

These news items represent a digest of information found in many of the publications and periodicals of the electronics industry or related industries. They appear in brief here for easy and concentrated reading. Further information on each can be found in the original source material. Sources will be forwarded on request.

(21)

Meetings with Exhibits

(Continued from page 8A)

November 14-16, 1960
Mid-American Electronics Convention (MAECON), Hotel Muehlebach, Kansas City, Mo.
Exhibits: Dr. L. R. Crissman, Trans World Airlines, Kansas City, Mo.

November 15-17, 1960
Northeast Electronics Research & Engineering Meeting (NEREM), Boston Commonwealth Armory, Boston, Mass.
Exhibits: Miss Shirley Whitcher, IRE Boston Office, 73 Tremont St., Boston, Mass.

December 1-2, 1960
PGVC Annual Meeting, Sheraton Hotel, Philadelphia, Pa.
Exhibits: Mr. E. B. Dunn, Atlantic Refining Co., 260 S. Broad St., Philadelphia 1, Pa.

December 11-14, 1960
Eastern Joint Computer Conference, Hotel New Yorker, New York, N.Y.
Exhibits: Mr. Alan D. Meacham, 120 E. 41st St., New York 17, N.Y.

January 9-11, 1961
Seventh National Symposium on Reliability and Quality Control in Electronics, Bellevue-Stratford Hotel, Philadelphia, Pa.
Exhibits: Mr. James H. Goodman, Burroughs Research Center, Building #3, Room 3307, Paoli, Pa.

March 20-23, 1961
International Radio and Electronics Show and IRE International Convention, Waldorf-Astoria Hotel and New York Coliseum, New York, N.Y.
Exhibits: Mr. William C. Copp, Institute of Radio Engineers, 72 West 45th Street, New York 36, N.Y.

April 26-28, 1961
Seventh Region Technical Conference and Trade Show. Westward Ho Hotel, Phoenix, Ariz.
Exhibits: Dr. Frank Holman, Boeing Airplane Co., 10708 39th Ave., S.W., Seattle 66, Wash.

May 8-10, 1961
National Aeronautical Electronics Conference, Miami and Biltmore Hotels, Dayton, Ohio.
Exhibits: Mr. Edward M. Lisowski, General Precision Lab., Inc., Suite 452, 333 West First St., Dayton 2, Ohio.

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.

Approximately actual size

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

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

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

Specifications

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

Data subject to change without notice.

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IRE News and Radio Notes

Current IRE Statistics

(As of May 31, 1960)

Membership—82,013
Sections*—107
Subsections*—26
Professional Groups*—28
Professional Group Chapters—269
Student Branches†—192

* See May, 1960, issue for a list.
† See June, 1960 issue for a list.

Calendar of Coming Events and Authors' Deadlines*

1960

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

Int'l. Information Theory Mtg., London, Eng., Aug. 29–Sept. 2.

EIA Conf. on Value Engrg., Disneyland Hotel, Anaheim, Calif., Sept. 7–8.

URSI 13th Gen. Assembly, Univ. of London, London, Eng., Sept. 5–15.

Joint Automatic Control Conf., M.I.T., Cambridge, Mass., Sept. 7–9.

Conf. on Communications, Roosevelt Hotel, Cedar Rapids, Iowa, Sept. 9–10.

4th Ann. Joint Mil. Ind. Electronic Test Equip. Symp., Chicago, Ill., Sept. 14–15.

8th Ann. Engrg. Management Conf. Morrison Hotel, Chicago, Ill., Sept. 15–16.

Benelux Section-PGCS Int'l Symp. on Data Transmission, Delft, Netherlands, Sept. 19–20.

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

Industrial Elec. Symp., Manager Hotel, Cleveland, Ohio, Sept. 21–22.

10th Ann. Broadcast Symp., Willard Hotel, Washington, D. C., Sept. 23–24.

Sixth Natl. Communications Symp., Hotel Utica and Utica Municipal Aud., Utica, N. Y., Oct. 3–5. (DL*: June 1, B. H. Balridge, 25 Bolton Rd., New Hartford, N. Y.)

PGNS 7th Ann. Mtg., Gatlinburg, Tenn., Oct. 3–5.

6th Conf. on Radio Interference Reduction, Chicago, Ill., Oct. 5–6.

Natl. Elec. Conf., Hotel Sherman, Chicago, Ill., Oct. 10–12. (DL*: May 1960 Prof. T. F. Jones, Jr., School of E.E., Purdue Univ., Lafayette, Ind.)

Engrg. Writing and Speech Symp., Bismark Hotel, Chicago, Ill., Oct. 13–14.

Symp. on Adaptive Control Systems, Garden City Hotel, Garden City, L. I., N. Y., Oct. 17–19, (DL*: June 31, 1960, H. Levenstein, W. L. Maxson Corp., 460 W. 34 St., N. Y.)

Symp. on Space Navigation, Deshler-Hilton Hotel, Columbus, Ohio, Oct. 19–21, (DL*: July 15, J. D. Kraus, Ohio State Univ. Radio Observatory, Columbus.)

* DL = Deadline for submitting abstracts.

(*Continued on page 15A*)

Call for Papers

1961 IRE International Convention

March 20–23, 1961

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

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

October 21, 1960

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 falls:

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

Note: Only original papers which have not been published or presented prior to the 1961 IRE International Convention will be considered; any necessary military or company clearance of paper is to be granted prior to submittal.

Address all material to: Dr. Gordon K. Teal, Chairman
1961 Technical Program Committee
The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

Purdue Announces Program for Symposium

Purdue University announces the program for the "Symposium on Engineering Applications of Probability and Random Function Theory," to be held at Purdue on November 15–16, 1960.

The speakers and their topics are as follows:

1) Dr. M. Kac, Cornell University, Professor of Mathematics, "Probability";
2) Dr. A. J. F. Siegert, Northwestern University, Professor of Physics, "Averages of Functionals for Various Processes";
3) Dr. R. B. Murphy, Bell Telephone Laboratories, Statistical Quality Control, "Reliability";
4) Dr. A. M. Freudenthal, Columbia University, Professor of Civil Engineering, "Structural Factors of Safety";
5) Dr. E. T. Jaynes, Stanford University, Associate Professor of Physics, "Application of Information Theory in Engineering";
6) Dr. I. Dyer, Bolt Beranek and Newman, Acoustical Consultant, "Noise Generation from Aerodynamic Sources";
7) Dr. R. Kalman, R.I.A.S., Research Mathematician, "New Methods in Wiener Filtering Theory";
8) Dr. E. Montroll, University of Maryland, Professor of Fluid Dynamics and Applied Mathematics, "Traffic Flow";
9) Dr. D. Slepian, Bell Telephone Laboratories, Research Mathematician, "Problems in Electrical Noise and Detection Theory."

Dr. S. S. Shu, Professor of Engineering Sciences, Purdue University, will introduce the program.

The purpose of the Symposium is to bring before the research engineer and scientist current techniques and thoughts concerning important practical applications of probability and random function theory. The unifying theme of the conference is the application of probabilistic models to engineering problems.

Requests for further information should be addressed to Professor J. L. Bogdanoff or Professor F. Kozin, co-chairmen of the Symposium, Division of Engineering Sciences, Purdue University, Lafayette, Ind.

Professional Group News

At its meeting on May 17, 1960 the IRE Executive Committee approved the following New Chapters: PG on **Audio**—North Carolina Chapter; PG on **Circuit Theory**—San Francisco Chapter; PG on **Engineering Management**—Binghamton Section; PG on **Instrumentation**—San Francisco Chapter.

Calendar of Coming Events and Author's Deadlines*

(Continued from page 14A)

East Coast Conf. on ANE, Lord Baltimore Hotel, Baltimore, Md., Oct. 24–26. (DL*: June 6, S. Hershfield, The Martin Co., Baltimore, Md.)

5th Ann. Conf. on Nonlinear Magnetics and Magnetic Amplifiers, Bellevue-Stratford Hotel Philadelphia, Pa., Oct. 26–28.

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

13th Ann. Conf. on Elec. Tech. in Med. and Bio., Sheraton Park Hotel, Washington, D. C., Oct. 31, Nov. 1–2.

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

Symp. on Communications, Queen Elizabeth Hotel, Montreal, Quebec Canada, Nov. 4–5.

6th Ann. Conf. on Magnetism and Magnetic Materials, New Yorker Hotel, N. Y., N. Y., Nov. 14–17. (DL*: Aug. 26. A. M. Clogston, R. C. Fletcher, Bell Tel. Labs., Murray Hill, N. J.)

Mid-Amer. Elec. Conv., Hotel Muehlebach, Kansas City, Mo., Nov. 15–16. (DL*: June 15, J. Austin, Bendix Aviation Corp., 95 and Troost, Kansas City, Mo.)

PGPT Ann. Conf., Boston, Mass., Nov. 15–16. (DL*: June 1, C. W. Watt, Raytheon Co., Waltham, Mass.)

Symp. on Engineering Applications of Probability and Random Function Theory, Purdue University, Lafayette, Ind., Nov. 15–16.

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

PGVC Ann. Mtg., Sheraton Hotel, Philadelphia, Pa., Dec. 1–2, (DL*: July 15, W. G. Chaney, American Telephone and Telegraph Co., 195 Broadway, N. Y. 7, N. Y.)

3rd EIA Conf. on Maintainability of Electronic Equipment, Hilton Hotel, San Antonio, Tex., Dec. 5–7.

Eastern Joint Computer Conf., New Yorker Hotel, New York, N. Y., Dec. 13–15. (DL, papers: Aug. 13, E. Kubie, Computer Usage Co., 18 E. 41 St., N. Y. 17, N. Y.)

1961

7th Natl. Symp. on Reliability and Quality Control, Bellevue-Strafford Hotel, Philadelphia, Pa., Jan 9–11 (DL*: May 8, 1960, R. E. Kuehn, IBM Corp., Owego, N. Y.) 1961.

Symp. on Space Instrumentation, Washington, D. C., Jan. 2nd PGMIL Conf., San Francisco, Calif., Feb. 1–3.

Int'l. Solid State Circuits Conf., University of Pennsylvania and Sheraton Hotel, Philadelphia, Pa., Feb. 15–17 (DL*: Oct. 21, 1960, J. J. Suran, Bldg. 3, Rm. 115 GE Co., Electronics Park, Syracuse, N.Y.)

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

SWIRECO, Dallas, Tex., April 19–21. 7th Region Tech. Conf. & Trade Show, Westward Ho Hotel, Phoenix, Ariz., April 26–28.

* DL = Deadline for submitting abstracts.

Tenth Annual Eastern Joint Computer Conference to be Held in New York City

The Annual Eastern Joint Computer Conference (EJCC) will be held December 13–15 at the Hotel New Yorker and Manhattan Center, New York City, it has been announced by N. Rochester of International Business Machines Corp., general chairman of the conference.

Abstracts of technical papers proposed for the conference should be submitted by August 13 to the technical program chairman, E. C. Kubie, Computer Usage Co., Inc., 18 East 41 Street, N. Y. 17, N. Y. Subject matter of the proposed papers should concern recent achievements or techniques in the design and use of computing equipment.

In an attempt to build a program of the highest possible quality, no parallel sessions are planned and a $300 prize will be awarded for the best presentation of a paper at the conference.

The EJCC is sponsored by the National Joint Computer Committee which consists of representatives of the Institute of Radio Engineers, the American Institute of Electrical Engineers, and the Association for Computing Machinery. The committee sponsors two meetings each year, one on the East, the other on the West Coast. More than 2500 members attended the 1960 WJCC in San Francisco this May. The 1960 EJCC will be the committee's eighth semi-annual conference.

In addition to N. Rochester and E. C. Kubie, members of the General Committee for the December meeting include: Publication Chairman, C. J. Rachel, Remington Rand Division of Sperry Rand Corp.; Local Arrangements Committee, B. W. Leavitt and A. L. Brown, General Telephone & Electronics Labs.; Finance, A. I. Schott, The National Cash Register Co.; Hotel, W. W. Ward, International Business Machines Corp.; Publicity and Printing, J. W. Heaney, Jr. and J. J. Lanigan, Sylvania Electric Products Inc.; Registration, J. Behr, Packard Bell Computer Corp.; Hospitality, R. P. Fopeano, Bendix Computer Division of Bendix Aviation Corp.; Exhibits, A. D. Meacham, Consultants to Industry Inc.; Exhibits Management, John Leslie Whitlock Associates.

Dr. C. N. Kimball (*center*), President of Midwest Research Institute and President of MAECON's Advisory Board, H. Stout (*left*), also of MRI and Chairman of the Kansas City Section of IRE, and L. Crissman (*right*) of Trans World Airlines and General Chairman of MAECON 1960, announce that plans have been finalized for the 1960 Mid-America Electronics Conference. MAECON 1960 technical papers sessions and exhibits will take place at the Hotel Muehlebach in Kansas City, Missouri, on November 15 16, 1960. The conference will focus as the theme, "The Semiconductor 60's." Other topics, presented concurrently with the semiconductor sessions, will feature microwaves, engineering education and management.

Space Electronics and Telemetry 1960 National Symposium

SHOREHAM HOTEL, WASHINGTON, D. C., SEPTEMBER 19–21, 1960

The Professional Group on Space Electronics and Telemetry will sponsor the fifth National Symposium concerning this important technological field at the Shoreham Hotel in the nation's capital on September 19–21. This year's Symposium will emphasize discussion of new design philosophies and advances in the state-of-the-art.

The program is organized around ten panel type sessions, each directed by a chairman prominent in the particular technical field. The panelists and the subject of their papers are selected by the panel chairman, thereby assuring a stimulating content and coherency to each panel area. It is planned to pre-publish the panelists' papers in the Symposium Proceedings and limit paper presentations to their highlights in order to have sufficient time for inter-panel discussion, questions and audience participation.

The program includes social events, such as the PGSET banquet, cocktail parties, luncheon, hospitality room and ladies' activities. There will also be a large and varied industrial exhibition.

Below is a listing of the panels being presented with respective subjects and chairmen. A complete listing of the individual panelists will be published in the next PROCEEDINGS. A brief summary indicates there are many well-known panelists on the program such as Dr. Lawrence Rauch, University of Michigan; Dr. Sonnett, Space Technology Laboratory; David Hogg, Bell Telephone Laboratory and Dr. Eberhardt Rechtin, Jet Propulsion Laboratory.

This year's Symposium is being sponsored by the Washington PGSET Chapter.

The Symposium committee is composed of:

- Symposium Chairman: J. E. Hinds, Jr., Ampex Corporation.
- Papers Chairman: H. W. Royce, Martin Company.
- Program Chairman: R. D. Briskman, NASA (Hq).
- Exhibits Chairman: L. H. King, Jansky and Bailey, Inc.
- Publicity Chairman: H. Gettings, Missiles and Rockets Magazine.
- Registration Chairman: R. W. Rochelle, NASA (GSFC).
- Arrangements Chairman: J. R. Kennedy, Ampex Corporation.
- Exhibits Manager: J. L. Whitlock, Whitlock Associates.
- Secretary-Treasurer: J. Joslow, Ampex Corporation.

Monday Morning September 19

Session I—Telemetry

Chairman: O. A. Hoberg, NASA, George Marshall Space Flight Center, Huntsville, Ala.

Monday Afternoon

PGSET Luncheon and Keynote Speech.

Session 2—Space Communications

Chairman: H. I. Butler, U. S. Army Signal Corps Engineering Laboratory, Fort Monmouth, N. J.

Session 3—Electronic Propulsion

Jointly sponsored by the American Rocket Society and the Institute of Radio Engineers.

Chairman: R. N. Edwards, General Electric, Cincinnati, O.

Tuesday Morning, September 20

Session 4—Signal Conditioning

Chairman: G. Ludwig, State University of Iowa, Iowa City.

Session 5—Navigation of Vehicles in Space

Chairman: Dr. W. E. Frye, Lockheed Missile System Div., Palo Alto, Calif.

Session 6—Effects of Space Environment on Men and Equipment

Chairman: Major General D. Flickinger, Headquarters Advanced Research and Development Command, Washington, D. C.

Session 7—Navigation on Earth by Means of Satellites

Chairman: W. C. Schoolfield, Advanced Research Projects Agency, Washington, D. C.

Wednesday Morning, September 21

Session 8—Propagation

Chairman: J. H. Chisholm, Massachusetts Institute of Technology, Lincoln Lab., Lexington, Mass.

Session 9—Tracking

Chairman: F. B. Smith, Headquarters, NASA, Washington, D. C.

Session 10—Guidance

Chairman: J. M. Bridges, Office of Secretary of Defense, Washington, D. C.

1960 Western Electronics Show and Convention

SPORTS ARENA, LOS ANGELES, CALIF., AUGUST 23–26

Tuesday Morning, August 23

Session 1—Systems and Maintainability

Contributed Papers

Chairman: *R. Whiteman, Project Director, General Analysis Corp., Los Angeles, Calif.*

"A Systematic Approach to Complex Electronic Equipment Maintenance," *J. J. Brown, J. H. Chin, G. W. Jacob, Sperry Gyroscope Co., Great Neck, L. I.*

"Economy Models for System Design Engineers," *E. S. Winlund, General Electric Co., Phoenix, Ariz.*

"Precision Film Potentiometers," *H. Adise, Computer Instruments Corp., Hempstead, L. I., N. Y.*

"Engineering Contribution to Product Quality," *W. C. Kraft, Sandia Corp., Albuquerque, N. M.*

Session 2—Pulse-Handling Techniques

Contributed Papers

Chairman: *N. Begovich, Hughes Aircraft Co., Fullerton, Calif.*

"A Theory of Enhancement Filters," *A. Norris, Varian Associates, Palo Alto, Calif.*

"Pulsed RF Storage in Long Delay, Broadband Closed Loop Systems," *O. A. Huettner, International Telephone and Telegraph Laboratories, Nutley, N. J.*

"The Problems and Solutions in the Navy's Program for Standardization of Video Processing and Distribution," *L. T. Rhodes, Naval Research Laboratories, Washington, D. C.*

"A Solid-State Video Processor with Pulse-for-Pulse AGC," *R. E. Segal, Packard-Bell Electronics Corp., West Los Angeles, Calif.*

Session 3—Communications: New Solutions to Some Old Problems

Contributed Papers

Chairman: *C. Lindholm, RAND Corporation, Santa Monica, Calif.*

"Effect of Link Elimination in Data Transmission Systems," *A. Machi and J. Hoffman, System Development Corp., Lodi, N. J.*

"Optimum Antenna Pattern for a Signal Burst Communication System," *P. A. Lux, Sandia Corp., Livermore, Calif.; H. M. Swarm and D. D. McNelis, Univ. of Washington, Seattle, Wash.*

"Linear Cancellation Technique for Suppressing Impulse Noise," *E. J. Baghdady, Research Laboratory of Electronics, Massachusetts Institute of Technology, Cambridge, Mass.*

Session 4—Management of Manned Machine Systems

Symposium

Chairman: *A. Small, Hughes Aircraft Company, Fullerton, Calif.*

"A Systems Management Appraisal of the Functions of Human Engineering," *T. Eason, Stromberg-Carlson, Co., Rochester, N. Y.*

"Human Factors Contribution to Management Control Procedures," *S. Deutsch, Douglas Aircraft Co., Inc., Santa Monica, Calif.*

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V	BL-236	54,000–57,000	10	UG385/U
V	BL-237	57,000–60,000	10	UG385/U
V	BL-221	69,000–70,500	10	UG385/U

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Session 5—Semiconductor Devices and Tubes

Contributed Papers

Chairman: *N. J. Golden, Hoffman Semiconductors, Inc., El Monte, Calif.*

"Power Output and Efficiency of Thermionic Converters," *I. T. Saldi, General Electric Co., Schenectady, N. Y.*

"High Power at 1000 MC Using Semiconductor Devices," *G. Leuttgenau and N. V. Duffin, Pacific Semiconductors, Inc., Culver City, Calif.*

"Equivalent Circuit of a Parametric Diode at Microwaves," *A. K. Kamal, K. E. Lytal and H. W. Pass, Purdue University, Lafayette, Ind.*

"Quality Assurance Procedures for Power Transistors," *J. S. Schaffner, Delco Radio Division, General Motors Corp., Kokomo, Ind.*

Tuesday Afternoon

Session 6—What are the Communication Values of the Technical Symposium?

Panel Discussion

Chairman and Moderator: *L. McConnell, System Development Corp., Santa Monica, Calif.*

"The Speaker," *I. J. Fong, Remington Rand Corp., UNIVAC Div., St. Paul, Minn.*

"The Writer," *E. R. Hagemann, Space Technology Laboratories, Los Angeles, Calif.*

"The Editor," *N. Horgan, The RAND Corp., Santa Monica, Calif.*

"The Publisher," *W. G. Stone, John Wiley & Sons, Inc., New York, N. Y.*

Session 7—Varactors and Tunnel Diode Applications

Contributed Papers

Chairman: *G. C. Messenger, Hughes Semiconductor Div., Newport Beach, Calif.*

"A Nonlinear Capacitor Harmonic Generator Suitable for Space Vehicle Applications," *P. M. Fitzgerald, T. H. Lee, M. S. Moy, E. J. Powers and J. J. Younger, Lockheed Aircraft Corp., Missile Systems Div., Sunnyvale, Calif.*

"Parametric Radio Frequency Amplifier," *A. Szerlip, Packard-Bell Electronic Corp., West Los Angeles, Calif.*

"Gain and Bandwidth Inconsistencies in Low Frequency Reactance Up-Converter Parametric Amplifiers," *A. K. Kamal and A. J. Holub, Purdue Univ., Lafayette, Ind.*

"A Compact Tunnel Diode Amplifier for Ultra High Frequencies," *G. Schaffner, Semiconductor Products Div., Motorola, Inc., Phoenix, Ariz.*

"Analysis and Design of the Twin-Tunnel-Diode Logic Circuit," *C. H. Alford, Lockheed Aircraft Corp., Missile Systems Div., Sunnyvale, Calif.*

Session 8—Instrumentation

Contributed Papers

Chairman: *A. Kaufman, Litton Industries, Beverly Hills, Calif.*

"Widely Separated Clocks with Microsecond Synchronization and Independent Distribution Systems," *T. L. Davis and R. H. Doherty, U. S. Dept. of Commerce, National Bureau of Standards, Boulder, Colo.*

"The Synthesis of Instrument Compensating Networks," *R. W. Kearns, Wayne State University, Detroit, Mich.*

"An Automatic Servomechanism Response Plotter," *D. Rice, Republic Aviation Corp., Farmingdale, L.I., N. Y.*

"Touch Detector," *G. T. Kemp, Texas Research Associates Corp., Austin, Tex.*

"Determination of Instantaneous Speed Error Data," *A. Updike, Ampex Data Products Co., Redwood City, Calif.*

Session 9—Circuit Theory

Tutorial Papers with Panel

Chairman: *L. Weinberg, Hughes Research Laboratories, Malibu, Calif.*

Panelists: *I. M. Horowitz, Hughes Research Laboratories, Malibu, Calif.; J. R. Burnett, Space Technology Laboratories, Los Angeles, Calif.*

"Analysis and Design of Feedback Systems with Gain and Time Constant Variations," *K. Chen, Westinghouse Electric Corp., Pittsburgh, Pa.*

"Measures of Sensitivity for Linear Systems with Large Multiple Parameter Variations," *S. L. Hakimi and J. B. Cruz, University of Illinois, Urbana, Ill.*

"A Sampled Data Technique for Realizing Network Transfer Functions," *L. E. Franks and I. W. Sandberg, Bell Telephone Laboratories, Inc., Murray Hill, N. J.*

"Delay Distortion Correction for Networks and Filters," *T. R. O'Meara, Hughes Research Laboratories, Malibu, Calif.*

Session 10—Semiconductor Devices

Contributed Papers

Chairman: *T. W. Griswold, Continental Device Corp., Hawthorne, Calif.*

"A New Semiconductor Memory Element with Non-Destructive Readout and Electrostatic Storage," *V. H. Grinich and David Hilbiber, Fairchild Semiconductor Corp., Mountain View, Calif.*

"Some Device Aspects of Multiple Microwave Reflections in Semiconductors," *H. Jacobs, F. A. Brand, J. Meindl and M. Benanti, U. S. Signal Army Research & Development Labs., Ft. Monmouth, N. J.; R. Benjamin, Monmouth College, W. Long Branch, N. J.*

"Base Turn-Off of PN-PN Switches," *R. H. Van Ligten and D. Navon, Transitron Electronic Corp., Wakefield, Mass.*

"Novel Adder-Subtractor Circuit Utilizing Tunnel Diodes," *R. A. Kaenel, Bell Telephone Labs., Inc., Murray Hill, N. J.*

"Transistor Scaling Theory," *W. E. Roach, Pacific Semiconductors, Inc., Culver, City, Calif.*

Wednesday Morning, August 24

Session 11—Computers—General

Contributed Papers

Chairman: *L. J. Craig, The RAND Corp., Santa Monica, Calif.*

"Digital Control Techniques for Space," *L. F. Jones and P. Margolin, Westinghouse Electric Corp., Baltimore, Md.*

"The Polymorphic Principle in Data Processing," *H. A. Heit, Thompson Ramo Wooldridge, Inc., Canoga Park, Calif.*

"An Aided Adaptive Character Reader for Machine Translation of Languages," *P. Baran and G. Estrin, University of California, Los Angeles, Calif.*

"A Multi-Addressable Random Access File System," *E. Coil, Librascope Div., General Precision, Inc., Glendale, Calif.*

Session 12—Stereo Multiplex Broadcasting

Papers and Panel Discussion

Chairman: *I. J. Kaar, Hoffman Electronics Corp., Los Angeles, Calif.*

Panelists: *C. Eilers, Zenith Radio Corp., Chicago, Ill.; W. H. Beaubien, General Electric Co., Utica, N. Y.; M. G. Crosby, Crosby-Teletronics Corp., Syosset, N. Y.; H. Parker, Calbest Engineering and Electronics, Los Angeles, Calif.; W. Halstead, Multiplex Development Corp., New York, N. Y.*

"Requirements for FM Stereophonic Radio Transmission," *R. J. Farber, Hazeltine Research Corp., Plainview, N. Y.*

"Progress of Field Tests for FM Stereophonic Broadcast Systems," *A. P. Walker, National Association of Broadcasters, Washington, D.C.*

Session 13—Microwave Theory and Techniques—I: Passive Elements

Contributed Papers

Chairman: *H. Saltzman, Kearfott Co., Inc., Van Nuys, Calif.*

"Misconceptions about Equivalent Circuits for Periodic Microwave Structures," *R. M. Bevensee, Varian Associates, Palo Alto, Calif.*

"A Fast Switching X-Band Circulator Utilizing Ferrite Toroids," *L. Levey and L. M. Silber, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

"Broadband Electronically-Tuned Microwave Filters," *K. L. Kotzebue, Watkins-Johnson Co., Palo Alto, Calif.*

"The Observed 50–90 KMC Attenuation of Two Inch Improved Waveguide," *A. P. King, Bell Telephone Laboratories, Red Bank, N. J.*

"A Noncontracting, Broadband Rotary Joint, and Four-Way Switch," *D. Alstadter and N. A. Dawson, Melpar, Inc., Falls Church, Va.*

Session 14—Analysis of Manned Machine Systems

Symposium

Chairman: *G. F. Rabideau, Norair Division of Northrop Corporation, Hawthorne, Calif.*

"The Vocal Adaptive Controller—Human Pilot Dynamics and Opinion," *D. T. McRuer and I. L. Ashkenas, Systems Technology, Inc., Inglewood, Calif.*

"An Analysis of the Decision Making Functions of a Simulated Air Defense," *A. Sweetland and W. Haythorn, The RAND Corp., Santa Monica, Calif.*

"Methodology of Manned Machine System Analysis," *R. W. Queal, Boeing Airplane Co., Seattle, Wash.*

"Optimizing Linear Dynamics for Human Operated Systems by Minimizing the Mean Square Tracking Error," *T. E. Leonard, Aeronutronic Systems, Inc., Newport Beach, Calif.*

"Encoding Techniques for Visual Displays in Computer-Aided Systems," *K. M. Neuman, U. S. Naval Electronics Lab., San Diego, Calif.*

Session 15—Microwave Tubes

Tutorial Papers

Chairman: *W. H. Christoffers, Microwave Tube Div., Hughes Aircraft Corp., Los Angeles, Calif.*

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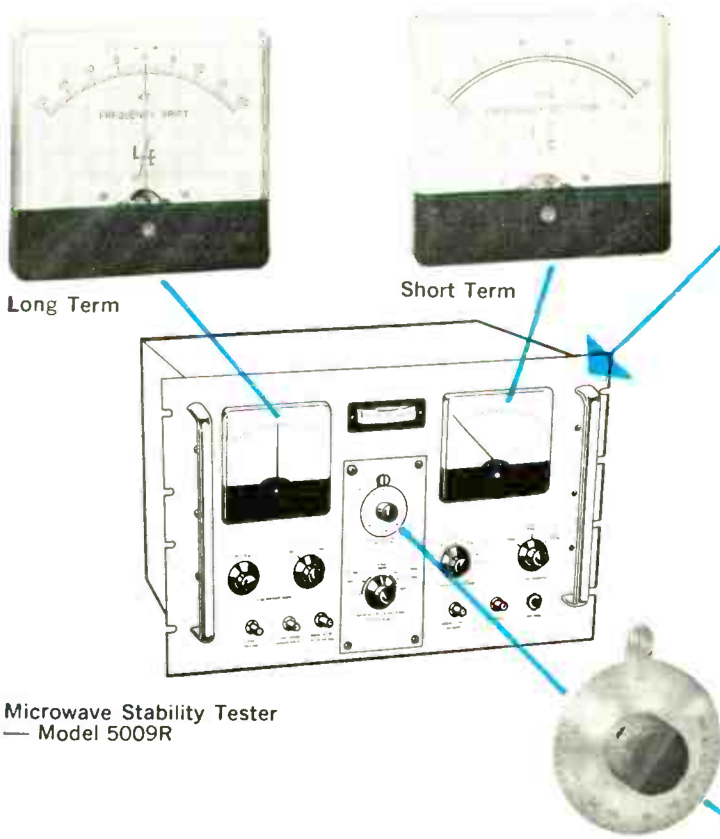

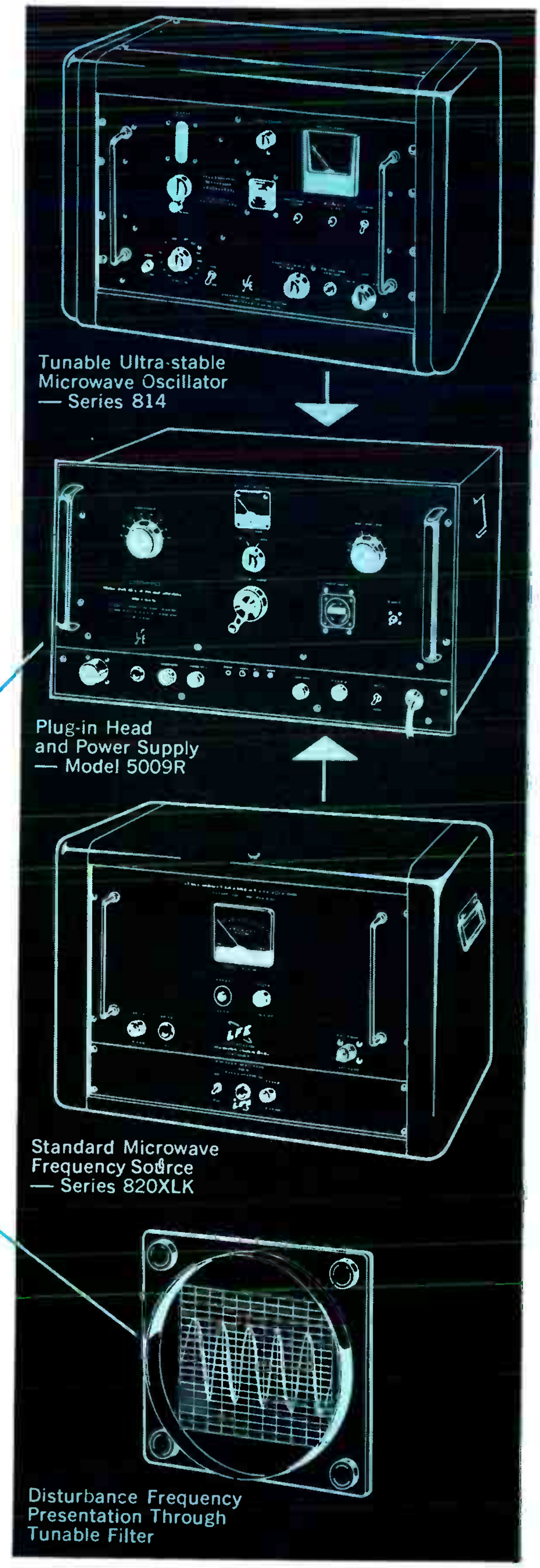

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"An Octave-Bandwidth Ultra Low Noise Traveling Wave Amplifier," *E. W. Kinaman and G. E. St. John, Watkins-Johnson Co., Palo Alto, Calif.*

"Very High Convergence Electron Guns," *D. V. Geppert, Sylvania Electronic Systems, Mountain View, Calif.*

"Cooling of the Slow Space-Charge Wave of an Electron Beam with Application to the Traveling-Wave Tube," *D. C. Forster, Hughes Research Labs., Culver City, Calif.*

"Arc Discharge Microwave Switch Tube," *S. J. Tetenbaum, R. R. Moats and D. Campbell, Sylvania Electronic Systems, Mountain View, Calif.*

"A Periodically Focused Backward-Wave Oscillator," *C. C. Johnson, Hughes Research Labs., Culver City, Calif.*

"A Four-Cavity, Electrostatically Focused, Ku-Band Klystron Amplifier," *R. G. Rockwell, Varian Asssociates, Palo Alto, Calif.*

Wednesday Afternoon

Session 16—Computer Circuits and Devices

Contributed Papers

Chairman: *G. Eisler, Eisler Associates, Los Angeles, Calif.*

"Diodeless Magnetic Core Logic," *S. B. Yochelson, Goodyear Aircraft Corp., Akron, Ohio.*

"A Fractional Microsecond Cycle Time Memory Using Low Coercive Ferrite Cores," *A. Lemack and J. E. Thomas, Sylvania Electronic Systems, Needham, Mass.*

"Adaptive Switching Circuits," *B. Widrow and M. E. Hoff, Stanford University, Palo Alto, Calif.*

"25-mc Clock-Rate Computer Circuits for Operation from −20°C to +100°C," *C. R. Cook, Jr., Texas Intruments, Inc., Dallas, Tex.*

"A Dynamic Logic Technique for Sixteen Megacycle Clock Rate," *T. P. Bothwell, J. DeClue, H. H. Hill and J. R. Longland, Computer Control Co., Framingham, Mass.*

Session 17—Magnetic Data Recording

Tutorial

Chairman: *W. R. Isom, Radio Corporation of America, Camden, N. J.*

"Extending the Bandwidth of a Conventional Instrumentation Recording System," *A. M. Wilson, Precision Instrument Co., San Carlos, Calif.*

"A Wideband Magnetic Recording System," *M. E. Anderson and J. A. Granath, Armour Research Foundation, Chicago, Ill.*

"The Sensitivity of Reproducing Heads in High-Frequency Magnetic Recording Systems," *W. T. Frost, Ampex, Data Products Co., Redwood City, Calif.*

"Mechanical Design of the cm-100 Instrumentation Tape Recorder," *J. T. Mullin, Mincom Div., Minnesota Mining and Mfg. Co., Los Angeles, Calif.*

"Electrical Design and Performance of the cm-100 Instrumentation Tape Recorder," *G. N. Johnson, Mincom Div., Minnesota Mining & Mfg. Co., Los Angeles, Calif.*

"Comparison of Wideband FM and Carrier Erase Techniques for Recording Data from DC to 10 kc," *G. Work and D. Lewis, Leach Corp., Compton, Calif.*

Session 18—Microwave Theory and Techniques—II: Active Elements

Contributed Papers

Chairman: *R. Jamison, Hughes Aircraft Co., Culver City, Calif.*

"Masers for System Applications," *H. R. Senf, Hughes Research Laboratories, Culver City, Calif.*

"Design and Operation of an S-Band Traveling-Wave Diode Parametric Amplifier," *C. G. Shafer, Raytheon Co., Waltham, Mass.*

"The Noise Figure of Iterative Traveling Wave Parametric Amplifiers," *C. V. Bell, Walla Walla College, Walla Walla, Wash.; and Glen Wade, Raytheon Co., Burlington, Mass.*

"Theory of TEM Diode Switching," *R. V. Garver, Diamond Ordnance Fuze Laboratories, Washington, D. C.*

"Tunnel Diode Microwave Oscillators with Milliwatt Power Outputs," *D. E. Nelson and F. Sterzer, Radio Corporation of America, Princeton, N. J.*

Session 19—Working with Engineers

Invited Speakers

Chairman: *N. H. Moore, Litton Industries, San Carlos, Calif.*

"Marketing," *G. P. Beiging, Packard-Bell Electronic Corp., West Los Angeles, Calif.*

"Patent Law," *W. R. Lane, North American Aviation, Los Angeles, Calif.*

"Accounting and Finance," *R. T. Silberman, Electronics Capital Corp., San Diego, Calif.*

Session 20—Vehicular Communications—I: Radiating Systems

Contributed Papers

Chairman: *D. L. MacDonald, Pacific Telephone & Telegraph, Los Angeles, Calif.*

"Theory and Performance of Vehicular Center-Fed Whip Antenna," *H. Brueckmann, U. S. Army Signal Research and Development Laboratory, Ft. Monmouth, N. J.*

"A Broad-Band 160 Megacycle Colinear Array," *R. F. H. Yang and H. H. Hansen, Andrew Corp., Chicago, Ill.*

"Effects of Tower and Guys on Performance of Side-Mounted Vertical Antennas," *R. F. H. Yang and F. R. Willis, Andrew Corp., Chicago, Ill.*

"Foamflex Coaxial Cable for Communications," *J. Arbuthnott, A. L. McKean and S. Trill, Phelps Dodge Copper Products Corp., New York, N. Y.*

Thursday Morning, August 25

Session 21—Component and Systems Reliability

Panel Discussion Following Presentation of Paper

Chairman: *W. R. Kuzmin, Packard-Bell Electronics Corp., Los Angeles, Calif.*

Panelists: *S. Gollin, Walter Darwin Teague Associates, New York, N. Y.; S. Kukawka, Bourne Laboratory, Inc., Riverside, Calif.; A. Wood, Relay Div., Leach Corp., Los Angeles, Calif.; C. C. Elrod, The Ralph M. Marsons Co., Pasadena, Calif.*

"Using Failure Rate Data for Component Part Derating," *I. Doshay, Aerojet General Corp., Azusa, Calif.*

Session 22—Air Traffic Control (ATC)—Session I

Related Papers

Chairman: *V. Weihe, General Precision, Inc., Washington, D. C.*

"Operational Consideration in ATC Design," *R. F. Link, Bureau of Research and Development, Federal Aviation Agency, Washington, D.C.*

"An Airline Pilot Looks at ATC," *Capt. J. D. Smith, Air Line Pilots Association, New York, N. Y.*

"ATC from the Aircraft Owners Viewpoint," *V. H. Kayne, Aircraft Owners and Pilots Association, Washington, D. C.*

"The Airlines and Air Traffic Control," *J. R. Dettman, Air Transport Association of America, Los Angeles, Calif.*

Session 23—Antennas, Session I

Contributed Papers

Chairman: *L. L. Bailin, Hughes Aircraft Co., Culver City, Calif.*

"A New Approach to Antenna Beam-Shaping—The 'Coke-Bottle' Antenna," *C. C. Phillips, Melpar, Inc., Falls Church, Va.*

"Application of Frequency Scan to Circular Arrays," *P. Shelton, Aero Geo Astro Corp., Alexanderia, Va.*

"Low Sidelobe Interferometer Antenna Patterns," *H. Pfizenmeyer and J. A. Kuecken, Avco Corp., Cincinnati, Ohio.*

"Design Techniques for a Light Weight, High Power, Spiral Antenna," *L. P. Jones, P. E. Taylor and C. W. Morrow, Melpar, Inc., Falls Church, Va.*

"Phase Distribution of Spiral Antennas," *N. Barbano, Sylvania Electronic Systems, Mountain View, Calif.*

Session 24—Synthesis and Design of Manned Machine Systems

Symposium

Chairman: *Col. L. Baker, U. S. Army, Chief Psychologist, Aberdeen, Md.*

"Human Factors in the Establishment of System Design Requirements," *R. H. Schneider, Dunlap and Associates, Inc., Santa Monica, Calif.*

"The Human Factors Laboratory as System Design Tool," *F. Marzocco, Thompson Ramo Wooldridge, Inc., Canoga Park, Calif.*

"On the Effect of CRT Transfer Function on Detection Threshold," *C. W. Miller and W. R. Minty, Cornell Aeronautical Laboratory, Inc., Buffalo, N. Y.*

"Introduction to Teaching Machines," *S. Levine, Litton Industries, Beverly Hills, Calif.*

"A High-Speed Color Display Unit," *W. H. Huntley, Jr., Stanford University, Stanford, Calif.*

Session 25—Microminiaturization

Tutorial Papers

Chairman: *T. Liiamatainen, Diamond Ordnance Fuze Laboratory, Washington, D. C.*

"Design and Fabrication of a Microelectronic IF Amplifier," *J. R. Black, Motorola Corp., Phoenix, Ariz.*

"A Packaged Micromodule Laboratory for Industry," *D. T. Levy, Radio Corporation of America, Somerville, N. J.*

"Semiconductor Packaging for High Component Density Application," *G. P. Walker, Rheem Semiconductors, Inc., Palo Alto, Calif.*

"Surface Passivation as Applied to Micro-Components," *T. C. Hall, Pacific Semiconductors, Inc., Culver City, Calif.*

"Laminar Junction Structures: A New Concept in Microcircuitry," *J. Alegreti, Merck Sharpe & Dohme, Rahway, N. J.*

"Solid State Micrologic Elements," *L. Kattner, J. Last, and J. Nall, Fairchild Semiconductor Corp., Palo Alto, Calif.*

Thursday Afternoon

Session 26—Government and Industry: Engineering Proposals

Panel Discussion

Moderator: *Cmdr. W. Ten Hagen, USN, Bureau of Weapons, Western District, El Segundo, Calif.*

Panelists: *J. Tassen, Contracts Div., Bureau of Naval Weapons, Washington, D. C.; C. E. Petrillo, U. S. Army Signal R & D Lab., Ft. Monmouth, N. J.; J. B. Lewi, Packard-Bell Electronics Corp., Los Angeles, Calif.; N. Klumph, Western Development Labs., Philco Corp., Palo Alto, Calif.; R. Nordlund, Wright Air Development Div., Dayton, Ohio.*

Session 27—Air Traffic Control (ATC)—Session II

Related Papers

Chairman: *G. Bieging, Packard-Bell Electronics Corp., Los Angeles, Calif.*

"Central Data Processing of ATC Systems," *L. L. Wahlman, Librascope Div., General Precision, Inc., Glendale, Calif.*

"Data Processing Requirements of the ATC System," *N. Pomerantz, General Precision Laboratories Div., General Precision Inc., Pleasantville, N. Y.*

"Automation in ATC," *T. L. Bartlett, Radio Corporation of America, Camden, N. J.*

"The Need for Automatic ATC," *H. K. Morgan, Bendix Aviation Corp., Detroit, Mich.*

"Future Trends in ATC," *G. Van Alstyne Gilfillan Bros., Inc., Los Angeles, Calif.*

Session 28—Antennas, Session II

Contributed Papers

Chairman: *C. E. Dunn, Convair Div. of General Dynamics, Inc., Pomona, Calif.*

"A Continuous Bistatic Echo Area Range," *J. W. Eberle, Ohio State Univ., Columbus, Ohio.*

"Fresnel Region Boresight Methods," *A. Bogush, RCA, Moorestown, N. J.*

"The Zone Plate as a Focussing Element," *C. E. Hendirix and L. F. Van Buskirk, U. S. Naval Ordnance Test Station, China Lake, Calif.*

"Beacon Antennas for Project Mercury," *D. F. Shea, D. Alstadter and W. O. Puro, Melpar, Inc., Falls Church, Va.*

"Miniaturized Cavity Fed Slot Antennas," *F. P. Brownell and D. F. Kendall, The Martin Co., Denver, Colo.*

Session 29—The Pioneer V Experiments

Symposium

Chairman: *C. P. Sonett, Space Technology Laboratories, Inc., Los Angeles, Calif.*

"Preliminary Results from the Space Probe Pioneer V," *C. Y. Fan, P. Meyer and J. A. Simpson, University of Chicago, Chicago, Ill.*

"Radiation Measurements Made by Space Probe Pioneer V," *R. L. Arnoldy, R. A. Hoffman and J. R. Winckler, University of Minnesota, Minneapolis, Minn.*

"Measurements of the Geomagnetic and Interplanetary Magnetic Fields: Pioneer V," *P. J. Coleman, D. L. Judge, E. J. Smith and C. P. Sonett, Space Technology Laboratories, Inc., Los Angeles, Calif.*

"Determination of the Astronomical Unit from a Least Square Fit to the Orbit of Pioneer V," *J. B. McGuire, D. D. Morrison and L. Wong, Space Technology Laboratories, Inc., Los Angeles, Calif.*

Session 30—Microminiaturization

Panel Discussion

Moderator: *W. V. Wright, Electro Optical Systems, Inc., Pasadena, Calif.*

Panelists: *W. B. Warren, Hughes Semiconductors Labs., Newport Beach, Calif.; M. Kahn, Sprague Electronics, North Adams, Mass.; J. S. Kilby, Texas Instruments, Inc., Dallas, Texas; D. Mackey, Radio Corporation of America, Somerville, N. J.; H. C. Lin, Westinghouse Electric Corp., Pittsburgh, Pa.; G. J. Selvin, Sylvania Electric Products, Inc., Waltham, Mass.; E. E. Maiden, Pacific Semiconductors, Inc., Culver City, Calif.; R. Norman, Fairchild Semiconductor Corp., Palo Alto, Calif.*

"Reliability of Superminiaturized Silicon Diodes," *E. E. Maiden, Pacific Semiconductors, Inc., Culver City, Calif.*

"The Hughes Type I Microelectronic Circuit Concept," *W. B. Warren, Hughes Semiconductor Labs., Newport Beach, Calif.*

Friday Morning, August 26

Session 31—Seeking a Logical Bioinstrumentation System

Panel Discussion

Chairman: *V. W. Blockley, Consultant: Environment Physiology, Santa Monica, Calif.*

Moderator: *M. Fishbein, System Development Corp., Santa Monica, Calif.*

Panelists: *D. Douglas, Spacelabs, Inc., Van Nuys, Calif.; L. Fields, Starling Corporation, Los Angeles, Calif.; T. McNeely, North American Aviation, Los Angeles, Calif.; M. McLennon, Chief of Medical Electronics—Bio-Medical Laboratory, Wright Air Development Center, Dayton, Ohio.*

"The Anesthetized Individual in a Normal Environment," *J. B. Dillon, M.D., University of California, Los Angeles, Calif.*

"The Unhealthy, Conscious Individual in a Normal Environment," *T. Winsor, M.D., Los Angeles, Calif.*

"The Healthy, Conscious Individual in an Abnormal Environment," *P. Meehan, M.D., University of Southern California, Los Angeles, Calif.*

"Computers and Programming in a Bioinstrumentation System," *P. Tiffany, System Development Corp., Santa Monica, Calif.*

Session 32—Military Electronics

Contributed Papers

Chairman: *Lt. Col. R. Isenson, Office Deputy Commander Army, Pacific Missile Range, Pt. Mugu, Calif.*

"System Implications of Electronic Ancestor Worship," *B. H. Baldridge, General Electric Co., Utica, N. Y.*

"Implementation of a Modern Communication System on National and Global Scales," *C. K. Chappuis, Los Angele s Calif.*

"Automatic Programming of Ground Support Checkout Equipment Using Computer Techniques," *M. Cook and C. Keeler, Convair Astronautics, San Diego, Calif.*

"The BMEWS Automatic Monitoring System," *E. L. Danheiser and M. Korsen, Radio Corporation of America, Moorestown, N. J.*

Session 33—Information Theory and Modulation Methods

Symposium

"PTM/AM," *C. Hoeppner, Radiation, Inc., Melbourne, Fla.*

"PCM/FM," *R. L. Sink, Consolidated Electro Dynamics, Pasadena, Calif.*

"PDM," *K. Uglow, Electromechanical Research, Inc., Orlando, Fla.*

"PACM/FM," *M. B. Rudin, Aeronutronic Systems, Inc., Newport Beach, Calif.*

"DSSB/AM," *J. W. Halina, International Telephone and Telegraph Co., Nutley, N. J.*

"Digilock," *R. Sanders, Space Electronics Corp., Glendale, Calif.*

"Sebit 25," *J. L. Hollis, Rixon Electronics, Silver Spring, Md.*

"Telebit," *J. Taber, Space Technology Laboratories, Inc., Los Angeles, Calif.*

Session 34—Operation and Training of Manned Machine Systems

Symposium

Chairman: *H. M. Parsons, System Development Corp., Santa Monica, Calif.*

"Model for Automating Maintenance Function," *D. Ellis, Hughes Aircraft Co., Culver City, Calif.*

"A Model for Relating Human Factors to ADP Systems Performance," *J. B. Teeple, Thompson Ramo Wooldridge, Sierra Vista, Ariz.*

"Human Maintenance Functions in Man-Machines," *M. Grodsky and G. W. Levy, The Martin Company, Baltimore, Md.*

"Human Factors in System Operations and Training," *J. W. Singleton, System Development Corp., Santa Monica, Calif.*

"Measuring Human Interactions in Man/Machine Systems," *A. M. Freed, System Development Corp., Santa Monica, Calif.*

Session 35—Vehicular Communications II: Mobile Radio and Paging System

Contributed Papers and Panel

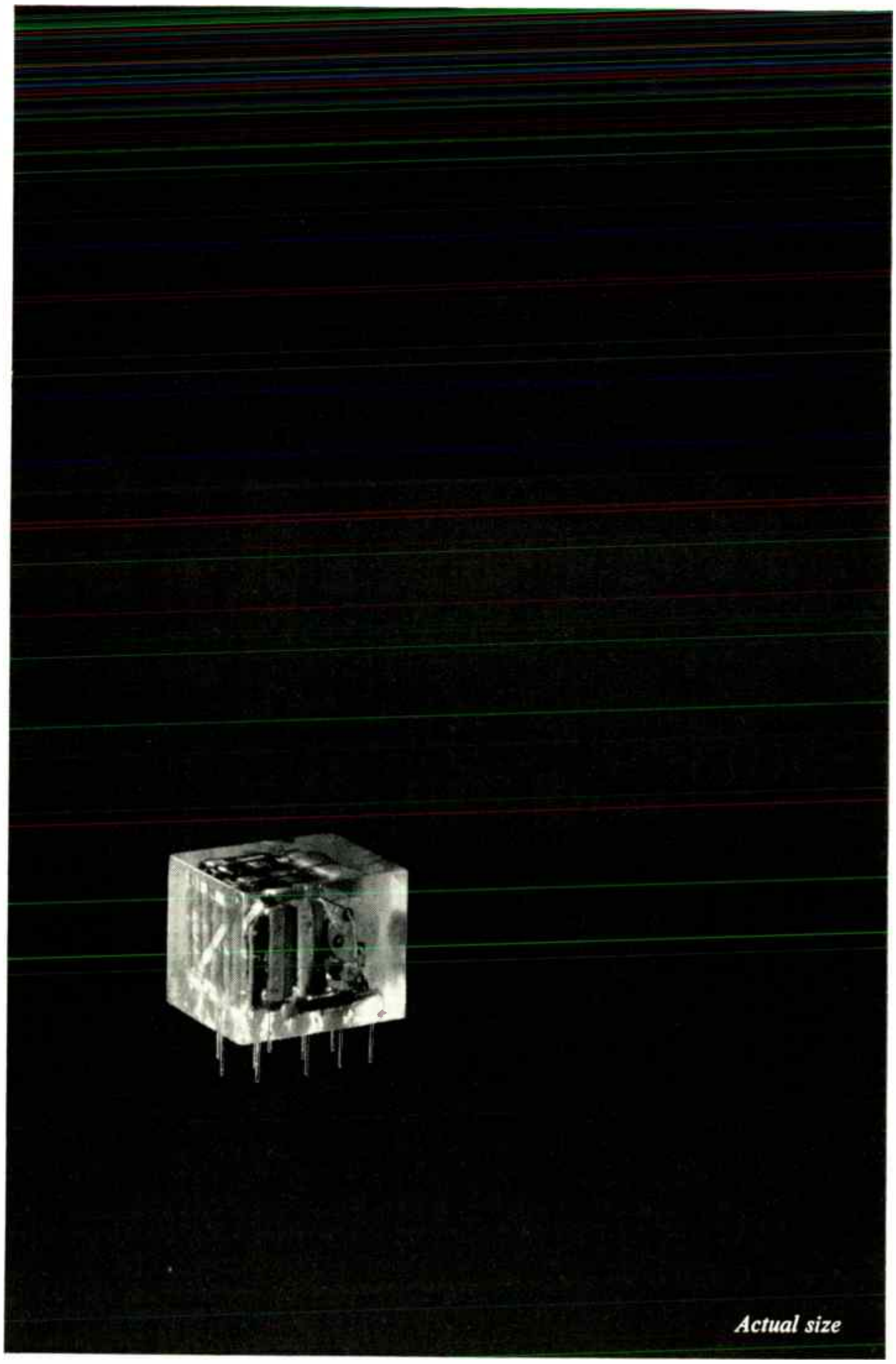
Chairman and Moderator: *K. T. Corner, Comm. Dept., City of Los Angeles, Calif.*

"System Performance, Compatibility and Standards," *R. T. Buesing and N. H. Sheperd, General Electric Co., Lynchburg, Va.*

"Personal Two Way Radio Communication System Featuring Modular Construction," *T. H. Yaffe, Bendix Radio Div., Bendix Aviation Corp., Baltimore, Md.*

"Personal Radio Paging in the VHF Band," *J. F. Mitchell, Motorola, Inc., Chicago, Ill.*

"Police and Fire Department Communication Centers: A System Approach to the Control Console and the Related Facilities," *G. A. Brookes, Westrex Corp., Los Angeles, Calif.*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Friday Afternoon

Session 36—Continuation of Session 31

Panel Discussion

Session 37—Coding Methods and Telemetry

Contributed Papers

Chairman: *A. V. Balakrishnan, Space Technology Laboratories, Inc., Los Angeles, Calif.*

"An Improved FM Discriminator Detector for Airborne Telemetry Receivers," *G. E. Reis and C. E. Land, Sandia Corp., Albuquerque, N. M.*

"Improved DOVAP Transponder," *F. M. Gardner, Gardner Research Co., Orange, Calif.*

"Optimized Data Systems," *J. C. O'Brien, Technical Specialist, Pomona, Calif.*

"Reliable Fail-Safe Binary Communication," *J. J. Metzner and K. C. Morgan, Research Div., New York University, New York, N. Y.*

"Data Compression," *H. Schwab, Applied Development Corp., Hawthorne, Calif.*

Session 38—Continuation of Session 33

Symposium

Session 39—Not scheduled

Session 40—Vehicular Communications III: New Ideas and Concepts for Mobile Telephone Operation

Contributed Papers and Panel

Chairman and Moderator: *A Culbertson, Lenkurt Corp., San Carlos, Calif.*

Panelists: *R. T. Crabb, Mobilfone Corp., Los Angeles, Calif.; A. R. Ogilvie, Secode Corp., San Francisco, Calif.; C. W. Schweiger Pacific Telephone and Telegraph Co., San Diego, Calif.*

"Application of Trunking Principles to Multichannel Mobile Telephone Service," *E. S. Randel, American Telephone and Telegraph Co., New York, N. Y.*

"System Concepts for Address Communication Systems," *D. H. Hamsher, U. S. Army Signal R & D Laboratories, Ft. Monmouth, N. J.*

"Push-Button Mobile Dial Radiotelephone: An Advanced Concept in Common Carrier Mobile Service," *J. R. Stewart, Motorola, Inc., Chicago, Ill.*

"A Three-Channel Single Sideband Multiplexed FM Mobile Radio System Using Transistorized Vehicle Termina Equipment," *W. S. Felch, American Telephone and Telegraph Co., New York, N. Y.*

"Guarde Tone Signalling," *W. B. Smith, Bendix Radio, Div. of Bendix Aviation Corp., Baltimore, Md.*

Tuesday Afternoon, August 23

Workshop I—Management of Manned Machine Systems

Roundtable Discussion with Panel

Moderator: *R. L. Clark, Department of Defense, Washington, D. C.*

Panelists: *R. Gilson, Stromberg-Carlson Co., San Diego, Calif.; E. Speakman, Radio Corporation of America, Camden, N. J.; W. Duke, Space Technology Laboratories, Los Angeles, Calif.; F. Seufert, Hoffman Electronics Corp., Los Angeles, Calif.* (This is a continuation of Session 4)

Wednesday Afternoon, August 24

Workshop II—Analysis of Manned Machine Systems

Roundtable Discussion with Panel

Moderator: *Lt. Col. A. Debbons, Rome Air Development Division, Rome, N. Y.*

Panelists: *L. Blumstein, Cornell Aeronautical Laboratory, Buffalo, N. Y.; L. Seale, Bell Aircraft Corp., Buffalo, N. Y.; M. Adelson, Hughes Aircraft Co., Fullerton, Calif.*; a fourth panelist to be announced. (This is a continuation of Session 14)

Thursday Afternoon, August 25

Workshop III—Synthesis and Design of Manned Machine Systems

Roundtable Discussion with Panel

Moderator: *D. T. McRuer, Systems Technology, Inc., Los Angeles, Calif.*

Panelists: *R. K. Ausbourne, Hughes Aircraft Corp., Culver City, Calif.; W. Evans, Aeronutronic Systems, Inc., Newport Beach, Calif.; L. Christie, System Development Corp., Santa Monica, Calif.; H. Van Cott, International Business Machines Corp., Bethesda, Md.* (This is a continuation of Session 24)

Friday Afternoon, August 26

Workshop IV—Operation and Training of Manned Machine Systems

Roundtable Discussion with Panel

Moderator: *J. Lyman, University of California, Los Angeles, Calif.*

Panelists: *J. Bialek, Stanford Research Institute, Palo Alto, Calif.; J. Mautsch, System Development Corp., Santa Monica, Calif.; L. Stovanoff, Hoffman Electronics Corp., Los Angeles, Calif.*; a fourth panelist to be announced.

Wednesday Afternoon, August 24

Women's Session

Special Session—Engineering: The Woman's Role

Invited Speakers with Audience Participation

Chairman: *N. E. Porter, Hewlett-Packard Co., Palo Alto, Calif.*

"The Woman's Position in Engineering," *R. M. Bernstein, Douglas Aircraft Co., Inc., Los Angeles, Calif.*

"Debugging the Engineer," *B. B. Leitner, Santa Monica, Calif.*

Other speakers to be announced.

The Fourth London Symposium on Information Theory

ROYAL INSTITUTION, LONDON, W.1, AUGUST 29–SEPTEMBER 2, 1960

The Fourth London Symposium on Information Theory is to be held at the Royal Institution in London, W.1, from Monday, August 29 to Friday, September 2. The Symposium is to consist mainly of discussions; it is expected that copies of all papers to be read will have been sent to all participants approximately six weeks before the meeting. Authors will then devote only 10 minutes each to introducing their papers.

The Proceedings will be published afterwards, by Butterworths' Scientific Press, London.

Monday Morning, August 29

Coding and Detection Theory and Statistical Theory

Opening remarks, explanation of plans.

"A Linear Method of Construction of Error-Correcting Codes," *M. Driml, The Academy of Information Theory and Automation, Czechoslovakia.*

"Digitalised Communication over Two-Way Channels," *J. A. Wozencraft and M. Horstein, M.I.T.*

"Error-Correcting Codes from Linear Sequential Networks," *N. M. Abramson, Stanford University.*

Monday Afternoon

"Entropy and Metric Spaces," *C. Rajski, Institute of Mathematics, Poland.*

"Physical Entropy, Information and the General Statistical Theory of Estimation," *B. Mandelbrot, I.B.M. Research Center.*

"On Decoding Group Codes," *N. Zierler, M.I.T.*

"A Self-Optimising System of Coding," *A. M. Andrew, National Physical Laboratory, Great Britain.*

Tuesday Morning, August 30

Telecommunication Systems

"Congestion in Telephone Exchanges," *R. Syski, Hivac Ltd., Great Britain.*

"Communication in Digital Systems," *I. L. Lebow, M.I.T.*

"Optimum Receivers for the Detection of Signals Transmitted through a Random Channel," *T. Kailath, M.I.T.*

"Weight of Evidence and False-Target Probabilities," *J. Good, Admiralty Research Laboratories, Great Britain.*

Tuesday Afternoon

Human Reaction to Information

"Choice Reaction-Time Experiments and Information Theory," *J. A. Leonard, M.R.C., Applied Psych. Laboratories, Great Britain.*

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"Is Human Choice Reaction-Time a Constant?" *G. H. Mowbray, Johns Hopkins University.*

"Human Performance in Perceptual Tasks," *J. F. Schouten, Inst. for Perceptual Studies, Netherlands.*

"Hesitation and Information in Speech," *F. Goldman-Eisler, University College, London, Great Britain.*

Wednesday Morning, August 31

Sensory Information and Biological Models

"Sequential Observations by Human Observers of Signals in Noise," *J. A. Swets and D. M. Green, M.I.T.*

"Visual Information Storage," *G. Sperling and E. Averbach, Bell Telephone Labs.*

"Binocular Depth Perception and Pattern Recognition," *B. Julesz, Bell Telephone Labs.*

"A Decision Theory Approach to Sound Lateralization," *H. B. Voelcker, Imperial College, London.*

Wednesday Afternoon

"Activity in Networks of Neuron-like Elements," *G. B. Farley and W. A. Clark, M.I.T.*

"Neural Discharge Patterns and Simulation of Synaptic Operations in Transmission of Sensory Information," *L. J. Viernstein and R. G. Grossman, John Hopkins University.*

"A Model for Neurophysiological Functions (an Improved 'Tortoise')," *H. Zemanek, The Technical College, Vienna, Austria.*

A paper not yet selected, possibly from the Soviet Union.

Thursday Morning, September 1

Learning Mechanisms and other Artefacts

"A Model for Cursive Writing Viewed as a Process of Pattern Generation," *M. Eden and M. Halle, M.I.T.*

"Machine-reading of Handwriting," *L. D. Harmon and L. S. Frishkopf, Bell Telephone Labs.*

"Studies of a Model of Human Categorizing Behavior," *J. C. R. Licklider, Bolt Beranek and Newman.*

"Adaptive Waveform Recognition," *C. V. Jakowatz, R. L. Shuey, and G. M. White, G.E. Co.*

Thursday Afternoon

"Task Simplification and Learning Devices," *J. Hartmansis, G.E. Co.*

"Learning in Random Nets," *M. Minsky and O. G. Selfridge, M.I.T.*

"A Learning Filter," *D. Gabor, Imperial College, London.*

"Mathematical Models of Perceptual Learning," *S. Papert, National Physical Laboratory, London.*

Friday Morning, September 2

Classification Theory, Syntactics and Semantics

"Basic Principles and Technical Variations in Sentence-Structure Determination," *D. G. Hays, RAND Corporation.*

"An Experimental Study of Strategies in 'Hypothesis-Formation' by Computer," *M. Kochen, I.B.M.*

"A Mathematical Theory of Discrete Classification," *S. W. Golomb, Cal. Inst. Tech.*

"Documentary Classification as a Self-Organizing System," *R. A. Fairthorne, Royal Aircraft Establishment, Great Britain.*

Friday Afternoon

"The Information Content of Biological Classifications," *G. A. Maccacaro and A. Rescignio, Milan, Italy.*

"Predictive Syntactic Analysis," *M. E. Sherry and A. G. Oettinger, Harvard University.*

"The Description of Finite Sequential Processes," *K. E. Iverson, Harvard University.*

"The Informational Analysis of Questions and Commands," *D. M. Mackay, King's College, London.*

Joint Automatic Control Conference

KRESGE AUDITORIUM, MASSACHUSETTS INSTITUTE OF TECHNOLOGY, CAMBRIDGE, MASS., SEPTEMBER 7–9, 1960

Conceived to reduce the overlap of conferences on control, sponsored by individual societies, the annual Joint Automatic Control Conference provides a unique high level conference on the theory and application of automatic control. This year's sponsoring society, The American Society of Mechanical Engineers, with the cooperation of the Boston Section, and the participating societies, IRE, American Institute of Electrical Engineers, Instrument Society of America, and American Institute of Chemical Engineers will conduct no other national conferences on control, except as part of their general meetings.

Only numbered papers in this program will be available in separate copy form. Copies may be obtained from: ASME Order Department, 29 West 39 St., New York 18, N. Y. Prices are 50¢ to members of ASME, AIEE, AIChE, ISA, and IRE, $1.00 to nonmembers. Papers must be ordered by the paper number listed in this program; all numbered papers will be available at the Conference. Although no conference proceedings will be published, each participating society will publish its transaction-quality papers.

Committees for the Conference are as follows: General Chairman: *W. D. Archibald, Energy Control Co., New York, N. Y.;* Program: *J. M. Mozley, Johns Hopkins Hospital, Baltimore, Md.;* Publicity: *W. E. Vannah, Control Engineering, New York, N. Y.;* Local Arrangements: *A. A. Erickson, Jr., R. E. Erickson Co., Boston, Mass.;* MIT Arrangements: *J. L. Shearer, MIT, Cambridge, Mass.*

Steering Committee: (ASME) *W. D. Archibald, Energy Control Co., New York, N. Y.;* (AIChE) *D. M. Boyd, Universal Oil Products Co., Des Plaines, Ill.;* (ASME) *E. R. Behn, Arma, Garden City, N. Y.;* (IRE/PGAC) *D. P. Lindorff, University of Connecticut, Storrs, Conn.;* (ISA) *R. K. Adams, Oak Ridge National Laboratory, Oak Ridge, Tenn.*

Program Committee: General Chairman: *J. M. Mozley, Johns Hopkins Hospital, Baltimore, Md.;* AIChE Program Chairman: *Dr. W. H. Abrahams, E. I. du Pont de Nemours & Co., Wilmington, Del.;* AIEE Feedback Controls Committee Program Chairman: *Dr. A. Fuchs, Boonshaft and Fuchs, Inc., Hatboro, Pa.;* AIEE Recording & Controlling Instrumentation Committee Program Chairman: *E. P. Davis, Leeds & Northrup Co., North Wales, Pa.;* IRE Program Chairman: *H. A. Miller, Taylor Instrument Co., Rochester, N. Y.,* ISA Program Chairman: *J. L. Harned, Research Labs., General Motors, Corp., Warren, Mich.;* ASME Program Chairman: *R. E. Kalman, R.I.A.S., Baltimore, Md.*

General Information

Registration

Registration will be conducted in Kresge Auditorium during the following hours:

Tuesday,	September 6—4:00 P.M.– 6:00 P.M.
Wednesday,	September 7—8:00 A.M.– 5:00 P.M.
Thursday,	September 8—8:00 A.M.– 5:00 P.M.
Friday,	September 9—8:00 A.M.–12:00 NOON

Personnel of conference registration committee and of chamber of commerce will be on hand at registration desk to advise on tours, shows, museums and restaurants.

Conference Fees

Members of ASME, AIChE, AIEE, ISA, IRE	$10.00
Nonmembers	$15.00
Ladies	No Charge

Conference Highlights

September 6
- 4:00 P.M.-6:00 P.M.—Early Bird Registration—Kresge Auditorium
- 5:00 P.M.-7:00 P.M.—Feedforward Cocktail Party—Foyer of Campus Room-Graduate House

September 8
- 5:30 P.M.—Feedback Dinner—New England Lobster Clam Buffet—Rockwell Cage. Ticket: $6.00, includes tax and gratuities.

Authors' Breakfast

By invitation, the authors, session chairmen and vice chairmen of each day's technical sessions will meet at breakfast to become better acquainted and receive last minute instructions. The breakfast will be served in the West Dining Room, Graduate House, at 7:30 A.M. each day of the Conference.

Ladies' Program

September 7–9
- 9:00 A.M.-10:00 A.M.—Coffee Hour—West Dining Room—Graduate House

Off Campus Accommodations

Those who have not already secured accommodations may obtain their accommodations in Cambridge (*e.g.*, Hotel Commodore) or Boston (*e.g.*, Sherry Biltmore or Statler Hilton). Breakfast and luncheon will be available in the main dining room of the Graduate House, MIT, (cafeteria style). No general dinner arrangements other than the Feedback Buffet Dinner, September 8, 1960, 5:30 P.M. to 7:30 P.M., have been planned, but Boston boasts fine eateries to satisfy any palate.

Wednesday Morning, September 7

Optimal Switching

Chairman: *R. E. Kalman, RIAS, Baltimore, Md.*

Vice-Chairman: *N. H. Choksy, Applied Physics Lab., Johns Hopkins University, Silver Spring, Md.*

"Dynamic Synthesis of Higher Order Saturating Systems,"[1] (ASME Paper No. 60-JAC-2) *F. Kurzweil, Jr., General Products Div., San Jose, Calif.*

"Solution Space Approach to the Design of Optimal Control Systems,"[1] (ASME Paper No. 60-JAC-11) *Yu-Chi-Ho, Computation Lab., Harvard University, Cambridge, Mass.*

"The Optimum Response of Second-Order Zero Seeking Velocity Controlled Systems with Contactor Controls,"[1] (ASME Paper No. 60-JAC-3), *I. Flugge-Lotz and Mih Yin, Div. of Engineering Mechanics, Stanford University, Stanford, Calif.*

"Pulse-Width Relay Control in Sampling Systems,"[1] (ASME Paper No. 60-JAC-4), *W. L. Nelson, Dept. of Elec. Engrg.*, Columbia University, New York, N. Y.

Chemical Process Dynamics

Chairman: *L. M. Zoss, Valparaiso University, Valparaiso, Ind.*

Vice-Chairman: *A. S. Foss, E. I. duPont de Nemours and Co., Wilmington, Del.*

"Survey of the Literature on Heat Exchanger Dynamics and Control," (AIChE Paper No. 1) *T. J. Williams and H. J. Morris, Monsanto Chemical Co., St. Louis, Mo.*

"The Dynamics and Control of Distillation Units and Other Mass Transfer Equipment," (AIChE Paper No. 2) *R. R. Rothfus and D. H. Archer, Carnegie Institute of Technology, Pittsburgh, Pa.*

"Dynamics of Chemical Reactors," (AIChE Paper No. 3) *L. Lapidus, Dept. of Chemical Engrg., Princeton University, Princeton, N. J.*

General Papers I

Chairman: *R. Knudsen, Engineering Staff, General Motors Technical Center, Warren, Mich.*

Vice-Chairman: *J. L. Harned, Senior Research Engineer, Research Labs., General Motors Technical Center, Warrren, Mich.*

"Optimization of Chemical Processes, (Paper No. ISA-1-60) *A. E. Hoerl and C. R. Hall, E. I. duPont de Nemours and Co., Newark, Del.*

"Accurate Pressure Regulation by Digital Servo System," (Paper No. ISA-5-60) *O. K. Kowallis, Wiancko Engineering Co., Pasadena, Calif.*

"Fractionation Control by Chromatography," (Paper No. ISA-7-60) *H. J. Maier, Perkin-Elmer Corp., Norwalk, Conn.*

"Application Factors in Transmitter Design," (Paper No. ISA-8-60) *V. V. Tivy, P. H. Drinker and M. C. Kessel, Foxboro Co., Foxboro, Mass.*

Wednesday Afternoon

New Techniques in Control System Theory

Chairman: *J. E. Gibson, Dept. of Elec. Engrg., Purdue University, Lafayette, Ind.*

Vice-Chairman: *B. Friedland, Dept. of Elec. Engrg., Columbia University, New York, N. Y.*

"Design of Optimum Multiveriable Control Systems,"[1] (ASME Paper No. 60-JAC-5) *E. B. Lee, Military Products Group, Minneapolis-Honeywell Regulator Co., Minneapolis, Minn.*

"Reduction of Dimensionality and the Dynamic Programming Treatment of Control Processes,"[1] (ASME Paper No. 60-JAC-6) *R. Bellman and R. Kalaba, The RAND Corp., Santa Monica, Calif.*

"Kinetic Lypaunov Functions for Stability Analysis of Nonlinear Control Systems,"[1] (ASME Paper No. 60-JAC-7) *S. S. L. Chang, Dept. of Elec. Engrg., New York University, New York, N. Y.*

"New Results in Linear Filtering and Prediction Theory,"[1] (ASME Paper No. 60-JAC-12) *R. E. Kalman, RIAS, Baltimore, Md. and R. S. Bucy, Applied Physics Lab., Johns Hopkins University, Silver Spring, Md.*

Dynamic Testing of Components and Systems

Chairman: *J. P. Lienesch, The Foxboro, Co., Foxboro, Mass.*

Vice-Chairman: *J. J. Hamrick, Burroughs Corp., Paoli, Pa.*

"Basic Survey of Methods Available for Dynamic Testing of Components and Systems," (AIEE Paper No. CP-60-964) *L. A. Gould and R. W. Rasche, Dept. of Elec. Engrg., MIT, Cambridge, Mass.*

"Dynamic Testing of Industrial Systems," (AIEE Paper No. CP-60-965) *A. R. Catheron, The Foxboro Co., Foxboro, Mass.*

"A Simulation Facility for the Study of Decision Making in Complex Military Systems," (AIEE Paper No. CP-60-966) *J. M. Doughty, Cambridge Research Center, U. S. Air Force, Bedford, Mass.*

"Dynamic Mechanical Measurements in Computer Systems," (AIEE Paper No. CP-60-967) *A. J. Fuimarello, IBM, Poughkeepsie, N. Y.*

"Dynamic Impedance Measurement of Electrical Contacts," (AIEE Paper No. CP-60-968) *E. S. Mathison, IBM, Poughkeepsie, N. Y.*

General Papers II

Chairman: *J. L. Harned, Senior Research Engineer, Research Labs., General Motors Technical Center, Warren, Mich.*

Vice-Chairman: *L. Taylor, Assistant Director of Research, Vickers, Inc., Detroit, Mich.*

"High Temperature Pneumatics—Its Use and Control," (Paper No. ISA-2-60) *J. Rivard and J. Pembleton, Vickers, Inc., Detroit, Mich.*

"High Performance Pneumatic Controllers," (Paper No. ISA-13-60) *F. J. Finegan, Jr., Sperry Gyroscope Co., Great Neck, Long Island, N. Y.*

"A New Technique for the Simulation of Transport Lags," (Paper No. ISA-6-60) *R. K. Sterns, Computer Systems, Inc., New York, N. Y.*

"Physical Implementation of Torque-Saturated, Second-Order Linear Servos With Optimum Switching Schemes," (Paper No. ISA-9-60) *R. M. Howe and L. L. Rauch, University of Michigan, Ann Arbor, Mich.*

Thursday Morning, September 8

Adaptive Control

Chairman: *K. Goff, Chief, Simulation and Analysis Group, Leeds and Northrup Co., North Wales, Pa.*

Vice-Chairman: *J. Schwartzenberg, Simulation and Analysis Group, Leeds and Northrup Co., North Wales, Pa.*

"Optimalizing Control with Process-Dynamics Identification," (IRE Paper No. 60AC-15) *P. Eykoff and O. J. M. Smith, Dept. of Engrg., University of California, Berkeley, Calif.*

"An Error Criterion for Adaptive Systems," (IRE Paper No. 60AC-13) *R. Van Wechel, Hallamore Electronics Co., Anaheim, Calif.*

"Adaptive Control through Sinusoidal Response," (IRE Paper No. 60AC-14) *K. C. Smith, Boeing Airplane Co., Seattle, Wash.*

Selected Papers on Automatic Control

Chairman: *T. F. Mahoney, Section Manager, Surveillance Systems Dept., Raytheon Company Equipment Division, Sudbury, Mass.*

Vice-Chairman: *R. E. Claflin, Jr., President, Claflin Associates, Newtonville, Mass.*

"Synchronous Networks," (IRE Paper No. 60AC-5) *G. Weiss, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.*

"The Application of Feedback Control Techniques to Organizational Systems," (IRE Paper No. 60AC-6) *R. B. Wilcox, Missile Electronics and Controls Division, RCA, Burlington, Mass.*

"A Simulation Study of Semi-Automatic Air Traffic Control Systems," (IRE Paper No. 60AC-9) *A. S. Jackson, S. Pardee and H. Ottoson, Data Processing and Controls Department, Thompson-Ramo-Wooldridge Products Co.*

"Control Concepts for Nuclear Ramjet Reactors," (IRE Paper No. 60AC-7) *R. E. Finnigan, University of California, Lawrence Radiation Lab., Livermore, Calif.*

"Stability and Control of Nuclear Rocket Propulsion," (IRE Paper No. 60AC-8) *R. R. Mohler, N-4 Group, University of California, Los Alamos, N. M.*

Control Components I

Chairman: *W. E. Sollecito, GE Co., Schenectady, N. Y.*

Vice-Chairman: *P. Troutman, GE Co., Schenectady, N. Y.*

"PERCOS—Performance Coding System of Methods and Devices Used for Measurement and Control," (AIEE Paper No. CP-60-972) *E. Mittlemann, Consulting Engineer, Chicago, Ill.*

"A-C Tachometer Specifications," (AIEE Paper No. CP-60-973) *D. Bloser, Daystrom Transicoil Co., Worchester, Pa.*

[1] Paper not available.

"Dynamic Response Testing, (AIEE Paper No. CP-60-974) *P. S. Buckley, E. I. duPont de Nemours and Co., Wilmington, Del.*

Thursday Afternoon

Synthesis and Programming of Digital Computer Control Systems

Chairman: *J. R. Ragazzini, Dean of Engineering, New York University, New York, N. Y.*

Vice-Chairman: *A. S. Robinson, Head, Advanced Electronics Lab., Bendix Corp., Teterboro, N. J.*

"Control Programming—Key to the Synthesis of Efficient Digital Computer Control Systems," (AIEE Paper No. TP-60-969) *A. S. Robinson, Head, Advanced Electronics Lab., Bendix Corp., Teterboro, N. J.*

"Information Handling Efficiency of a Digital Control Computer," (AIEE Paper No. TP-60-970) *H. Freeman, Sperry Gyroscope Co., Great Neck, Long Island, N. Y.*

"Logical Organization of the Honeywell D-290," (AIEE Paper No. TP-60-971) *J. J. Eachus, Datamatic Div., Minneapolis-Honeywell Regulator Co., Boston, Mass.*

Control Components II

Chairman: *A. R. Aikman, Schlumberger Oil Well Surveying Corp., Ridgefield, Conn.*

Vice-Chairman: *R. M. Howe, Dept. of Aeronautical and Astronautical Engineering, University of Michigan, Ann Arbor, Mich.*

"Standard Servo Package System," (Paper No. ISA-4-60) *R. Spencer, Vickers, Inc., Detroit, Mich.*

"Analysis of Electrohydraulic Valves and Systems," (Paper No. ISA-11-60) *R. C. Cataldo, Research Labs., General Motors Technical Center, Warren, Mich.*

"The Dynamics of Common Magnetically Damped Instruments," (Paper No. ISA-12-60) *M. M. Gibbs, Minneapolis-Honeywell Regulator Co., Boston Div., Boston, Mass.*

General Papers III

Chairman: *J. L. Shearer, Dept. of Mechanica Engineering, MIT, Cambridge, Mass.*

Vice-Chairman: *F. D. Ezekiel, Dept. of Mechanical Engineering, MIT, Cambridge, Mass.*

"Approximate Method for Calculating the Time Response in Linear Time-Varying and Nonlinear Automatic Control Systems," (ASME Paper No. 60-JAC-10) *B. Naumov, Institute of Automatics and Telemechanics, Moscow, U.S.S.R.* (To be presented by *G. Newton, Dept. of Elec. Engrg., MIT, Cambridge, Mass.*

"An Approach to the Design of Power Servo-mechanisms," (ASME Paper No. 60-JAC-1) *D. V. Stallard, Feedback Controls, Natick, Mass.*

"Improvement of the Power Efficiency of a Hydraulic Control System by the Use of a Gain Compensated Control Valve," (ASME Paper No. 60-JAC-8) *S. Y. Lee, Dept. of Mechanical Engrg., MIT, Cambridge, Mass.*

Thursday Evening

Report on the First IFAC Congress at Moscow, U.S.S.R. June 27–July 2, 1960

Chairman: *H. Chestnut, First President of IFAC, Control Systems Engineer, General Engrg. Lab., General Electric Co., Schenectady, N. Y.*

At this evening meeting, a team of outstanding automatic control specialists will report on technical developments observed first-hand at the First IFAC Congress at Moscow, U.S.S.R. These reports will include an analysis of the present status and future trends in the areas of automatic control theory, scientific and industrial applications, and control system components. A period of open discussion will follow presentation of the formal report.

Friday Morning, September 9

Special Topics

Chairman: *W. Van der Velde, Dept. of Aeronautics and Astronautics, MIT, Cambridge, Mass.*

Vice-Chairman: *H. Mori, Hydel, Inc., Waltham, Mass.*

"Integral Transforms for Algebraic Analysis and Design of a Class of Linear-Variable and Adaptive Control Systems," (IRE Paper No. 60AC-10) *G. W. Johnson, Advanced Systems Research, IBM, Oswego, N. Y.*

"Regression Techniques in Multivariate Adaptive Control Systems," (IRE Paper No. 60AC-11) *A. B. Bishop, Dept. of Industrial Engrg., Ohio State University, Columbus, O. and H. R. Chope, Industrial Nucleonics Corp., Columbus, O.*

"Optimum Design of Passive-Adaptive, Linear Feedback System with Varying plants," (IRE Paper No. 60AC-12) *P. E. Fleischer, Elec. Engrg. Dept., New York University, New York, N. Y.*

"Investigation of Periodic Modes of a Sampled Data Control System Containing a Saturating Element," (ASME Paper No. 60-JAC-9) *W. E. Meserve and H. C. Torng, Dept. of Elec. Engrg., Cornell University, Ithaca, N. Y.*

Case Histories and Computers for On Line Control Systems

Chairman: *R. G. Lex, Leeds and Northrup Co., North Wales, Pa.*

Vice-Chairman: *H. R. Koen, Minneapolis-Honeywell Regulator Co., Industrial Div., Philadelphia, Pa.*

"Programming for Process Control," (AIEE Paper No. CP-60-975) *E. Borgers, Thompson-Ramo-Wooldridge Products Co. Beverly Hills, Calif.*

"Computer Control System for a Continuous Annealing Line," (AIEE Paper No. CP-60-976) *J. T. Bradford, Jr., Jones and Laughlin Co., Pittsburgh, Pa. and R. W. Kirkland, General Electric Co., Schenectady, N. Y.*

"Process Control Computer System for Vinyl Chloride Manufacturing at B. F. Goodrich Chemical Company, Calvert City, Ky.," (AIEE Paper No. CPA-60-5041) *H. Flum, Thompson-Ramo Wooldridge Products Co., Beverly Hills, Calif.*

"Hybrid Computers for Process Control," (AIEE Paper No. CP-60-978) *G. Birkel, Jr., Radiation, Inc., Melbourne, Fla.*

Panel Discussion—Automatic Control Education

"The Role of the University in Control Technology," a panel discussion by four educators and a representative from industry on these questions: a) Is there such a thing as an academically prepared control engineer? b) What would be the ideal academic preparation for this field? c) What kind of graduate control specialists does industry want? d) What kind of research in control technology can universities undertake? e) What kind of aid can industry give to universities to help develop competence in control technology?

Moderator: *L. E. Slater, Executive Director, Foundation for Instrumentation Education and Research, New York, N. Y.*

Panelists: *D. P. Eckman, Director, Systems Research Center, Case Institute of Technology, Cleveland, Ohio.; E. F. Johnson, Professor of Chemical Engineering, Princeton University, Princeton, N. J.; J. L. Shearer, Associate Professor of Mechanical Engineering, MIT, Cambridge, Mass.; S. W. Herwals, Vice-President, Research, Westinghouse Electric Corp., Pittsburgh, Pa.; J. R. Ragazzini, Dean of Engineering, New York University, New York, N. Y.*

Friday Afternoon

Nonlinear Systems

Chairman: *G. S. Axelby, Fellow Engineer, Westinghouse Air Arm Division, Baltimore, Md.*

Vice-Chairman: *H. A. Miller, Manager, Electronic Development Div., Taylor Instrument Co., Rochester, N. Y.*

"On Minimum of Maximum Expected Deviation from an Unstable Equilibrium Position of a Randomly Perturbed Control System," (IRE Paper No. 60AC-1) *M. Aoki, Numerical Analysis Research, University of California, Los Angeles, Calif.*

"A Nonlinear Analysis Technique for an On-Off Servo," (IRE Paper No. 60AC-2) *R. Farrah, Res. Labs. Div., Bendix Corp., Detroit, Mich.*

"Synthesis of High Order Nonlinear Control Systems with Ramp Input," (IRE Paper No. 60AC-3) *C. Shen, Dept. of Mech. Engrg., Rensselaer Polytechnic Institute, Troy, N. Y.*

"A Phase Space Investigation of Bi-Stable Systems by Means of Vectors," (IRE Paper No. 60AC-4) *R. V. Halstenberg, Convair, San Diego, Calif.*

Optimum Control of Chemical Processes

Chairman: *T. J. Williams, Monsanto Chemical Company, St. Louis, Mo.*

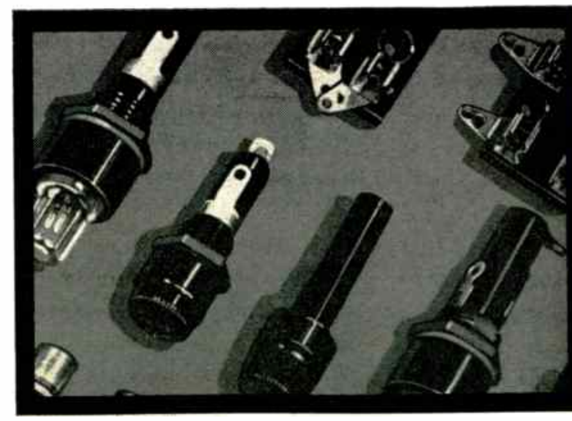
Vice-Chairman: *A. M. Fuchs, Boonshaft and Fuchs, Hatboro, Pa.*

"Dynamic Solution to a Generalized Chemical Processing Model," (AIEE Paper No. CP-60-979) *R. E. Boydston, Information Systems, Inc., Skokie, Ill.*

"Considerations in the Design of a Dynamic Control System for Generalized Chemical Processing Model Considered as a Nonlinear System," (AIEE Paper No. CP-60-980) *A. E. Beecher and L. A. Gould, Systems Lab., MIT, Cambridge, Mass.*

"Optimization of a Chemical Processing System," (AIEE Paper No. CP-60-981) *J. H. Decanini and L. A. Gould, Systems Lab., MIT, Cambridge, Mass.*

"Dynamic Optimization and Control of a Stirred Tank Chemical Reactor," (AIEE Paper No. TP-60-982) *W. Kipiniak and L. A. Gould, Systems Lab., MIT, Cambridge, Mass.*



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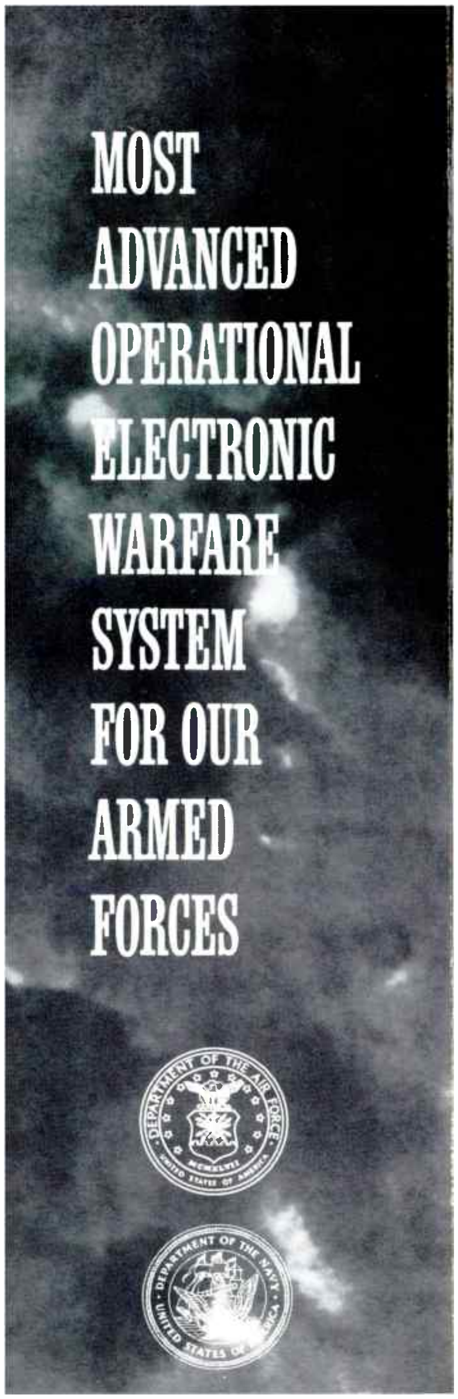
"Big Stick" of the Strategic Air Command, Boeing B-52-G Missile Bomber can reach overseas targets and return without refueling.

The most potent airborne Countermeasures Equipments yet devised are being produced in quantity by Hallicrafters for our military aircraft. Teamwork with our armed forces continues to pay off for America–in more effective, more economical electronic warfare systems.

Please turn the page . . . for examples of Hallicrafters facilities, and Quick Reaction Capability in Airborne, Ground, Naval and Missile electronic systems for counter-measures . . . reconnaissance . . . maintenance and technical support . . . communications.

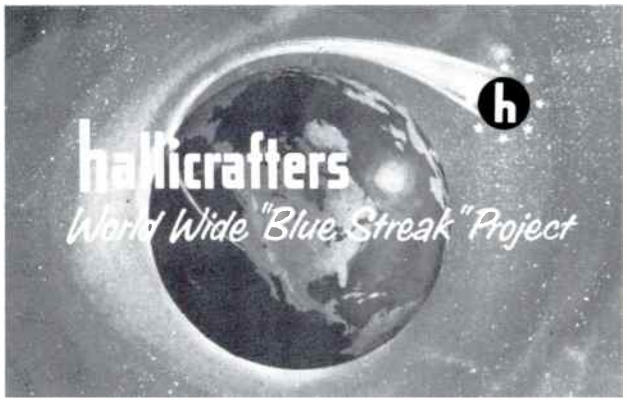
hallicrafters company

Military Electronics Division
Chicago 24, Illinois

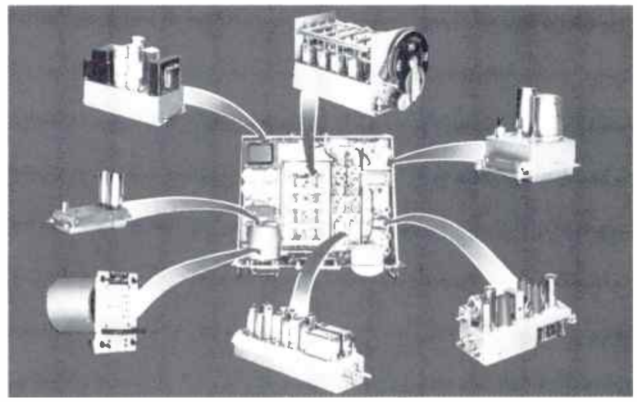


"There is but one lamp by which our steps are guided—the lamp of experience"
–Patrick Henry

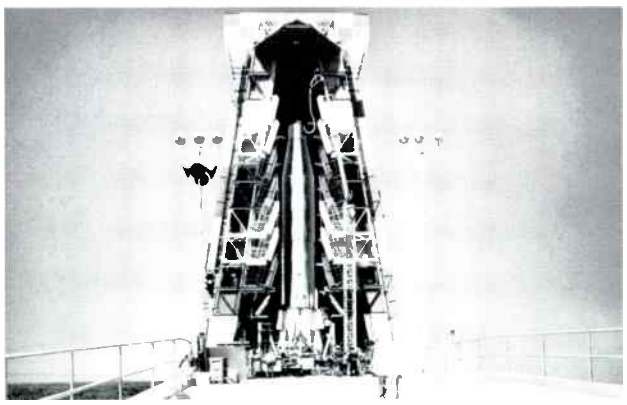
For more than a quarter century, Hallicrafters has worked in close partnership with our armed forces on fast solutions to critical military electronics problems. Out of this priceless experience are emerging startling new ideas and hard-hitting, fast-moving techniques to keep our country one jump ahead in electronic warfare . . .



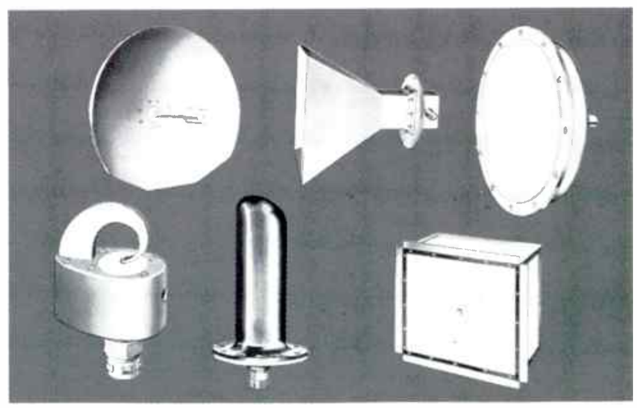
New levels of speed and efficiency are being reached in equipment modernization, retrofit and technical support programs with Hallicrafters' radical new "Blue Streak" project. Specially-trained Maintenance and Technical Support Teams, close-knit and flexible, can be tactically deployed to accomplish maintenance, installation and testing of electronics weapons systems anywhere in the world.



Hallicrafters communications leadership is exemplified by new high frequency Single Sideband receiver, (model no. S.W.E.). 100% modular design permits simple modification for compatability with existing and future communications systems. Stability, with proper available plug-ins, is better than one part in 10,000,000 per month. Hallicrafters also offers an existing capability in receiving and transmitting techniques up to frequencies of 50,000 megacycles.



Hallicrafters participation in the Atlas missile project helped to develop capability for many areas of the complex missile field, including code translator data systems; ground support equipment; ECM testing and antenna systems. Current explorations involve latest Infra Red techniques.



Airborne antennas and micro-wave components with power capability in excess of 1,000 watts, can be made available to solve tomorrow's very high power handling requirements. Testing of microwave components is possible with special high power generators, designed and built by Hallicrafters.

company

For further information on Hallicrafters facilities and experience in military electronics research, development and production, please write to:

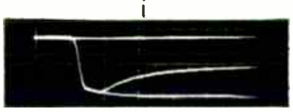
hallicrafters company

Military Electronics Division, 4401 W. 5th Ave., Chicago 24, Ill.

Looking for a challenging new opportunity? We are interested in qualified engineers at all levels. For full details in confidence, contact William F. Frankart, Director of Engineering.

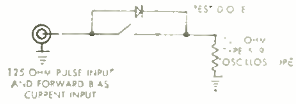
for clear display of:

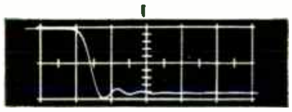
Single-shot nuclear events • Fast diode turn-on • Avalanche transistor waveforms • Tunnel diode waveforms
Recovery time of fast diodes • High-speed circuit response • Transistor switching waveforms • Radio frequency waveform



2 nsec/cm

Switching and storage times in fast transistors and diodes can be measured using the outstanding characteristics of the Type 519. In this typical diode-recovery-time waveform, the upper trace is +45 ma reference, the middle trace shows the diode turn-off, and the lower trace shows the diode shorted.





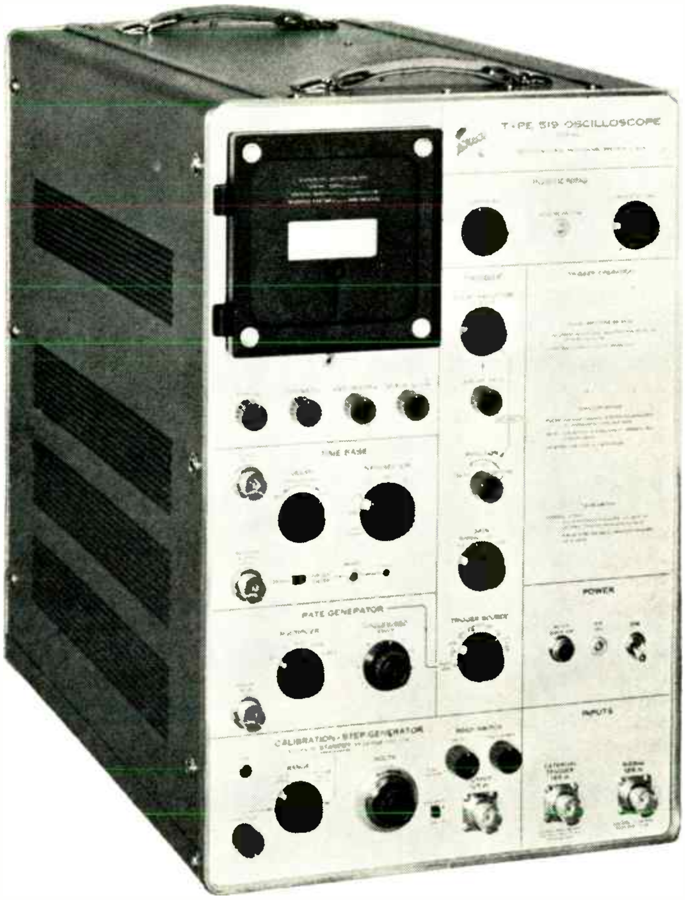
5 nsec/cm

The Type 519 Oscilloscope is an invaluable tool for testing active or passive wideband circuits. In this wideband amplifier waveform little or no correction is necessary for the inherent risetime of the oscilloscope.

The NEW

TEKTRONIX

KMC OSCILLOSCOPE



Type 519
DC to over 1000 Megacycles
24-kv Accelerating Potential

The new Tektronix Type 519 Oscilloscope is a calibrated high-speed laboratory instrument designed for observation, measurement, and photographic recording of fractional nanosecond risetimes. A 2 x 6 cm viewing area coupled with 24-kv accelerating potential affords bright displays with excellent definition.

Performance features include: passband from dc to beyond 1000 mc, risetime less than 0.35 nsec, sensitivity of 10 v/cm, linear sweeps to 2 nsec/cm, sweep delay to 35 nsec, and a sensitive wideband triggering system. All features are fully compatible with the signal bandwidth capabilities of the instrument. The single unit houses a fixed signal delay line, a convenient sweep-delay control, a pulse-rate generator, a standard amplitude and waveform generator, and electronically-regulated power and high voltage supplies. Only two connections are necessary for normal operation—signal input and power cord.

Combining simple operation with laboratory precision and reliability, the Tektronix Type 519 Oscilloscope is ideally suited to single-shot or random nuclear events. In addition, the extreme bandwidth of the Type 519 permits application to general measurements where oscilloscope risetime must be much faster than signal risetime.

PRICE, Type 519 . . . . **$3800**
f.o.b. factory

Please call your Tektronix Field Engineer for complete specifications and a demonstration of the Type 519 in your own applications.

CAREER OPPORTUNITIES now exist at Tektronix in the following fields: Instrument design, Circuit design and engineering, Cathode ray tubes, Electron physics, Solid state and semi-conductor devices. For information write to Irving Smith, Personnel Director.

Tektronix, Inc.

P. O. Box 500 • Beaverton, Oregon
Phone MItchell 4-0161 • TWX—BEAV 311 • Cable: TEKTRONIX

TEKTRONIX FIELD OFFICES: Albuquerque, N. Mex. • Atlanta, Ga. • Baltimore (Towson, Md.) • Boston (Lexington, Mass.) • Buffalo, N.Y. • Chicago (Park Ridge, Ill.) • Cleveland, Ohio • Dallas, Texas • Dayton, Ohio • Denver, Colo. • Detroit (Lathrup Village, Mich.) • Endicott (Endwell, N.Y.) • Greensboro, N.C. • Houston, Texas • Kansas City (Mission, Kan.) • Los Angeles, Calif., area (East L.A., Encino, West L.A.) • Minneapolis, Minn. • New York City Area (Albertson, L.I., N.Y. • Stamford, Conn. • Union, N.J.) • Orlando, Fla. • Philadelphia, Pa. • Phoenix, (Scottsdale, Ariz.) • Poughkeepsie, N.Y. • San Diego, Calif. • San Francisco (Palo Alto, Calif.) • St. Petersburg, Fla. • Syracuse, N.Y. • Toronto (Willowdale, Ont.) Canada • Washington, D.C. (Annandale, Va.)

TEKTRONIX ENGINEERING REPRESENTATIVES: Hawthorne Electronics· Portland, Oregon • Seattle, Washington. Tektronix is represented in twenty overseas countries by qualified engineering organizations.

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

SEE THE TYPE 519 AND OTHER NEW TEKTRONIX INSTRUMENTS AT WESCON, BOOTHS 817 AND 818

NEWS New Products

Peak Following Amplifier

Model V63 peak following amplifier, a product of **Video Instruments Co., Inc.** 3002 Pennsylvania Ave., Santa Monica, Calif., is used to amplify signals from crystal type accelerometers. The unique feature of this amplifier is that it produces a proportional dc level for each peak of acceleration. This makes it possible to use peaks of sine waves or random waves as control functions. For example in vibration tests, "safe" limits can be established through the use of a reference voltage. Whenever the dc output level of the V63 exceeds the reference voltage, a relay becomes energized and can be used to shut down the test, or sound an alarm, before the test becomes destructive

At its output, the V63 provides up to 100 ma. This high current output makes it possible to drive high frequency recorders with the V63. If the chart speed is sufficiently slow, the resulting curve will be the envelope of the vibration curve. Thus this amplifier also performs a task of data reduction.

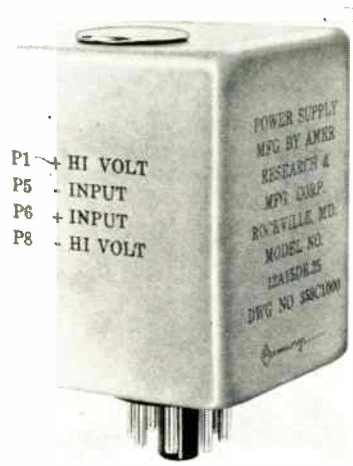
The input impedance of the V63 is in excess of 1,000 megohms. The bandwidth is from 5 cps to 5 kc. The output however is dc coupled, at an impedance of less than 1 ohm, from dc to 5 kc. The unit has an integral, regulated power supply.

Transistorized DC To DC Converter

American Research & Manufacturing Corp., 920 Halpine Ave., Rockville, Md., has developed a line of miniature dc to dc power supplies providing a range of high voltages at low current levels, suitable for such applications as Geiger tubes, photomultipliers, cathode ray tubes, photo flash units, infra-red detection and other low-power consuming equipment.

A ringing choke-type converter, using a single germanium transistor, provides up

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



to 0.75 watt power output, up to 80% efficiency. Output voltage of unregulated units is adjustable via external resistors. Proper model selector makes possible unregulated output voltages of from 350 to 2100 volts dc, at load currents of up to 250 microamperes. The units are designed to operate from 6 volts or 12 volts dc power sources, each unit equipped with a standard 8-pin octal socket. With exception of externally removable transistors, the units are completely encapsulated. Maximum ripple is 1% rms.

Utilizing corona-type voltage regulator tubes to achieve regulation of output voltages, the units provide regulated outputs at 400, 500, 600, 700, 800, 900, 1000, 1200, 1500, 1800 and 2000 volts. Typical variations in load voltage are ±7 volts for a 400 volt unit when the load current is varied from 5 to 150 ma, and a ±35 volt change in the output voltage for an 1800 volt unit when the load current changes from 15 to 250 ma.

Varying with size of order, prices of unregulated units range from $65.25 to $81.50, regulated units from $71.00 to $88.85.

Telemetering Test Oscillator

Crosby-Teletronics Corp., 54 Kinkle St., Westbury, L. I., N. Y., has developed an advanced telemetering test oscillator which is being used to ground check components in the Polaris missile program.

The Model TO-258 Telemetering Test Oscillator, designed with high frequency stability, provides accurate calibration of sub-carrier units in the FM/FM telemetering system.

High frequency stability and deviation control calibrated directly in percent make this instrument suited for production testing and other applications utilizing standard test frequencies. Models can be supplied with any 20 frequencies from 20 cps to 100 kc. Selection of frequencies is made by push-button.

Wilson President of Hazeltine

Directors of **Hazeltine Corporation** have announced the election of Webster H. Wilson as president. Wilson had been executive vice president for operations of Hazeltine Electronics Division and a director of Hazeltine Corporation since 1958. He succeeds William A. MacDonald, who continues as chairman of the board.

Wilson joined Hazeltine in 1946 and became a chief project administrator and later vice president of the Government and Commercial Department. The new Hazeltine president was graduated from Phillips Academy, Andover, Massachusetts and received his Bachelor of Science degree from Massachusetts Institute of Technology in 1936, where he majored in Business Administration and Aeronautical Engineering. As a naval officer during World War II, he supervised the installation and maintenance of shipboard electronic equipment. Prior to the war, he was an engineer with the Boston Edison Company.

Wilfred M. McFarland was elected vice chairman of the board, a position which was recently established. He has been a director since 1953. Mr. McFarland retains his post as executive vice president of Hazeltine and as president of Hazeltine Technical Development Center, Inc., the company's Indiana subsidiary.

(Continued on page 38A)

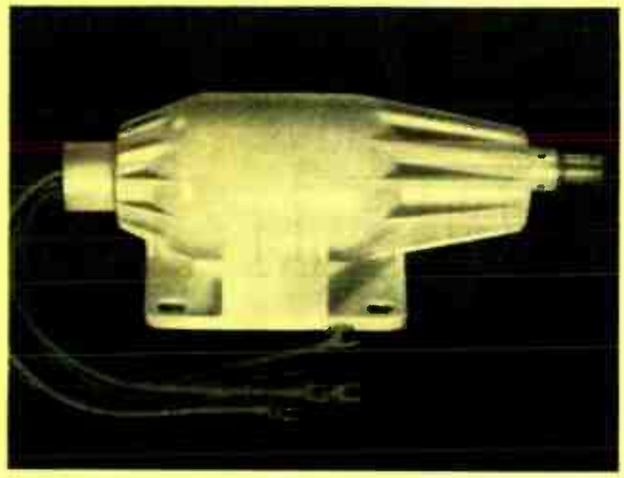
Creative Microwave Technology

Published by MICROWAVE AND POWER TUBE DIVISION, RAYTHEON COMPANY, WALTHAM 54, MASS., Vol. 2, No. 2

A TOTALLY NEW CONCEPT IN "O"-TYPE BWO CONSTRUCTION

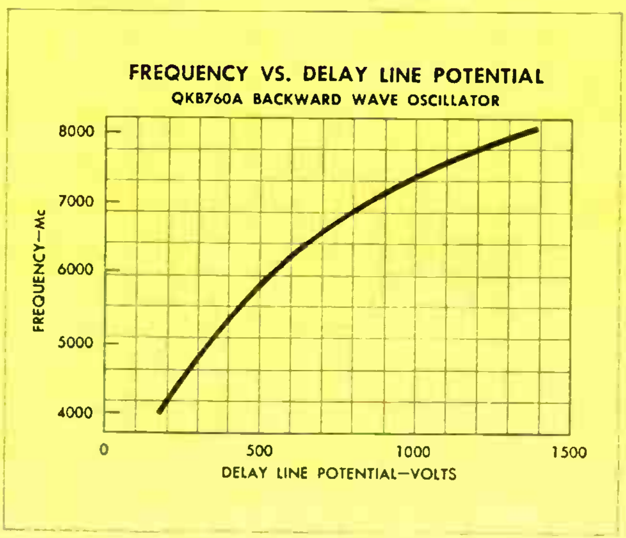
--Interdigital-type delay line affords maximum heat dissipation at high power outputs

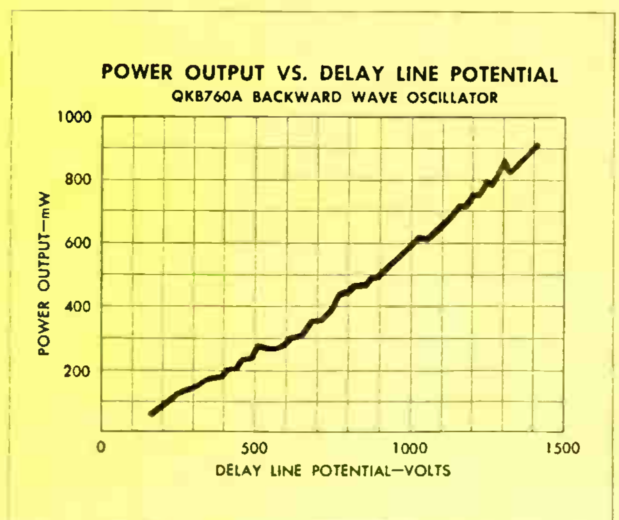
These broadband voltage tunable backward wave oscillators are the smallest, lightest and most reliable of their kind. They were developed especially for modern airborne and ground-based applications utilizing swept oscillator and frequency diversity techniques. Four compatible types are available. They cover a continuous frequency range of 1 to 12.4 KMC. They are magnetically shielded and are insensitive to the effects of external fields. They exhibit a minimum of fine-grain power output variations. Potted leads permit operation at high altitudes over a wide temperature range. Raytheon-perfected laminating techniques make possible interdigital construction which results in maximum heat dissipation. Under normal operating conditions, no forced-air cooling or protective circuitry is required. Laminate-thickness held to extremely close tolerances assures improved fine-grain frequency characteristics with optimum line matching and consistently reproducible characteristics from tube to tube.



Typical Operating Characteristics

	QKB786	QKB816A	QKB760A	QKB776
Frequency Range	1.0-2.0KMC	2.0-4.0KMC	4.0-8.0KMC	8.0-12.4KMC
Power Output	100 mW Min.	70 mW Min.	30 mW Min.	50 mW Min.
Delay Line (Tuning) Voltage	100-1500 Vdc			
Filament Voltage	6.3 V.			
Cathode Current	45 mA Max.			
Anode Voltage	60-150Vdc	100-200 Vdc	60-130 Vdc	60-130 Vdc
Control Grid Cut-off	-150 Vdc	-100 Vdc	-100 Vdc	-100 Vdc





Excellence in Electronics

You can obtain detailed application information and special development services by contacting: Microwave and Power Tube Division, Raytheon Co., Waltham 54, Mass. In Canada: E. Waterloo, Ontario. In Europe: Zurich, Switzerland.

RAYTHEON

A LEADER IN CREATIVE MICROWAVE TECHNOLOGY

SEE THESE TUBES AT RAYTHEON'S WESCON BOOTH.

(Continued from page 36A)

Beetham VP and Manager of Knights

The James Knights Co., Sandwich, Ill., announces the election of four new vice presidents: Robert Beetham becomes vice president and manager of the Knights filter products department. Beetham joined James Knights in 1954. He holds post-graduate degrees from Ohio State University and is widely known as an electronic scientist.

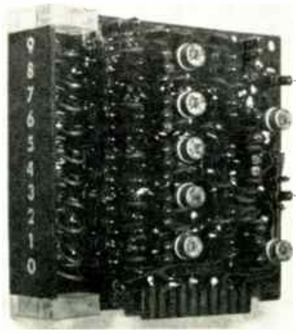
Louis Dick was elected vice president and chief crystallographer. Mr. Dick's experience includes work in piezoelectricity at Colorado A & M College. He was manager of Motorola's crystal activity prior to joining Knights in 1954.

David Larsen, a 1952 graduate in electrical engineering from Colorado University, becomes vice president and production manager of the crystal plant.

Glen Munro becomes vice president and general sales manager. Munro has an extensive background in the electronics market, including 10 years as a regional sales manager of Motorola.

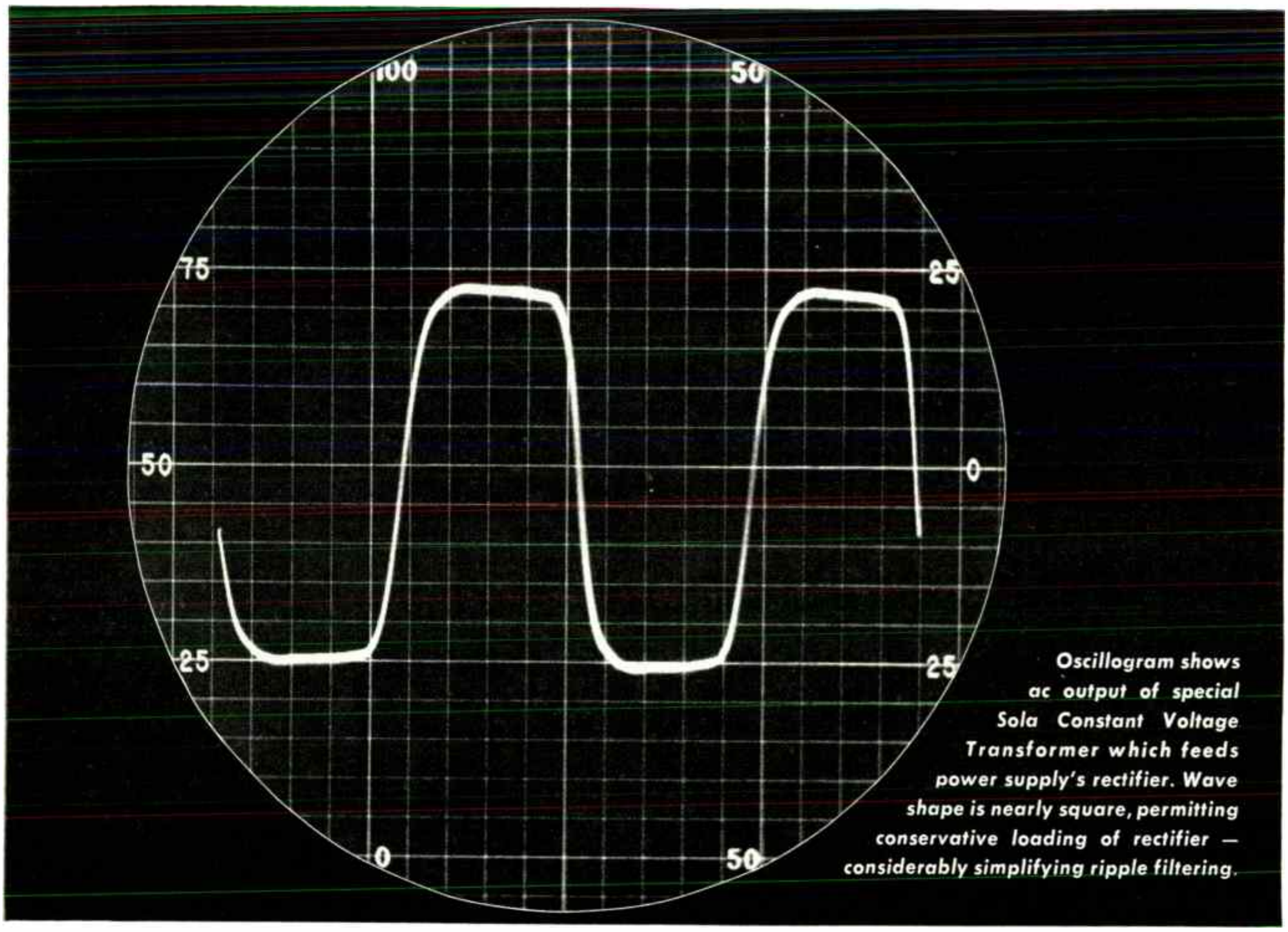
Add-Subtract Decade Counter

A new transistorized 100 kc add-subtract decade counter which features simple switching control is available from Robotomics, Inc., 4624 E. Garfield, Phoenix, Ariz. Grounding one of two control lines determines whether decade will add or subtract. Available in single voltage model 1303 requiring +100 volts at 40 ma; or lower dual voltage model 1203 requiring +30 volts at 35 ma., and 60 volts ac at



1 ma. Bright, ten number display, electrical zero reset, and optional coincidence output. Plugs into standard 10-pin PC connector. Size: $3\frac{9}{16}$ H. × 1 W. × $3\frac{3}{4}$ inches D. Weight is 3 ounces. Availability: 2–4 weeks after receipt of order in quantities of 1 to 24. Model 1203, production quantities of 100 & up $63.00, Model 1303, $73.00.

(Continued on page 110A)



Square-wave output of special transformer gives high efficiency in Sola's regulated dc power supply

Sola engineers (men with a keen eye for a trim wave shape) designed a special constant voltage transformer having nearly a square-wave output. Then they linked the transformer with two other components to produce a regulated dc power supply which has notable efficiency.

They fed the regulated output of this transformer into a semiconductor rectifier . . . the low-peak characteristic of the square wave results in a conservative loading on the economical rectifier assembly. It can deliver considerable amounts of current as long as you don't over-voltage it—and over-voltaging just doesn't happen when the input to the rectifier is Sola-regulated to within ±1%.

The rectified voltage feeds into the third component in this happy combination—the high-capacitance filter. The capacitor's filtering job is made easier because the rectified square wave contains a comparatively small amount of ripple. Final dc output from the filter has less than 1% rms ripple . . . for many applications there is no need for a voltage-dropping, efficiency-cutting choke coil.

The Sola Constant Voltage DC Power Supply has output in the ampere range, regulates within ±1% even under ±10% line voltage variations, and is suitable for intermittent, variable, and pulse loads. It has low output impedance, is very compact, and provides about all you could ask for in maintenance-free dependability.

Hundreds of ratings of these dc power supplies have been designed and produced to meet widely varying electrical and mechanical requirements of equipment manufacturers. In addition, there are six stock variable-output models and six stock fixed-output models with ratings from 24 volts at six amps to 250 volts at one amp.

For complete data write for Bulletin 1H-DC

SOLA

A DIVISION OF BPC BASIC PRODUCTS CORPORATION

SOLA ELECTRIC CO., Busse Rd. at Lunt, Elk Grove, Illinois, HEmpstead 9-2800 • In Canada, Sola-Basic Products Ltd., 377 Evans Ave., Toronto 18, Ont.

IRE People

The appointment of **Harland A. Bass** (M'44–SM'52) as Chief Engineer, Data Handling, was announced today by Mr. R. M. Bukaty, Vice President, Electronics and Controls Operation, Crosley Division, Avco Corporation.

H. A. BASS

He had been serving as Assistant Chief Engineer, Data Handling, and first joined Crosley in 1940 as an engineer on military products.

A native of Waterloo, Iowa, Mr. Bass attended Iowa State Teachers College and was graduated with a Bachelor of Science in Electrical Engineering from the University of Iowa.

❖

With the appointment of **Joseph D. Bianco** (M'56) to the newly created post of sales manager, Components for Research, Inc., Palo Alto, California, marks the completion of four years of activity in the potting and encapsulation of electronic assemblies. The appointment is part of a major expansion program which will provide a broad line of specialty epoxy-resin components for ultra-high-voltage application.

Mr. Bianco leaves Hughes Aircraft Company's Vacuum Tube Products division in Oceanside, Calif., where he was plant manager, to accept this post. Prior to his position at Hughes, he was in the research and development laboratories of Eitel-McCullough, Inc., serving as a project engineer on the large klystron program for the BMEWS radar.

❖

Appointment of **Howard A. Bond** (SM'57) as manager of reconnaissance systems in Stromberg-Carlson's Electronics Division has been announced by R. J. Gilson, director of systems management. Stromberg-Carlson is a division of General Dynamics Corporation.

H. A. BOND

In this capacity he will have responsibility for management of a 27-million-dollar contract under which Stromberg-Carlson is designing and building a complex electronic reconnaissance system for the U. S. Air Force.

He has been with Stromberg-Carlson since 1957, when he joined the company as manager of electronic systems design. For 10 years previously, he was engaged in security work for the U. S. government, first with the Naval Security Agency, and later as chief of a radio frequency division of the National Security Agency.

From 1945 to 1947 he served in the U. S. Army with the rank of lieutenant, first in the Signal Corps, and later as a project engineer developing speech privacy equipment for the Army Security Agency. He now holds the rank of major in the U. S. Army Reserve, assigned to headquarters of the Army Security Agency.

Mr. Bond is an electrical engineering graduate of Northwestern University, Evanston, Ill., and did graduate work at the University of Maryland, College Park. He is a member of Tau Beta Pi, honorary electrical engineering society.

❖

Hugh C. Bream (M'56–SM'57) has been named President and General Manager of Western Design, Santa Barbara Airport, Goleta, California, a division of U. S. Industries, Inc., according to an announcement by John I. Snyder, Jr., Chairman of the board and President of U. S. Industries, Inc.

H. C. BREAM

Previously, he was manager of Western Design's Santa Barbara division. He has had extensive experience in engineering and technical management. Before joining Western Design, he was Vice President-Marketing, Hoover Electronics Company, and Manager of the Electronics Division, Rheem Manufacturing Company. Earlier, he was engaged in the development and engineering of instrumentation programs for the Ralph M. Parsons Company.

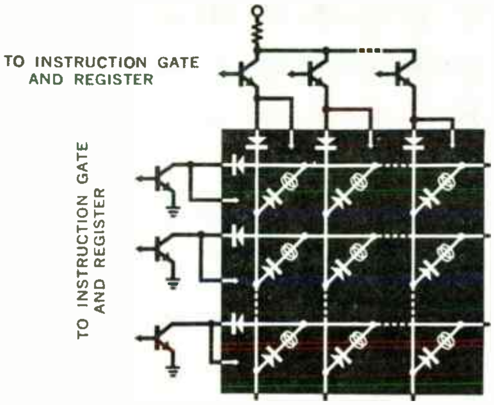
Mr. Bream attended Marin Junior College, Kentfield, Calif., San Jose College, San Jose, Calif., the California Institute of Technology, Pasadena, and the University of California at Los Angeles. He is a member of the Electronics Committee of the Los Angeles Chamber of Commerce, the American Rocket Society, Armed Forces Communications and Electronics Association and the American Management Association.

❖

Dr. Cledo Brunetti (A'37–SM'46–F'49) was elected President of Grand Central Rocket Company at a meeting of the Board of Directors on June 6, 1960, it has been disclosed by Major General J. W. Sessums, Chairman. Grand Central, third

(Continued on page 44A)

NEW



Typical n-junction matrix for n-stage matrix configuration. Fairchild 2N1613 transistors and FD200 diodes, used throughout, guarantee acceptable leakage, switching speed and conductance values up to 125°C.

ANSWER

TO COMPUTER MATRIX PROBLEMS

LOW LEAKAGE TRANSISTORS AND FAST RECOVERY, LOW CAPACITANCE DIODES FROM FAIRCHILD

Approach to the ideal matrix. 2N1613 silicon transistors and FD200 silicon diodes from Fairchild are unique in making feasible the ideal matrix. They give you low leakage and low capacitance with high conductance and high speed, even at high ambient temperatures. These characteristics are combined only in Fairchild Planar devices. With them you can now largely ignore stray leakage or capacitance build-up across the matrix. Temperature effects and long-term performance decay are no longer critical. You can eliminate complex circuitry previously necessary in designing around these losses.

Fairchild's Planar structure for transistors and diodes features the industry's most advanced diffusion and surface passivation techniques. Current leakage is reduced to 10 mμA maximum (2N1613) and 0.1μA maximum (FD200) at 25°C. Maximum values at 150°C are 10μA and 100μA.

Surface passivation also prevents significant degeneration of parameters during circuit life which could introduce error or failure in the matrix. This technique also lends itself to precisely controlled manufacture, assuring excellent product uniformity.

2N1613 ELECTRICAL CHARACTERISTICS (25°C except as noted)

Symbol	Characteristic	Min.	Typical	Max.	Test Conditions	
h_{FE}	D.C. Current Gain	40		120	I_C = 150 mA	V_{CE} = 10V
$V_{BE(sat)}$	Base Saturation Voltage			1.3V	I_C = 150 mA	I_B = 15 mA
$V_{CE(sat)}$	Collector Saturation Voltage			1.5V	I_C = 150 mA	I_B = 15 mA
C_{ob}	Collector Capacitance		18	25μμf	I_E = 0	V_{CB} = 10V
I_{CBO}	Collector Cutoff Current		0.8mμ A 1.0μ A	10mμ A 10μ A	V_{CB} = 60 V_{CB} = 60	T = 25°C T = 150°C

FD200 ELECTRICAL SPECIFICATIONS (25°C except as noted)

Symbol	Characteristic	Min.	Typical	Max.	Test Conditions	
V_F	Forward Voltage			1.0V	I_F = 100 mA	
I_R	Reverse Current			0.1μA	V_R = −150V	
I_R	Reverse Current (150°C)			100μA	V_R = −150V	
B_V	Breakdown Voltage	200 V			I_R = 100μA	
t_{rr}	Reverse Recovery Time			50.0 mμ sec	I_f = 30 mA I_r = 30 mA	R_L = 150Ω
C_o	Capacitance			5.0μμf	V_R = 0V	f = 1 mc
RE	Rectification Efficiency	35%			100 mc	
	Forward Voltage Temperature Coefficient		−1.8 mV/oC			

FAIRCHILD SEMICONDUCTOR CORPORATION

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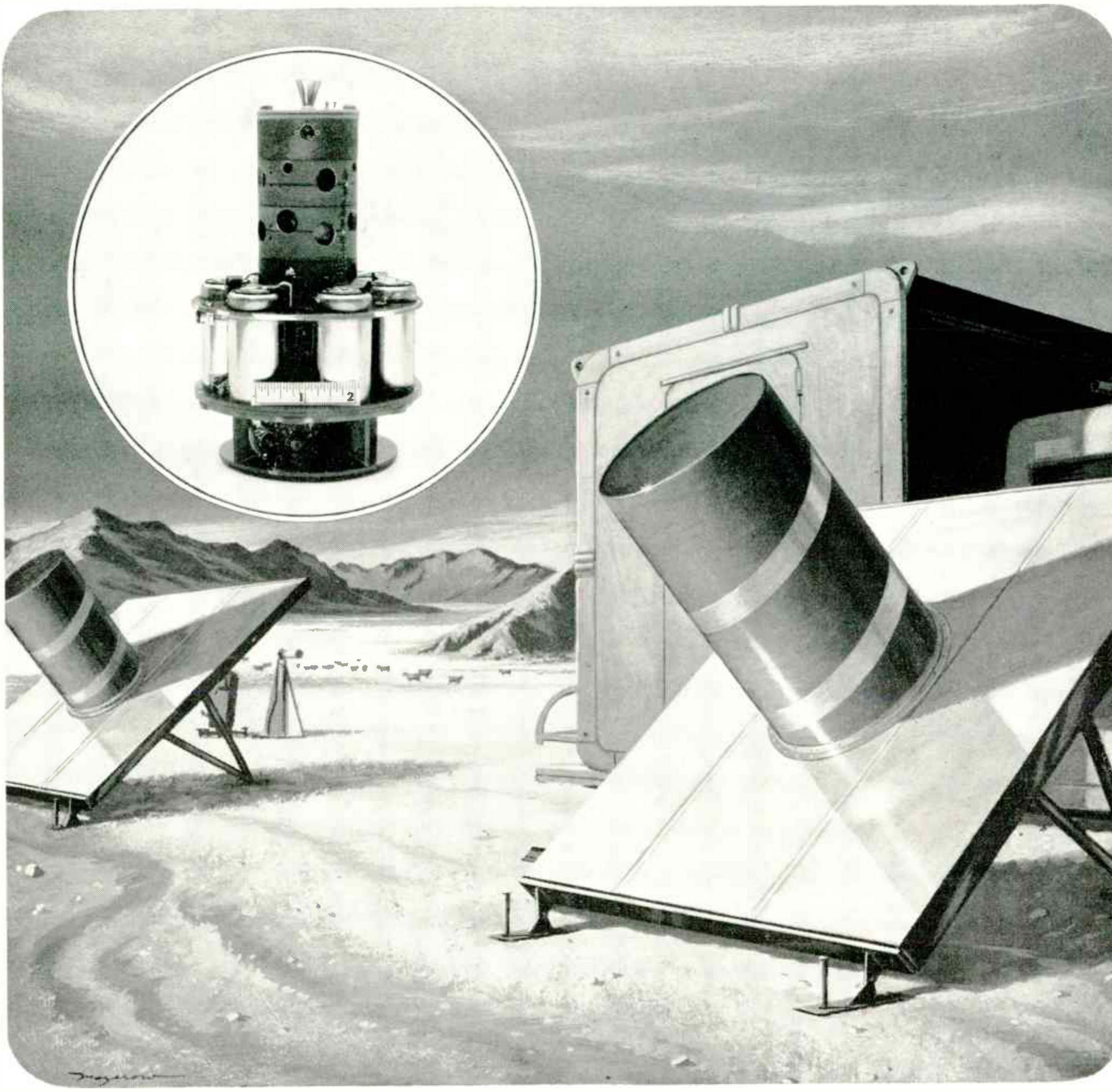
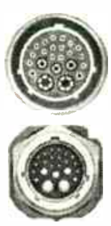


NEW FROM BENDIX

QWLD HUS-KEY ELECTRICAL CONNECTORS

FOR MISSILE GROUND SUPPORT EQUIPMENT



WITH SUCH OUTSTANDING ADVANTAGES AS:

FIVE INTEGRAL KEYS AND KEYWAYS that provide for positive polarization and positive mating—even in blind locations. ***No possible contact damage.*** QWLD connectors can be fully mated and unmated by hand.

HEAVY-DUTY CONSTRUCTION through use of extra-heavy machined or forged aluminum shell components, resilient inserts, silver-plated copper alloy contacts, and rugged cable accessories with new superior gasket design.

TWO NEW SERIES AVAILABLE with the QWLD having standard solder or solderless contacts and the QWLG having provisions for grounding one contact to the shell.

PLUS . . . IMPROVED WATERPROOFING • CLOSED ENTRY SOCKET CONTACTS • SELF-EJECTING COUPLING ACTION • DESIGNED TO MEET MILITARY SPECIFICATIONS • WIDEST RANGE OF CABLE ACCESSORIES.

QWLD is the latest development in Scintilla's long line of multiple conductor cable connectors—and is specially designed to meet the rugged environmental conditions of missile launching equipment, ground radar, or power and control circuits—and heavy-duty industrial applications such as are found in oil fields or mining. Be sure to investigate the new QWLD HUS-KEY* Connectors from Scintilla.

*TRADEMARK

Scintilla Division
SIDNEY, NEW YORK

THE Bendix CORPORATION

IRE People

(Continued from page 40A)

largest company of its type in the United States, is one of the pioneers in the field of solid propulsion.

General Sessums said that the strong leadership that Dr. Brunetti, as Vice President and General Manager, has provided has had a great influence in building the Grand Central Rocket Organization into a team of the nation's most outstanding solid propellant scientists. His vast technical experience at Stanford Research, the Bureau of Standards and General Mills has been invaluable to the vigorous growth of GCR.

He has been Managing Director of Engineering, Research and Development for General Mills, Inc., Minneapolis, Minn., and Associate Director of Stanford Research Institute of Stanford University, Stanford, Calif. He spent eight years at the National Bureau of Standards in Washington as head of Ordnance Development, Production Engineering and finally as Chief of Engineering Electronics. He was one of the leaders of the Proximity Fuze in Washington, a major, secret weapon of World War II. He served as a member of a group of national specialists assigned to Woods Hole, Mass., two years ago to study the future scientific aspects of the U. S. military picture for the national Academy of Sciences under sponsorship of the U. S. Air Force. Prior to his assignment with Grand Central, he was an executive of Food Machinery and Chemical Corporation.

Dr. Brunetti is special consultant to the Assistant Secretary of Defense and has been chairman or member of many defense committees. A man of many distinctions and honors, he was nationally identified in 1942 by Eta Kappa Nu, national honorary fraternity, as the outstanding electrical engineer in the United States.

❖

Arthur B. Buchanan (A'26–M'32–SM'32), Assistant Communication Engineer, The Detroit Edison Company, retired August 1, 1960, to establish an independent practice as an Engineering Consultant. He was engaged primarily in radio and electronic engineering work with Edison for 38 years. He has been active in IRE affairs for the past 34 years, and has been on the Executive Committee of the Detroit Section, IRE during that entire period. He was Section Chairman from July, 1929 to June, 1930, and again during 1935, when the IRE Summer Convention was held in Detroit. He was General Chairman of the IRE Summer Convention in Detroit in 1941.

He was the organizer and first Chairman of the IRE Professional Group on Vehicular Communication, and has served on the Administrative Board ever since. He edited the PGVC NEWSLETTER from 1953 to 1959.

(Continued on page 46A)

IRE People

(Continued from page 44A)

Mr. Buchanan was born in Sault Ste. Marie, Mich., in July, 1895, and obtained the B.S. degree in E.E. from the University of Michigan, Ann Arbor, in 1922. He is a Registered Professional Engineer and plans to specialized in mobile radio systems.

❖

Monte Cohen (A'30–VA'39–M'55), new Vice Chairman of the Board of Directors of General Instrument Corporation, is a veteran of 44 years in the electronics industry, whose career dates back to the earliest days of "earphone radio" and the old Marconi Company (now Radio Corporation of America) where he got his first job at the age of 16.

M. COHEN

President of General Instrument from 1953 until his election as Vice Chairman (he retains his active operating post as Chairman of the Operations Committee), he has been most closely identified with the Company's F. W. Sickles Division at Chicopee, Massachusetts, which is today the largest manufacturer in the U. S. of TV deflection components and one of the leading producers of television tuners, as well as of a wide variety of other components.

With the Sickles Division for 31 years (he joined it in 1929 when it was an independent company), he became General Manager of the firm in 1931, remained with it when it was merged into General Instrument in 1945, was named President of the Sickles Division in 1951 and in the same year elected Executive Vice President of the parent General Instrument. In December, 1953, he was elected President of the Corporation, a post he held until his election as Vice Chairman.

One of the best known men in the electronics industry, Mr. Cohen is a former director of the Radio and Television Manufacturers Association (now the Electronics Industries Association), a member of the Veteran Wireless Operators Association and a director of the Springfield (Mass.) National Bank. He served as an advisor to the War Production Board during World War II.

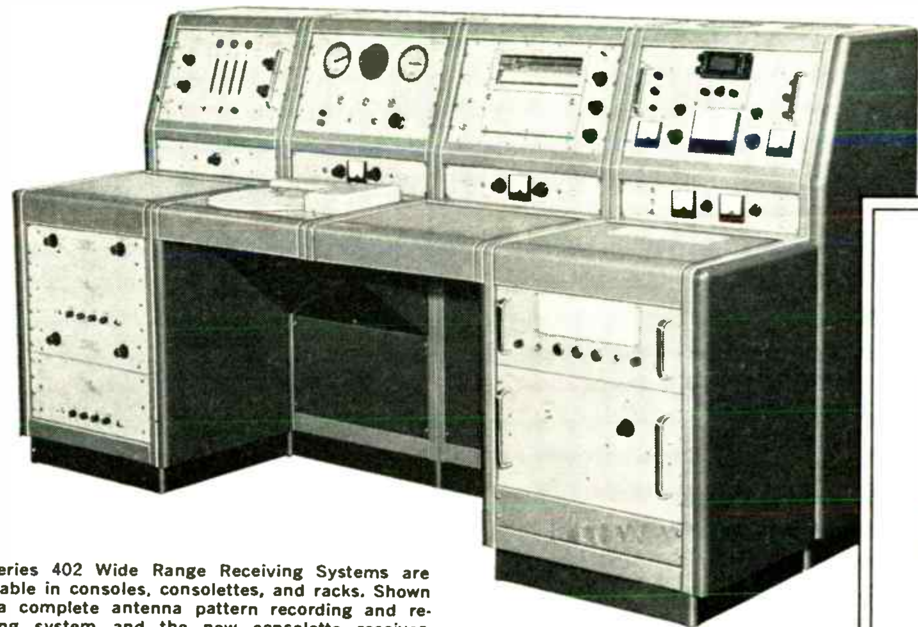
❖

The appointment of **Bernard Elbinger** (S'49–A'50–M'55) as Head of the Electronic Instrumentation Section of the Rheem Semiconductor Corporation has been announced by Dr. E. M. Baldwin, Vice President and General Manager of the Mountain View, California silicon mesa transistor, diode, and special assemblies manufacturing company.

(Continued on page 48A)

MAJORS and MINORS

. . . A Message to the Antenna Designer




● Series 402 Wide Range Receiving Systems are available in consoles, consolettes, and racks. Shown are a complete antenna pattern recording and receiving system and the new consolette receiver.

A crowded spectrum plus high power radar and communication systems critically compound the problems of the antenna design engineer.

More than ever, the complete pattern including all the major and minor lobes of every radiating element must be graphed for sound engineering evaluation.

S-A Receiver Gets the Whole Signal

Scientific-Atlanta Series 402 Wide Range Receiving Systems are specifically designed for antenna pattern measurements. Unique in design, these receivers combine maximum sensitivity and linearity from 30 mc to above 100 kmc. They are also useful as multipurpose laboratory instruments for microwave testing, monitoring, and measuring applications.

Only from S-A, 1 db Linearity over Full 60 db Dynamic Range

A recent development, S-A's P-4 modification adds 20 db to the normal 40 db dynamic range. The modification takes advantage of the gain vs AGC voltage characteristics of the Series 402. Existing receivers can be modified at the factory.

New Modification Z Broadens Use

Modification Z adds a precision IF attenuator and VTVM to the Series 402. Now RF and microwave signal level, gain, and isolation measurements can be made with fewer components and instruments. *For instance,* an X band 80 db attenuator can be calibrated to within ±0.5 db with a 1 mw signal source, a flap attenuator, a mixer, and an S-A Series 402Z Receiver. Antenna gain can be measured by direct comparison with a standard gain antenna. Signal levels can be compared against a reference standard.

Other Features

One coaxial cable from antenna to receiver eliminates costly lossy waveguides and rotary joints. Antenna can be located up to 75 feet away with negligible loss in sensitivity ☆ One receiving system covers 30 mc to above 100 kmc without plug-ins ☆ Reception of cw signals from simple sources eliminates need for precise modulation ☆ High sensitivity means low source power and long ranges ☆ High selectivity reduces interference and cross talk between adjacent test ranges ☆ Positive AFC action over full dynamic range provides pattern recording in deep nulls.

PRICES

Series 402, 2 to above 100 kmc	$7500
Series 402A, 2 to above 100 kmc with AGC	8000
Series 402B, 30 mc to above 100 kmc	8500
Series 402C, 30 mc to above 100 kmc with AGC	9000
Modification P-4	500
Modification Z	1000

NEW DATA FOLDER READY

For complete information ask for our new data folder from your nearby S-A engineering representative or write directly to Dept. 86.

SCIENTIFIC-ATLANTA, INC.
2162 PIEDMONT ROAD, N. E.
ATLANTA 9, GEORGIA
PHONE: TRinity 5-7291

3

IRE People

(Continued from page 46A)

Mr. Elbinger holds the Bachelor of Science degree in Electrical Engineering from the University of Southern California, Los Angeles and has taken graduate courses from the University of California at Los Angeles and the Navy Airborne Radio-Radar Maintenance School.

He was an Electronic Development Engineer for Aerojet-General Corporation working on the design and development of electronics equipment, infra-red missile guidance, tracking and detection systems. At Hughes Aircraft Company, he headed the Electronic Tooling Section, developing and fabricating semiconductor testing and processing equipment for manufacturing and engineering sections. His background also includes positions as Manager, Special Products Section for Electronic Control Systems and Head, Electronic Instrumentation for Fairchild Semiconductor Corporation.

Mr. Elbinger has applied for a patent in special signal detecting circuit.

❖

Thomas A. Combellick (A'52–SM'56), formerly Chief Engineer for Military Products Development, has been appointed Government Marketing Manager for Lenkurt Electric Co., Inc.

T. A. Combellick

He joined Lenkurt in 1955 as a project engineer, and during the next two years served as manager of several development and applications engineering departments. In November, 1957, he was appointed manager of the company's Washington D. C., sales and engineering offices, with responsibility for customer relations in the eastern U. S. He returned to San Carlos and the Chief Engineer position in March, 1959.

Before joining Lenkurt he spent five years with Westinghouse, where he was in charge of microwave system development, and five years with Western Electric, where he was project engineer on the manufacture and test of tubes for the TD-2 microwave system.

He was graduated from Hampden Sydney College, Va., with the B.S. degree and pursued postgraduate studies at the Harvard Graduate School of Engineering and at Massachusetts Institute of Technology.

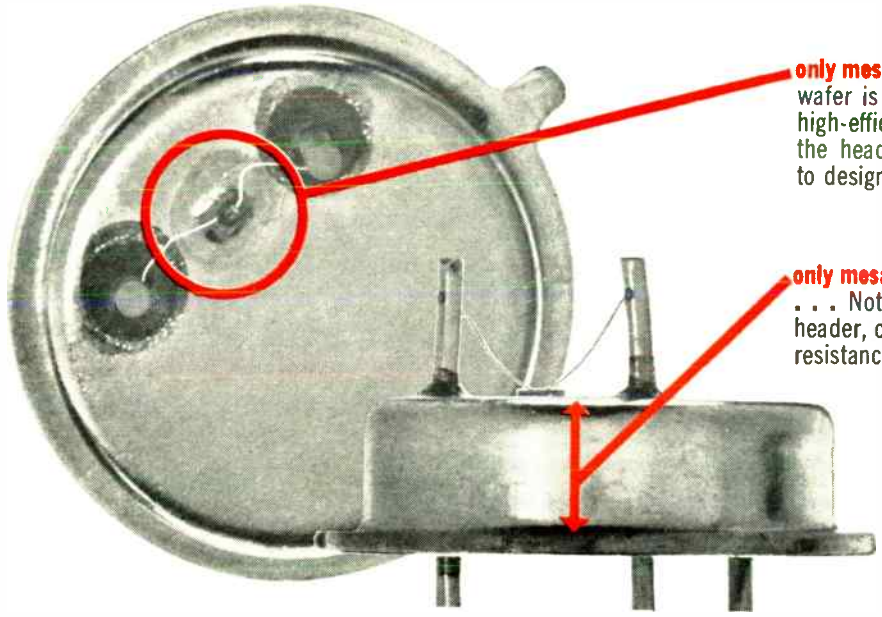
He is a member of the American Institute of Electrical Engineers and the Armed Forces Communications and Electronics Association. He is the author of several papers on microwave radio subjects.

❖

John C. Geist (SM'57) has been named Associate Director at the Silver Spring Laboratory, Vitro's electronic research

(Continued on page 50A)

NEW TI GENERAL-PURPOSE SILICON MESA TRANSISTORS



only mesas give you maximum dissipation . . . Note how wafer is bonded directly to header, forming a direct, high-efficiency metal-to-metal thermal path through the header. High dissipation capabilities permit you to design conservatively for maximum reliability!

only mesas give you maximum mechanical ruggedness . . . Note how active element is bonded directly to header, close to unit's center of gravity—for maximum resistance to vibration and shock.

TI 2N1564 series GUARANTEES -55°C beta, 600-mw dissipation and gain at 30mc

ACTUAL SIZE

Design now with industry's first small-signal silicon mesa transistors... the new TI 2N1564-series! Take advantage of guaranteed −55°C betas of 12, 20 and 40 . . . guaranteed 600-mw free-air dissipation . . . guaranteed current gain at 30 mc. Apply the design flexibility of 1 to 50 ma collector current operating range; 20-50, 40-100 and 80-200 beta spreads at 25°C and 60-v collector-emitter breakdown voltage to your audio, medium-power and higher frequency amplifier and switching designs . . . Specify the new TI 2N1564-series.

absolute maximum ratings at 25°C ambient (unless otherwise noted)

Collector-Emitter Voltage (see note 1)	60 v
Emitter-Base Voltage	5 v
Total Device Dissipation at 25°C Case Temperature (see note 2)	1.2 w
Total Device Dissipation at 25°C Ambient Temperature (see note 3)	0.6 w
Collector Junction Temperature	175°C
Storage Temperature Range	−65°C to +200°C

Note 1: The voltage at which h_{FB} approaches one when the emitter-base diode is open circuited. This value can be exceeded in applications where the dc circuit resistance (R_{BE}) between base and emitter is a finite value.
Note 2: Derate linearly to 175°C case temperature at the rate of 8.0 mw/°C.
Note 3: Derate linearly to 175°C ambient temperature at the rate of 4.0 mw/°C.

Available TODAY in production quantities through all TI Sales Offices and Authorized TI Distributors.

	Parameter	Test Conditions		2N1564 Min.	2N1564 Typ.	2N1564 Max.	2N1565 Min.	2N1565 Typ.	2N1565 Max.	2N1566 Min.	2N1566 Typ.	2N1566 Max.	Unit
I_{CBO}	Collector Reverse Current	V_{CB} = 40 v	I_E = 0			1			1			1	μa
BV_{CBO}	Collector-Base Breakdown Voltage	I_C = 10 μa	I_E = 0	80			80			80			volt
BV_{CEO}*	Collector-Emitter Breakdown Voltage	I_C = 10 ma	I_E = 0	60			60			60			volt
h_{fe}	A-C Common-Emitter Forward Current Transfer Ratio	V_{CE} = 5 v, f = 1 kc	I_E = −5 ma	20		50	40		100	80		200	
		V_{CE} = 5 v, T_A = −55°C	I_E = −5 ma, f = 1 kc	12			20			40			
		V_{CE} = 5 v, f = 30 mc	I_E = −5 ma	1	4		2	4.5		2	5.0		

the FIRST silicon transistor manufacturer

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

IRE People

(Continued from page 48A)

and development facility in Silver Spring, Md.

J. C. Geist

He has been with Vitro since 1948 and has come up through the ranks to his present position in top management. He has directed technical work at all levels from project leader to department head, and before his recent promotion was Technical Operations Manager responsible for managing all of the Laboratory's technical work and for integrating projects assigned to various technical departments.

Before joining Vitro in 1948, he was an engineer in charge of telemetering development at the Applied Physics Laboratory, The Johns Hopkins University, Baltimore, Md.

From 1942 to 1946, he was a contracting officer in the U. S. Army Signal Corps and was responsible for technical aspects of radio and radar equipment procurement. He held the rank of Major in 1946.

He is a specialist in communications systems design, particularly in the application of VHF and FM equipment. He holds a patent on the design of a frequency modulation radio transmitter. Also, he is experienced in television servicing, color television system techniques, and missile telemetry.

During the last eight years he has published more than a dozen technical articles in *Electronics*, *Popular Electronics*, *Radio and Television Service Dealer*, *Service*, and *QST*.

Recently, he won second prize in an essay contest sponsored by the Professional Group on Engineering Management, Institute of Radio Engineers' Washington Chapter. Subject of the contest was "Problems the Engineering Manager Will Face in the Next Decade."

Mr. Geist is a Registered Professional Engineer. He received the B.S.M.E. degree in 1937 from the University of Delaware, Newark.

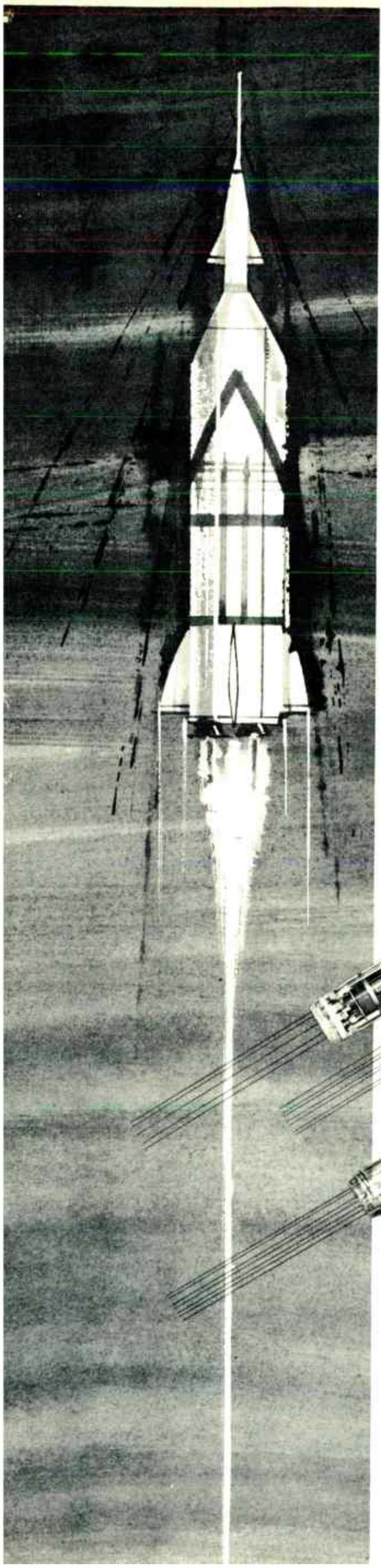
❖

Robert J. Gleason (A'35–M'39–SM'43) has been elected Executive Vice President of Aeronautical Radio, Inc. and of ARINC Research Corporation, both of Washington, D. C. He has been Vice President of Aeronautical Radio, Inc. since 1951, after joining that company in 1949 as Director, Operations, and is also an elected member of the Board of Directors of both companies.

R. J. Gleason

(Continued on page 56A)

ELECTRON TUBE NEWS ...from SYLVANIA



SYLVANIA
GOLD BRAND SUBMINIATURE TUBES

add a high degree of reliability to your critical designs

...new tests prove it!

New manufacturing techniques *build* reliability *into* Sylvania Gold Brand Subminiature Tubes.

New survival rate criteria provide *quantitative definition of Subminiature Tube reliability,* aid designer compute reliability of end-equipment.

New—four Gold Brand Subminiature types—featuring *rugged-design heater for 26.5V applications*—increase versatility of line, widen designer's choice.

SYLVANIA INCREASES SHOCK TEST LEVELS!

• 750g for Gold Brand Premium Subminiature Types
• 1000g for Gold Brand Guided Missile Subminiature Types

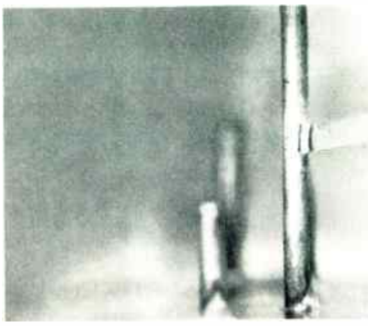
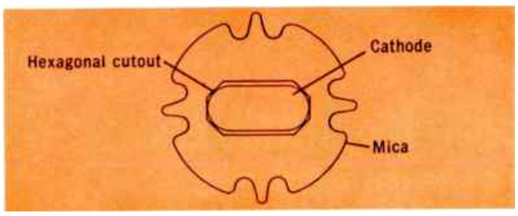


Photo shows the result of Sylvania advanced welding methods. Weld area is extremely rugged and free of weld splatter and oxidation. As a result, catastrophic failures under severe environmental conditions are minimized.



Hexagonal cutout in mica provides firm 6-point contact with cathode, offers increased resistance to shock.

Sylvania has significantly improved the design and manufacture of subminiature type tubes. Now, Gold Brand Subminiature Tubes are capable of withstanding greatly increased impact acceleration tests. For example, newly designed micas provide tight 6-point contact with the cathode. A reducing welding method produces an exceptionally sturdy, clean weld area. Special flared-lip envelopes assure that mica points are not damaged in insertion, maintain the tube structure rigidly within the bulb.

In addition to the increased shock of 1000g applied to Guided Missile Subminiature Tubes, the shock intensity pattern has been changed by eliminating the usual ½″ synthetic rubber pad between the hammer and striking plate of the high impact machine. *Although shock tests are increased, rigid control of end points has not been relaxed.*

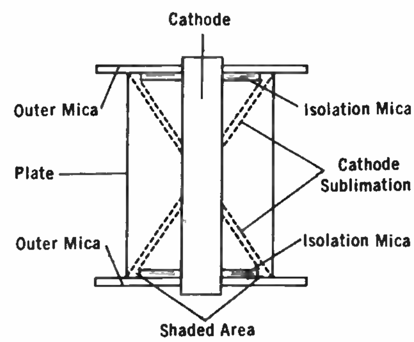
Too, low-frequency vibration tests assure low signal to noise ratio. Vibration tests for "random" or "white" noise are made over a frequency range of 100 to 5000 cps and read up to 10,000 cps to control harmonics. Additional checks include tests for low voltage stability and fatigue.

SYLVANIA INCREASES LIFE TESTS TO 1000 HOURS! NEW CONTROLS ADDED TO 100-HOUR TEST!

Now, Sylvania Gold Brand Subminiature Tubes are tested for 1000 as well as 500 hours. They must meet the same tight limits at 500 and 1000 hours for such end points as: inoperatives, grid current, filament current, Gm, heater-to-cathode leakage, electrical insulation, and cathode interface impedance.

These end points are controlled during manufacture by such operations as: chemically etching the cathode sleeve to provide a good bonding surface for the cathode coating which helps reduce interface impedance, provides improved electrical levels, especially at reduced voltage conditions; use of isolation micas to increase insulation resistance; coating the inside of the cathode sleeve with a nonconductive material to minimize heater-to-cathode leakage.

Further controls are included in the 100-hour life test to assure early-hour stability. For example, new specifications are added for grid current, heater-to-cathode leakage and insulation resistance. The 100-hour life test is performed at room temperature—a critical level for cathode sublimation and resultant leakage paths—and on concurrent samples at various operating temperatures.



Isolation mica "shades" outer mica from sublimation, forms laminated path and greatly increases dc resistance of leakage paths.

SYLVANIA TIGHTENS GLASS AQL LIMITS!

Sylvania has lowered the Acceptance Quality Level from 6.5% to 4% for combined glass defects. Individual glass defects must now meet a 1.5% AQL. This is made possible by increased manufacturing controls to maintain strain-free glass envelopes. Strains that may occur in manufacture are eliminated by annealing glass of Gold Brand Subminiature Tubes after envelopes are sealed. "After-manufacture" annealing is made possible by a special process that keeps the tube structure relatively cool during the annealing. Gold Brand Guided Missile Subminiature types utilize high-resistivity glass. Tubes are capable of withstanding operating temperatures of 250°C, electrolysis caused by heat is virtually eliminated.

HEATER TEST AT ELEVATED VOLTAGE ASSURES FAST WARM-UP TIME!

Sylvania Gold Brand Subminiature Tubes with 6-volt heaters are sample-tested at a heater voltage of 10 volts and a peak heater-to-cathode voltage of 150 volts—cycled 10 seconds "on" and 4 minutes "off" for a total of 300 cycles. In addition, all Gold Brand Subminiature types are tested at normal heater voltages cycled 1 minute "on," 4 minutes "off" for 2000 cycles. To more closely correspond to equipment variations, heaters are designed to operate in a wider voltage range. Ratings for heater voltage variations have been increased from ±5% to ±10%.

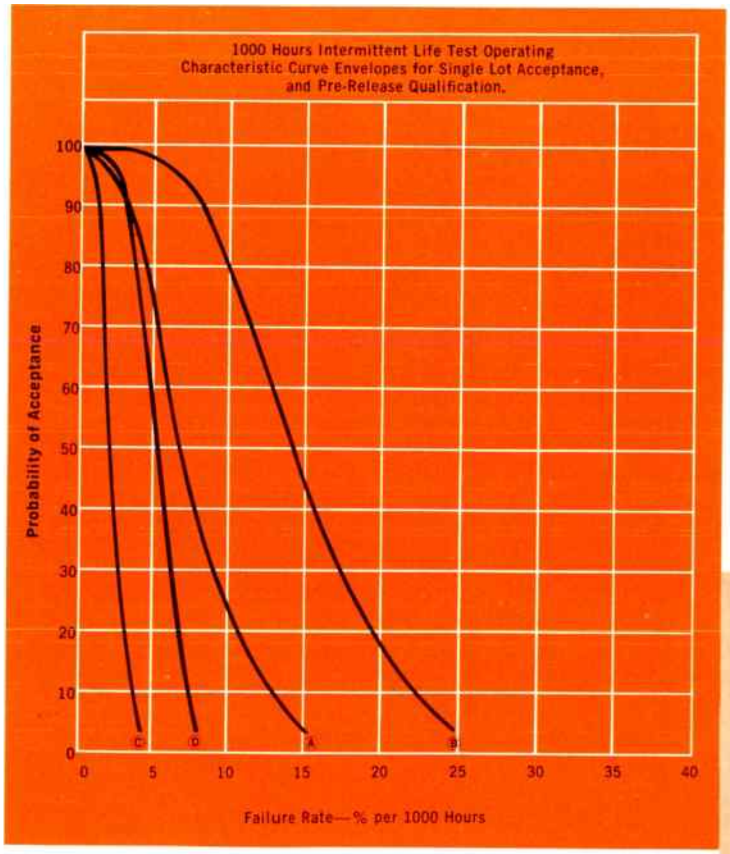
SYLVANIA ADDS INTENSE RADIATION TESTS

Gold Brand Subminiature Tubes are capable of withstanding radiation dose rates (fast neutrons) of 10^{12} NV and accumulated radiation of 10^{16} NVT – further proof of Gold Brand reliability under the most severe environmental conditions.

SYLVANIA "GLEAM PROJECT" INCREASES TUBE RELIABILITY

Initiated 15 years ago, "Gleam" is contributing to Gold Brand Subminiature Tube reliability by – welding in a reducing atmosphere to eliminate weld splatter and oxidation • use of special flared-lip bulbs to allow easy insertion of tube structure into bulb without damaging and flaking mica points • ultrasonic cleaning of critical parts • specially processed getter material which resists flaking • air-conditioning in factories • lint-free clothing, enclosed cloakrooms • individual hooded worktables • lint-free parts containers • microscopic examination of completed tubes for loose particles

SYLVANIA INITIATES NEW SURVIVAL RATE CRITERIA ON GOLD BRAND SUBMINIATURE TUBES!



Sylvania rigorous acceptance criteria is based on the average number of *cumulative* failures for a *five*-lot moving average—instead of one—tested for 1000 hours. The first five lots are tested and the cumulative number of inoperatives and combined failures are plotted with their respective bogey rates. Inoperatives and failures for the sixth lot are added to the cumulative figure and the first lot figures deleted. Sampling consists of 40 tubes per lot. The result is a more stringent control over a wide range of production as well as giving the customary lot by lot results. Too, percent failure rate in 1000 tube hours can be statistically predicted with a high degree of accuracy and provide a quantitative measure of reliability.

	Acceptance Numbers for all Sample Sizes: Inoperatives	Total incl. Inops.
Single Lot	2	5
Five-Lot Moving Sum	5	14
Pre-Release at 500 Hours:		
Five-Lot Moving Sum at 1000 Hours	4	12
Current Lot at 500 Hours	1	2

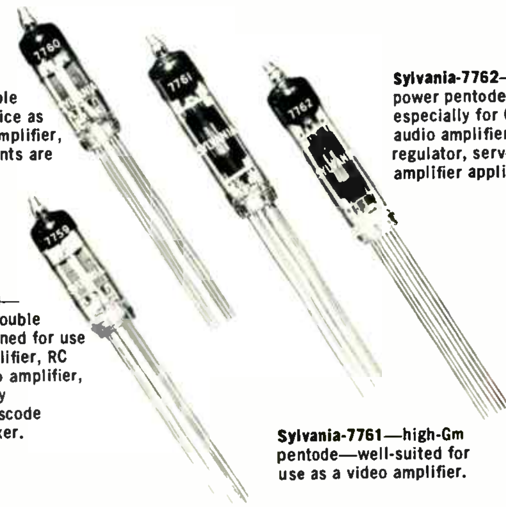
Base Scale for Exemplary Curves Shown Relates to

Single Lot Acceptance	AFR	IFR	RFR
(A) Single Lot for Inops.: n=40, c=2	2.0	6.5	13
(B) Single Lot for Total: n=40, c=5	6.6	14.0	22
Pre-Release Qualification			
(C) Five-Lot Moving Sum for Inops. at 1000 hours and current lot at 500 hours: n=200, c=4 and n=40, c=1	.80	2.0	3.3
(D) Five-Lot Moving Sum for Total at 1000 hours and current lot at 500 hours: n=200, c=12 and n=40, c=2	2.4	5.0	7.2

SYLVANIA ANNOUNCES 4 NEW GOLD BRAND SUBMINIATURE TYPES FOR 26.5 VOLT APPLICATIONS

These remarkable new Gold Brand Subminiature Tubes utilize a rugged-design heater that combines very low heater power with excellent mechanical strength. A heavy mandrel coated with a high-temperature insulator forms the base of the heater. A fine heater wire is wound over the coating and the entire assembly recoated to form a sturdy, efficient, folded coil heater. Your Sylvania Sales Engineer has complete technical data on all four types.

Average Characteristics and Typical Operation	7759 (Each Section)	7760 (Each Section)	7761 Class A Video Amplifier	7762 Class A1 (Single Tube)	Unit
Plate Voltage	100	26.5	—	110	Vdc
Plate Supply Voltage	—	—	200	—	Volts
Cathode Resistor	0.15	—	0.1	0.27	Megohms
Grid Resistor	—	2.2	0.47	—	Megohms
Plate Current	6.5	3.0	—	—	mAdc
Transconductance	5400	5000	—	—	μmhos
Amplification Factor	35	20	—	—	—
Grid Voltage for Ib=100μAdc Max.	—6.5	—	—	—	Vdc
Grid Voltage for Ib=50μAdc	—	—3.5	—	—	Vdc
Grid #2 Voltage	—	—	100	110	Volts
Signal Voltage (rms)	—	—	1.6	6.4	Volts
Zero Signal Plate Current	—	—	19	30	mAdc
Max. Signal Plate Current	—	—	18.5	29	mAdc
Zero Signal Grid #2 Current	—	—	4.0	2.2	mAdc
Max. Signal Grid #2 Current	—	—	4.5	5.5	mAdc
Voltage Output (Peak to Peak)	—	—	135	—	Volts
Load Resistance	—	—	4.7	3.0	Megohms
Power Output	—	—	—	1	Watts
Total Harmonic Distortion	—	—	—	10	%



Sylvania-7760—medium-mu double triode—for service as an RC coupled amplifier, mixer. All elements are 26.5 volts.

Sylvania-7762—beam power pentode—especially for Class A audio amplifier, series regulator, servo-amplifier applications.

Sylvania-7759—medium-mu double triode—designed for use as a UHF amplifier, RC coupled audio amplifier, low frequency oscillator, cascode amplifier, mixer.

Sylvania-7761—high-Gm pentode—well-suited for use as a video amplifier.

GOLD BRAND PREMIUM SUBMINIATURE TYPES for 26.5-Volt Applications

Type	Description
5903	UHF Double Diode
5904*	UHF Medium-Mu Triode
5905*	UHF Sharp Cutoff Pentode
5906	UHF Sharp Cutoff Pentode
5907*	UHF Remote Cutoff Pentode
5908*	UHF Pentode
5916	Dual-Control
7759	Medium-Mu Double Triode
7760*	Medium-Mu Double Triode
7761	High Gm Video Pentode
7762	Beam Power Pentode

*All elements 26.5 volts

GOLD BRAND PREMIUM SUBMINIATURE GUIDED MISSILE TYPES

Type	Description
6943	Sharp Cutoff RF Pentode
6944	Semi-Remote Cutoff RF Pentode
6945	AF Beam Power Pentode
6946	Medium-Mu Triode
6947	Medium-Mu Double Triode
6948	High-Mu Double Triode
6788	Sharp-Cutoff AF Pentode

GOLD BRAND PREMIUM SUBMINIATURE TYPES

Type	Description
5636	Dual Control Pentode
5639	Video Pentode
5641	Diode
5643	Tetrode Thyratron
5644	Cold Cathode Diode
5647	UHF Diode
5718	UHF Medium-Mu Triode
5719	High-Mu Triode
5840	UHF Sharp Cutoff Pentode
5896	UHF Double Diode
5899	UHF Semi-Remote Cutoff Pentode
5902	Beam Power Pentode
5977	Medium-Mu Triode
5987	Low-Mu Power Triode
6021	Medium-Mu Double Triode
6110	UHF Double Diode
6111	Medium-Mu Double Triode
6112	High-Mu Double Triode
6205	UHF Sharp Cutoff Pentode
6206	UHF Semi-Remote Cutoff Pentode
6308	Cold Cathode Diode
6352	Double Diode
6814	Medium-Mu Triode
7327	Medium-Mu Double Triode (Pulse Tube)
7550	Medium-Mu Double Triode (Pulse Tube)

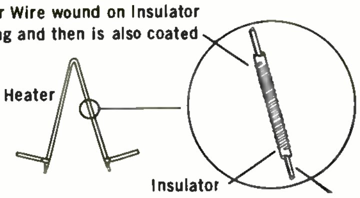


Diagram shows enlarged view of rugged new 26.5-Volt heater for Gold Brand Subminiature Tubes.

Gain the benefits of Gold Brand Subminiature Tubes in your military and industrial designs. Call your nearest Sylvania Field Office for the new specifications and delivery information. For data on individual types, write Electronic Tubes Division, Sylvania Electric Products Inc., Dept. H, 1100 Main St., Buffalo, N. Y.

See the Sylvania Exhibit at Wescon—Booth #2009-2011, 2058-2061, 2108-2111.

SYLVANIA

Subsidiary of **GENERAL TELEPHONE & ELECTRONICS**

CONVECTION COOLED GUARANTEED 5 YEARS

LAMBDA Transistorized Power Supplies

LA Series

5 and 10 AMP • 0-34 VDC

3½" Panel Height on 5 AMP Models

CONDENSED DATA ON LA SERIES

LA 50-03AM	(with meters)	0-34 VDC, 0-5A	$425.
LA 100-03AM	(with meters)	0-34 VDC, 0-10A	540.
LA 50-03A	(without meters)	0-34 VDC, 0-5A	395.
LA 100-03A	(without meters)	0-34 VDC, 0-10A	510.

MODEL	VOLTAGE STEPS
LA 50-03A, LA 50-03AM	2, 4, 8, 16 and 0-4 volt vernier
LA100-03A, LA100-03AM	2, 4, 8, 16 and 0-4 volt vernier

Regulation: *Line* Better than 0.15 per cent or 20 millivolts (whichever is greater). For input variations from 100-130 VAC. *Load* Better than 0.15 per cent or 20 millivolts (whichever is greater). For load variations from 0 to full load.

AC Input: 100-130 VAC, 60 ± 0.3 cycle. This frequency band amply covers standard commercial power lines in the United States and Canada.

Ripple and Noise: Less than 1 millivolt rms.

Ambient Temperature: 50°C—continuous duty.

Remote DC Vernier: Provision for remote operation of DC Vernier.

Remote Sensing: Provision is made for remote sensing to minimize effect of power output leads on DC regulation, output impedance and transient response.

Size:

LA 50-03A	3½" H x 19" W x 14⅜" D
LA 100-03A	7" H x 19" W x 14⅜" D

LT Series

1 and 2 AMP • 0-32 VDC

Compact 3½" Panel Height

CONDENSED DATA ON LT SERIES

LT 1095M	(with meters)	0-32 VDC, 0-1 AMP	$315.
LT 2095M	(with meters)	0-32 VDC, 0-2 AMP	395.
LT 1095	(without meters)	0-32 VDC, 0-1 AMP	285.
LT 2095	(without meters)	0-32 VDC, 0-2 AMP	365.

MODEL	VOLTAGE BANDS
LT 1095, LT-1095M	0-8, 8-16, 16-24, 24-32
LT 2095, LT-2095M	0-8, 8-16, 16-24, 24-32

Regulation: *Line* Better than 0.15 per cent or 20 millivolts (whichever is greater). For input variations from 105-125 VAC. *Load* Better than 0.15 per cent or 20 millivolts (whichever is greater). For load variations from 0 to full load.

AC Input: 105-125 VAC, 50-400 CPS.

Ripple and Noise: Less than 1 millivolt rms.

Ambient Temperature: 50°C—continuous duty.

Remote DC Vernier: Provision for remote operation of DC Vernier.

Remote Sensing: Provision is made for remote sensing to minimize effect of power output leads on DC regulation, output impedance and transient response.

Size:

LT 1095	3½" H x 19" W x 14⅜" D
LT 2095	3½" H x 19" W x 14⅜" D

SEND TODAY FOR COMPLETE DATA.

LAMBDA ELECTRONICS CORP.
11-11 131 STREET • DEPT. 3 • COLLEGE POINT 56, N. Y. • INDEPENDENCE 1-8500
See Us at Wescon. Booths 2114 & 2115

LA110

IRE People

(Continued from page 50A)

Prior to joining Aeronautical Radio, Inc., he was the Communications Superintendent for Pan American World Airways, Inc. (Alaska Division, 1932–1941; Pacific-Alaska Division, 1945–1947; Latin American Division, 1947–1949).

Military duty included service as Army Airways Communications Service Regional Control Officer for Alaska; Commanding Officer, 1st AACS Tactical Group serving 20th Bomber Command India-China; and Commanding Officer 63rd AACS Group in China. He was awarded the Bronze Star Medal.

Mr. Gleason received the B.S.E.E. degree from the University of Washington, Seattle, in 1931.

❖

Election of new vice presidents has been announced by R. E. Krafve, president of Raytheon Company, following a meeting of the board of directors.

Among the new vice presidents are **Fritz A. Gross** (A'36–VA'39) of Weston, Mass., assistant general manager of the Equipment Division; and **Thomas L. Phillips** (M'51) of Wayland, Mass., assistant general manager of the Missile Systems Division.

Mr. Gross, one of the nation's outstanding authorities in the radar field, is a 26-year veteran with Raytheon. His leadership in development of the first microwave search radar was a major factor in the American naval victory at Guadalcanal. American ships were the first afloat to have this radar and thus, the first to be able to "see" at night. For this and his accomplishments in development of fire control radar, he was awarded the U. S. Navy's Certificate of Merit Commendation.

He has served in a number of managerial and engineering positions. Following the war, he headed the group which designed the Mariners Pathfinder radar for commercial vessels. Later, after serving as manager of the division's Wayland, Mass. laboratory, he was promoted to what is now Surface Radar and Navigation Operations of the division. In October, 1959, he was appointed to his present post.

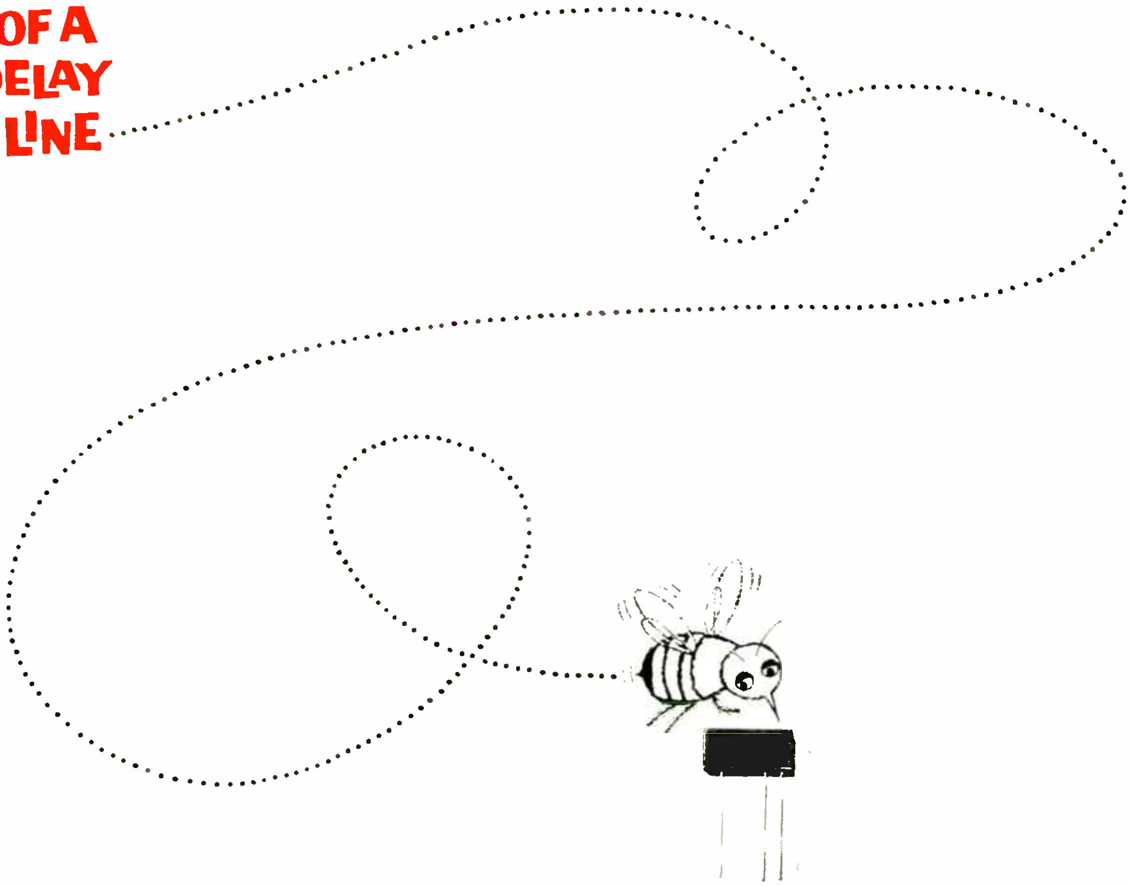
Mr. Phillips, who joined Raytheon in 1948, held positions in design of servomechanisms, special radar circuits and systems until his appointment as manager of the Missile Systems Department in 1953.

In 1955, he was appointed assistant laboratory manager for the division and two years later was named manager of the division's Bedford Laboratories, a post he continued to hold while serving as assistant general manager of the division. Outstanding work on the Navy's Sparrow III missile system, for which Raytheon is prime contractor, won for him the Navy Meritorious Public Service Citation in 1958.

❖

(Continued on page 60A)

A HONEY OF A DELAY LINE



ESC'S NEW SUBMINIATURE LUMPED CONSTANT DELAY LINE*

Model 16-92 is the latest example of creative versatility from ESC, America's largest producer of custom-built and stock delay lines. The specifications: 1/10 usec. delay, 1,600 ohm impedance, ¼" x ¼" x ½" dimensions. Only ESC produces so many different delay lines, for so many varied applications. From the largest to the smallest, ESC has the best, most economical answer to **your** particular delay line problem. Write today for complete technical data.

**shown actual size.*

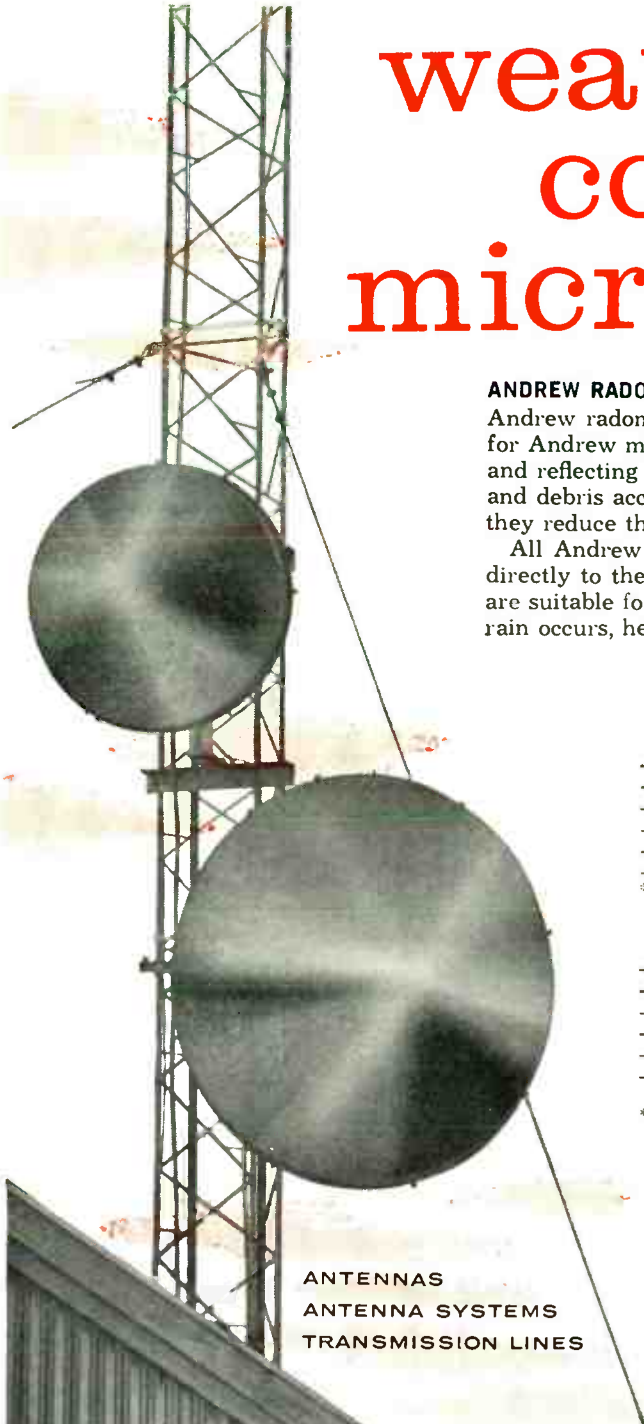
See You at the Wescon Show — Booth # 906

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

ESC ELECTRONICS CORP. 534 Bergen Boulevard, Palisades Park, New Jersey

Distributed constant delay lines • Lumped constant delay lines • Variable delay networks • Continuously variable delay lines • Step variable delay lines • Shift registers • Video transformers • Filters of all types • Pulse-forming networks • Miniature plug-in encapsulated circuit assemblies

weather conditioned microwave



ANDREW RADOME EQUIPPED ANTENNAS DEFY ICE...SNOW...WIND

Andrew radomes provide excellent 2-way year-round protection for Andrew microwave antenna systems. First, they protect feed and reflecting surface against the attenuating effects of snow, ice and debris accumulation. Secondly, for tower mounted antennas they reduce the effects of wind thrust by 35%.

All Andrew radomes are lightweight and easy to install—clip directly to the dish rim of existing antennas. Unheated radomes are suitable for all but exceptional cases. In areas where freezing rain occurs, heated radomes can be provided.

SPECIFICATIONS

STANDARD RADOMES

Dia. Feet	Type No.	Attenuation @ 6 kmc. db	VSWR Contribution @ 6 kmc	Thrust at* 30 psf (Flats), lbs.
10	R10	0.4	0.02	1,990
8	R8	0.4	0.02	1,270
6	R6	0.4	0.02	714
4	R4	0.4	0.02	320
2	R2	0.4	0.02	75

*Including antenna

HEATED RADOMES

Dia. Feet	Type No.	Attenuation @ 6 kmc. db	VSWR Contribution @ 6 kmc.	Thrust at* 30 psf. (Flats), lbs.	Power** Reqmts.
10	HR10	0.7	0.02	1,990	3,400 watts
8	HR8	0.7	0.02	1,270	2,400 watts
6	HR6	0.7	0.02	714	1,200 watts
4	HR4	0.7	0.02	320	550 watts
2	HR2	0.7	0.02	75	150 watts

*Including antenna

**Power requirements for HR10 and HR8 are 3 wire single phase 60 cycle 220 volts.

Power requirements for HR6, HR4 and HR2 are single phase 60 cycle 115 v.

For further details on ANDREW Microwave Antennas, Radomes, Wave Guides write for new Andrew Catalog M.

ANTENNAS
ANTENNA SYSTEMS
TRANSMISSION LINES

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"We have paid particular attention to antennas during high wind conditions of gusts up to 40-60 m.p.h. It is very obvious that these radomes quite materially reduce the wind loading on the parabolas—due to their shape factor." *Washington State Patrol, Kennewick, Washington*

"We have had up to four inches of ice on the radome with practically no reduction of antenna effectiveness." *KLIX-AM-TV, The KLIX Corporation, Twin Falls, Idaho*

"Our field forces report that the radomes produce a signal loss of less than 1 db per antenna. Several radomes were removed and antennas inspected following a heavy snow storm and no snow or ice was found in the antennas." *Natural Gas Pipeline Company of America*

IRE People

(Continued from page 56A)

The Military Systems-Stavid Division of Lockheed Electronics Company has appointed **Donald L. Gunter** (M'47) to the new post of Director of Marketing. He will be responsible for sales, contract administration, market planning, and customer relations activities for the division.

D. L. Gunter

A native of West Virginia, the new marketing head was previously employed for 12 years by RCA. His last position was Manager of Market Development for the Astro-Electronic Products Division. He also served as Manager of Planning, Special Systems and Development Department; Manager of Defense Projects Coordination; Administrator of Guided Missile Projects; Manager of Air Force Contracting; Manager of Army Contracting; and Manager of Research Contracts.

Other positions throughout his careers include Contract Representative for the Applied Physics Laboratory of Johns Hopkins University, Silver Spring, Md.; Project Engineer for the Bureau of Ships, Washington, D. C.; and Power Sales Engineer for Appalachian Electric Power Company, Charleston, W. Va. He holds the B.S.E.E. degree from Virginia Polytechnic Institute, Blacksburg, and attended Navy Radio and Radar Schools at Virginia Polytechnic Institute, Harvard and Massachusetts Institute of Technology.

Mr. Gunter is affiliated with Tau Beta Pi, Phi Kappa Phi, American Ordnance Association, American Society of Naval Engineers, U. S. Naval Institute, American Rocket Society, and the Armed Forces Communications and Electronics Association.

❖

Foto-Video Electronics, Inc., of Cedar Grove, N. J., announces appointment of **Robert D. Hamilton** (M'58) formerly consulting engineer with International Business Machines Corporation, at Oswego, N. Y., as Head of the Systems Engineering Department at the Company's expanding Cedar Grove operations.

R. D. Hamilton

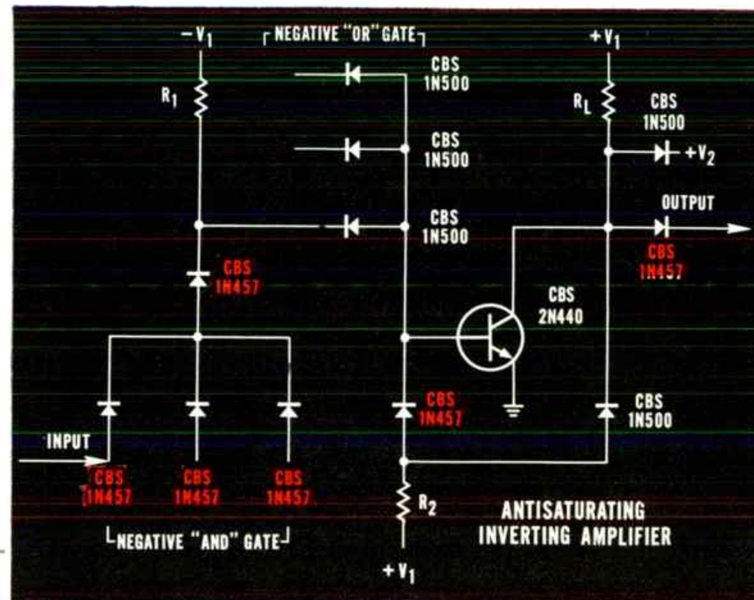
Mr. Hamilton comes to Foto-Video with a wide background in mechanical and electronic design and systems engineering.

While serving IBM, he was in charge of planning and specification writing for the Bomb-Nav System Test involving Sighting

(Continued on page 62A)

NEW CBS DIFFUSED SILICON DIODES

featured in more efficient economical switching block



This single building block for computer switching achieves increased efficiency, flexible cascading, and simple maintenance. New CBS high back-resistance diffused silicon diodes used in the "And" gate assure positive switching. The relatively large voltage drop developed by these current switching devices drives the phase inverter transistor efficiently at high switching speeds, and minimizes cooling problems.

New CBS high back-resistance diffused silicon diodes for positive switching now join CBS high-conductance and fast-recovery types. Efficient and flexible switching is made possible (see circuit). CBS diffusion techniques offer three major advantages over the alloying method: Close process control of all parameters, great uniformity, and high reverse voltage through the graded junction.

The new CBS 1N456, 1N457, 1N458, 1N459 are particularly designed for efficient computer operation in missiles, rockets, airborne and industrial equipment. Typical applications include: switching, pulse, flip-flop, modulator, demodulator, discriminator, clamping, gating, and detector circuits. Write for data sheet E-387. Order direct, from your local sales office, or MWD distributor.

Check These Characteristics

Type	Min. Rev. Voltage @ 100 µa (volts)	Min. Forward Current		Maximum Reverse Current				Avg. Rect. Fwd. Current
				@ 25°C		@ 150°C		
		I_F (mA)	E_F (volts)	I_R (µA)	E_R (volts)	I_R (µA)	E_R (volts)	(mA)
1N456	−30	40	1.0	0.025	−25	5	−25	90
1N457	−70	20	1.0	0.025	−60	5	−60	75
1N458	−150	7	1.0	0.025	−125	5	−125	55
1N459	−200	3	1.0	0.025	−175	5	−175	40

Other CBS Diffused Silicon Types

Type	Min. Reverse V @ 100 uA	Min. Avg. Forward @ 1V mA @ 25°C	Bulletin
	High Conductance Types		
1N482	−40	100	E-373
1N483	−80	100	E-373
1N484	−150	100	E-373
1N485	−200	100	E-373
	Fast Recovery Types		
1N625	−35	20	E-374
1N626	−50	20	E-374
1N627	−100	20	E-374
1N628	−150	20	E-374
1N629	−200	20	E-374

ADVANTAGES OF CBS 1N456, 1N457, 1N458 AND 1N459

Efficient computer switching
High back resistance
Sharp back-voltage characteristic
Excellent forward conductance
Low current saturation
Wide storage and operating temperature ranges

CBS ELECTRONICS

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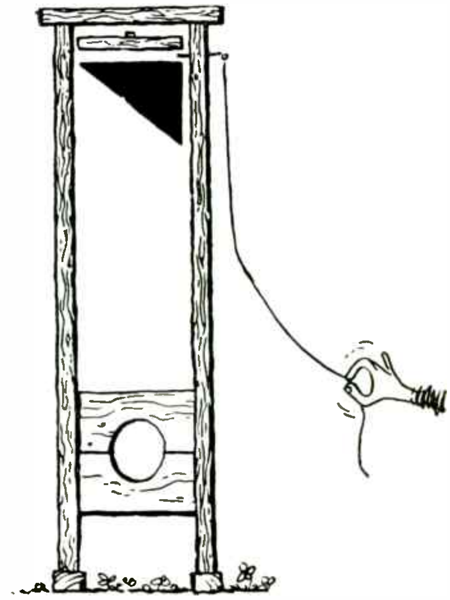
More Reliable Products through Advanced Engineering

CBS® semiconductors



IRE People

(Continued from page 60A)

Radar, Doppler Radar, Data-Processing, Digital-Computing operations, Radar Display and Stellar Inertial operations.

Prior to that he was a senior engineer with Allen B. DuMont Laboratories, holding the post of Electronic Systems Engineer, responsible for the scheduling and writing of creative industrial and military-systems proposals including tactical reconnaissance by television and other communications.

Before holding those posts, he served G. M. Giannini and Company, of Caldwell, N. J., in component engineering; Eclipse-Pioneer Division at Teterboro; Kearfott Company at Little Falls, N. J., and other electronic firms. He served in the U. S. Army Signal Corps overseas and in this country between 1940 and 1945.

Mr. Hamilton is a member of a number of professional societies.

❖

Appointment of **Charles Hittner** (A'45–SM'54), formerly Staff Director of Engineering at Radio Receptor as a Divisional Vice President and Chief Staff Engineer of the Engineering Products Division has been announced by General Instrument Corporation, of which Radio Receptor Company is a subsidiary.

C. Hittner

After joining Radio Receptor in 1945, he held the post of Chief of the Advanced Development Section and Chief Engineer of the Engineering Products Division before being named Staff Director in 1958. In these posts, he supervised the design and development of ground communication equipment widely used by the U. S. Air Force, air traffic control systems for the F.A.A., and vehicular and gun-mounted identification systems used extensively by the U. S. Army Signal Corps. Prior to joining Radio Receptor, he was with Hudson American Corporation.

Mr. Hittner is a graduate of Polytechnic Institute of Brooklyn, Brooklyn, N. Y.

❖

The Semiconductor Division of Sylvania Electric Products Inc. has announced the appointment of **William D. Hogan** (S'50–A'51–M'55) to the newly created post of manager of field engineering.

He has been section head in charge of field engineering since 1958. He joined Sylvania in 1950 as a junior engineer in the division's Engineering Test Laboratory in Boston.

Previously, he was a civil engineer with

(Continued on page 64A)

Philco announces the only MICRO-ENERGY SWITCH

the industry's first LOW ENERGY, HIGH SPEED switching transistor

T-1930 . . . MICRO-ENERGY SWITCH . . . TO-18 CASE

MAXIMUM RATINGS

Storage Temperature	100° C
Total Device Dissipation at 25° C	35 mw

CHARACTERISTICS	MIN.	TYP.	MAX.
DC Current Amplification Factor, h_{FE} ($V_{CE} = -0.20$ v, $I_C = -2$ ma)	25	40	
Collector Voltage, V_{CE} ($I_C = -2$ ma, $I_B = -0.2$ ma)		.095	.13 V
Gain-bandwidth Product, f_T ($V_{CE} = 1$ v, $I_C = 1$ ma)	125	175	mcs

The Philco T-1930 is a new concept in the design of switching transistors for high speed computer logic circuits! All internal device capacities are exceedingly small . . . and its static characteristics are optimized for operation at low collector voltages and collector currents. *It permits the design of high-speed logic circuits with an overall power consumption only 1/3rd to 1/10th that of circuits with conventional transistors.* It will operate at pulse rates in excess of 10 mc with collector currents as small as 1 ma from collector supply voltages as small as 1 V.

This new micro-energy switch is of great importance in the design of ultra-reliable, high density, high speed equipment. In micro-energy circuits, the total device dissipation is reduced to an absolute minimum . . . 250 microwatts . . . a prime consideration in achieving maximum reliability. The T-1930 is an important step toward microminiaturization . . . permitting high packing densities without excessive internal heat generation. For complete information write Dept. IRE-860.

Immediately available from your Philco Industrial Semiconductor Distributor

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IRE People

(*Continued from page 62A*)

the Commonwealth of Massachusetts and with Raytheon Manufacturing Company at Waltham, Mass.

A native of Boston, Mass., Mr. Hogan graduated summa cum laude from the University of Massachusetts, Amherst, Mass., where he received a Bachelor of Science degree in Electrical Engineering. He has also followed courses in industrial electronics at Northeastern University.

❖

Dr. Ralph P. Johnson (SM'58) has been appointed vice president, Electronics Divisions, Thompson Ramo Wooldridge, Inc., according to an announcement by Dr. D. E. Wooldridge, president, TRW Inc.

R. P. Johnson

In his new capacity, which is effective immediately, he will be responsible for activities within TRW's five electronics divisions: Commercial Electronics Group, Electronic Components Group, Pacific Semiconductors, Inc., Ramo-Wooldridge Division, and The Thompson-Ramo-Wooldridge Products Company. He will be located in TRW's west coast headquarters in Canoga Park, California.

He was formerly TRW vice president and general manager of Ramo-Wooldridge, a division of Thompson Ramo Wooldridge, Inc. He will be succeeded as R-W general manager by **Milton E. Mohr** (M'45–SM'53), who was vice president, operations for Ramo-Wooldridge.

Dr. Johnson joined Ramo-Wooldridge in 1954, as director of research and development. Prior to this, he held a similar position with Hughes Aircraft Company for four years. From 1947 to 1950, he was deputy director of the Research Division, U. S. Atomic Energy Commission; and from 1936 to 1947, he was a research physicist with the General Electric Company.

❖

Boonton Electronics Corporation, Morris Plains, N. J. has announced the appointment of **Raymond E. Lafferty** (M'51–SM'56) to the position of Chief Engineer. In his new position he will investigate new applications and improvements for their present line of test equipment, as well as contribute to the development of new equipment. He brings twenty years of experience in broadcasting, and component and instrument design.

R. E. Lafferty

(*Continued on page 67A*)

ARNOLD: WIDEST SELECTION OF MO-PERMALLOY POWDER CORES FOR YOUR REQUIREMENTS

For greater design flexibility, Arnold leads the way in offering you a full range of Molybdenum Permalloy powder cores . . . 25 different sizes, from the smallest to the largest on the market, from 0.260" to 5.218" OD.

In addition to pioneering the development of the cheerio-size cores, Arnold is the exclusive producer of the largest 125 Mu core commercially available. A huge 2000-ton press is required for its manufacture, and insures its uniform physical and magnetic properties. This big core is also available in three other standard permeabilities: 60, 26 and 14 Mu.

A new high-permeability core of 147 Mu is available in most sizes.

These cores are specifically designed for low-frequency applications where the use of 125 Mu cores does not result in sufficient Q or inductance per turn. They are primarily intended for applications at frequencies below 2000 cps.

Most sizes of Arnold M-PP cores can be furnished with a controlled temperature coefficient of inductance in the range of 30 to 130° F. Many can be supplied temperature stabilized over the MIL-T-27 wide-range specification of −55 to +85° C . . . another special Arnold feature.

Graded cores are available upon special request. All popular sizes of Arnold M-PP cores are produced to a standard inductance tolerance of + or −8%, and many of these sizes are available for immediate delivery from strategically located warehouses.

Let us supply *your* requirements for Mo-Permalloy powder cores (*Bulletin PC-104C*). Other Arnold products include the most extensive line of tape-wound cores, iron powder cores, permanent magnets and special magnetic materials in the industry. • Contact ***The Arnold Engineering Co., Main Office and Plant, Marengo, Illinois.***

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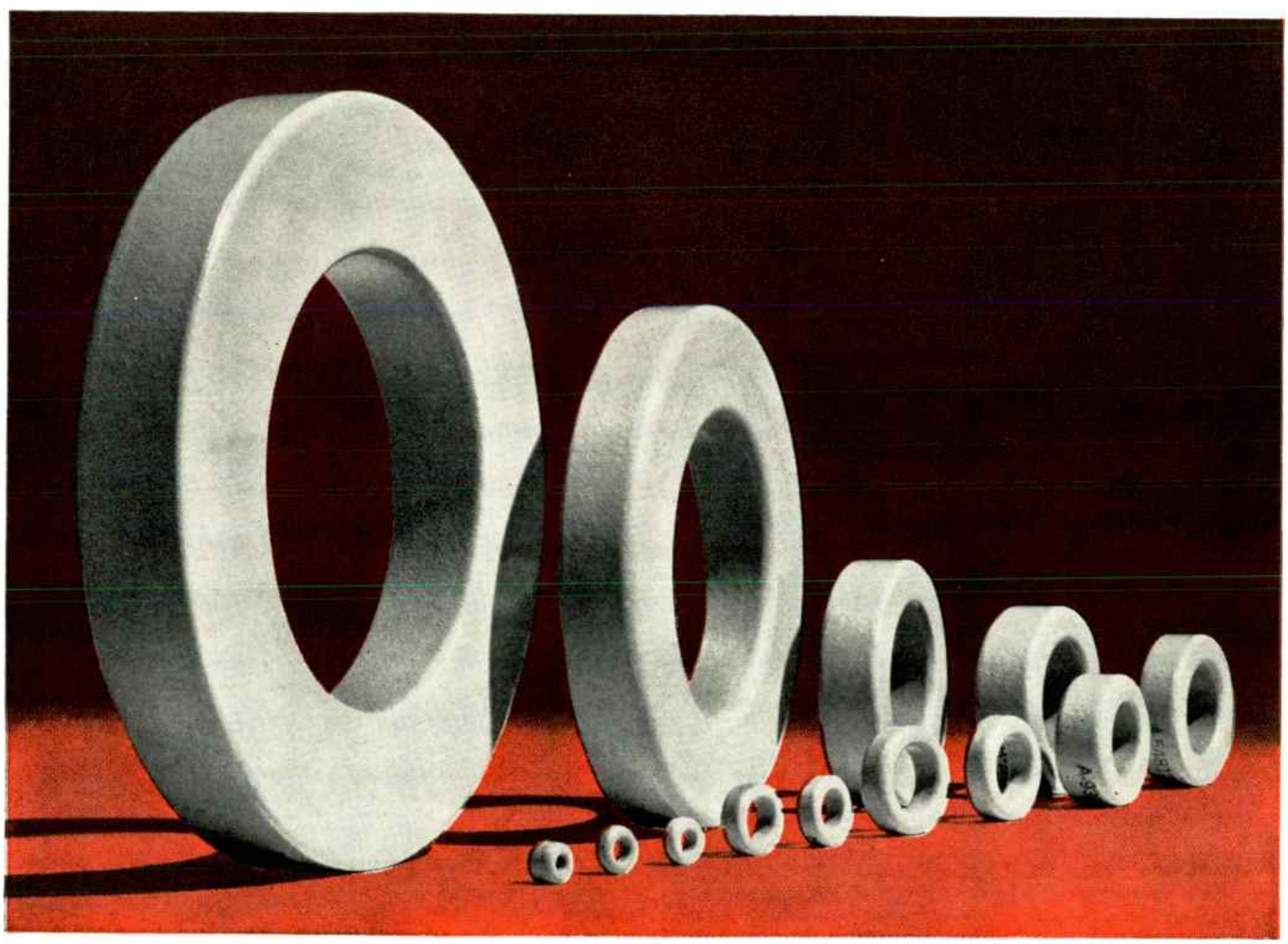
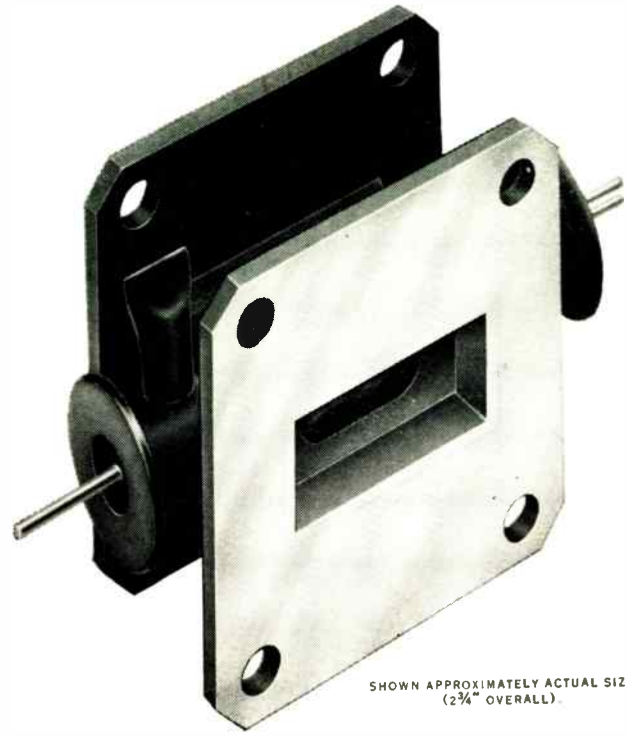


Microwave Component News from SYLVANIA

New broad-band high-speed microwave switch



SHOWN APPROXIMATELY ACTUAL SIZE (2¾" OVERALL)

Now—a superior new device for your switching, duplexing and crystal-protecting applications

SYLVANIA'S RESEARCH in microwave components has resulted in an important new development which overcomes many of the disadvantages of existing microwave switching and duplexing devices. It is a hot-cathode, grid-initiated arc discharge switch with this unique combination of features:

- Suitable for high and low power applications
- Firing time of .1 -.3 microsecond
- Recovery time of the order of 1-10 microseconds
- Band width comparable to full waveguide band
- Consistent, reproducible phase of transmission and reflection

It has these specific advantages:

SWITCHING APPLICATIONS

High-speed   Controlled firing   High isolation   Low loss

DUPLEXING AND CRYSTAL PROTECTION APPLICATIONS

Negligible spike leakage
Insertion loss as low as 0.2 db
Maintains performance at low temperatures
No noise contribution to receiver (no keep-alive current)
Protects against RF power at lower levels than TR's

For engineering samples in C or X bands or information on units in other bands, contact your Sylvania sales office or write to
Sylvania Special Tube Operations
500 Evelyn Avenue, Mountain View, California

SYLVANIA

Subsidiary of **GENERAL TELEPHONE & ELECTRONICS**

GENERAL SYSTEM

(Continued from page 64A)

He was most recently Assistant Chief Engineer for the Daven Company. He joined that organization in 1957 as Section Head in charge of low frequency LC filters and test equipment, including some government projects. His sphere was later enlarged to encompass delay lines and RF attenuator design. In June, 1959 he was made Assistant Chief Engineer.

From 1948 to 1957, he was a Development Engineer for the National Broadcasting Company. In this position he was concerned with receiver and transmitter design, automatic audio gain control amplifiers, magnetic tape, regulated power supplies, synthetic sound effects, and transistorization of portable equipment.

Prior to 1948, he was with WSLB as Chief Engineer, the Boonton Radio Corporation as Project Engineer, and an Instructor with the New York State Signal Corps Schools.

He has contributed over a dozen papers in the field of measurements and delivered a number of talks, including a paper presented at the 1956 IRE Convention.

Mr. Lafferty was a member of the Registration Committee for the IRE National Convention for several years and Chairman of that Committee in 1958 and 1959. He also served as Chairman of the Hospitality Committee for the 1960 IRE International Convention. Mr. Lafferty is a Fellow of the Radio Club of America.

❖

Announcement of the appointment of **Robert D. Lavin** (A'54) as manager of the Systems Engineering Department of Gulton Industries' Ortholog Division was made May 27, 1960, by Dr. Leslie K. Gulton, president of the electronics engineering and manufacturing firm. He will be in charge of the recently-created department's project work in advanced military and industrial instrumentation systems which are being applied in fields such as missile and space vehicular information and control, sonar techniques, and industrial control instrumentation.

R. D. Lavin

Prior to joining Gulton Industries, he had wide experience in systems management and development. Among his previous associations were positions with the RCA Missile Test Project at Patrick Air Force Base, Florida; the Applied Science Corporation of Princeton; and the Ampex Corporation in California. Earlier he served as an instructor in communications for the Air Force.

Mr. Lavin is a member of the Professional Group on Space Electronics and Telemetry. He holds both the Bachelors and Masters degrees from the University of Minnesota, Minneapolis.

❖

(Continued on page 68A)

IRE People

(Continued from page 67A)

Dr. Harry Letaw, Jr. (SM'56) has been named marketing manager for Surface Radar and Navigation Operations of Raytheon Company. He succeeds **A. E. Keleher, Jr.** (M'47), new manager of marketing planning for the electronic firm's Equipment Division.

In his new post, Dr. Letaw will manage marketing efforts for Surface Radar and Navigation Operations, one of five Equipment Division units. The operation is currently developing and producing surface radar and navigation equipment for the military as well as storm and flight tracking radars for the U. S. Weather Bureau and Federal Aviation's Agency's nation-wide network.

He joined Raytheon in 1955 as a senior research staff member of the Research Division. In June, 1958, he was appointed manager of the Equipment Division's Space Technology Department, and in July, 1959, he became manager of systems requirement for the division.

Prior to joining Raytheon he was a research assistant professor in the University of Illinois' Electrical Engineering Department.

He holds the Bachelors and Masters degrees from the University of Florida, Gainesville, and received the Ph.D. degree there in 1952. He held Atomic Energy Commission and Research Corp. fellowships while doing graduate study.

He is a member of the American Physical Society, Phi Beta Kappa, Phi Kappa Phi, and Sigma Xi.

❖

The appointment of **Albert E. Linden** (A'53–SM'57) to the position of Department Staff Engineer was recently announced by Hugh E. Webber, Manager, Ground-Electronics Department of the Martin Company, Orlando, Florida.

A. E. LINDEN

Mr. Linden is best known to IRE members, however, as Chairman of the newly-organized Orlando Section. In early 1959, he was the sparkplug of the Steering Committee which organized the Section, and he personally obtained 100 signatures on the petition for its formation. During his term as Chairman, IRE membership in the area, which six years ago was only 12, rapidly grew to over 325.

He came to Martin-Orlando from the Missile Guidance Section of American Bosch Arma Corporation, where he was Senior Project Engineer in the Ground Equipment Department. There, he supervised the development of equipment for dynamic testing of digital computers associated with the inertial guidance system of the "Atlas" and "Titan" ICBM's. Prior to

(Continued on page 70A)

THE FINEST INSTRUMENT FOR ALL-PURPOSE TESTING . . .

WESTON Model 901 Group Now better than ever

Weston's matched line of Model 901 portables are well known for sustained accuracy and dependability under general test conditions. This modern group of AC and DC multi-range instruments consists of ammeters, voltmeters and wattmeters covering a wide range of measurement.

Designed for critical use, the Model 901 DC series is now accurate to 0.25%. Hand-calibrated mirror scales are combined with knife edge pointers to eliminate parallax errors. Widely-spaced markings on 5.5-inch long scales facilitate readability. The Model 904 AC series is now stocked in multi-range, frequency-compensated versions only.

Excellent for field use, these portables are housed in rugged plastic.

Other features include wide, shadow-reducing windows which are specially treated against electrostatic effects, and self-shielded mechanisms that offer positive protection against external magnetic influences.

Call your Weston representative for complete information, or write for Catalog 06-203. Daystrom, Incorporated, Weston Instruments Division, Newark 12, New Jersey. *International Sales Division, 100 Empire St., Newark 12, New Jersey. In Canada: Daystrom Ltd., 840 Caledonia Rd., Toronto 19, Ontario.*

Weston Model 901 Group consists of: Model 901 DC Instruments; Model 904 AC Instruments; Model 902 AC Rectifier-Types; and Model 905 AC and DC Single Phase Wattmeters. Protective leather carrying cases are available for all models.

DAYSTROM, INCORPORATED
WESTON INSTRUMENTS DIVISION

Reliability by Design

(Continued from page 68A)

this, he was a Senior Electronics Design Engineer for Kaiser Metal Products, Inc., and still earlier, held engineering supervisory positions with National Union Radio Corporation and Raytheon Manufacturing Company. At Kaiser, one of his outstanding developments was a "Precision Digital Delay Generator" which has a measurement accuracy of two millimicroseconds. Also, he was involved in the design of numerical controls for machine tools, "tinkertoy" modules, and advancements in the development of modular printed wiring.

He received his education in Electrical Engineering at Temple University, Philadelphia, Pa., and the University of Pennsylvania, University Park. He has done graduate work at the University of Pennsylvania, Massachussetts Institute of Technology and the Brooklyn Polytechnic Institute.

After his discharge from the service, where he served as a Communications Warrant Officer during World War II, he established his own business devoted to Television Sales and Service. He returned to military electronics at the start of the Korean War. He holds FCC Licenses for Radiotelegraph, Radiotelephone and Amateur Radio operation.

Since joining Martin-Orlando in May of 1958, he has held various supervisory positions within the Ground-Electronics Department. He has been successively: Coordinator of Ground Check-Out Equipment for the Lacrosse, Pershing, and Bullpup guided missiles; Acting Section Head; Chairman of the Research Committee of Electronic Packaging; Member of the Product Review Board; Lecturer at Orlando Junior College; and Chairman of the Committee for Design Coordination of Ground Equipment.

❖

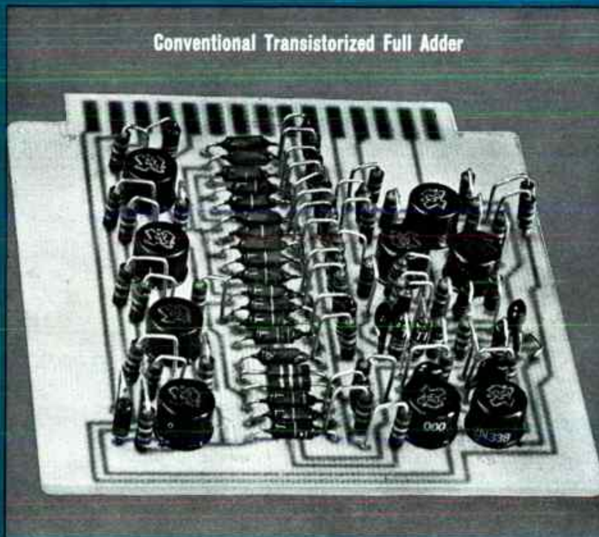
Dr. Leonard C. Maier, Jr. (M'52), has been named manager of engineering for the General Electric Company's Semiconductor Products Department at Electronics Park, Syracuse, N. Y.

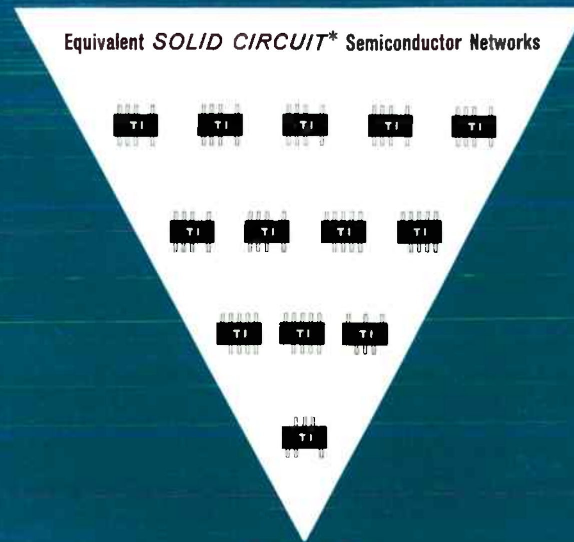
L. C. Maier, Jr.

In his new position, he will direct and be responsible for all semiconductor engineering and advanced research studies which include the Advanced Semiconductor Laboratory, Engineering Administration, Transistor Advance and Design Engineering, Transistor Product Engineering and Semiconductor Metals Engineering.

A native of LaGrange, Ill., he joined G.E. as an engineer in the Communications Systems Unit in 1950. During the next eight years, he served in various supervisory and managerial capacities in

(Continued on page 73A)





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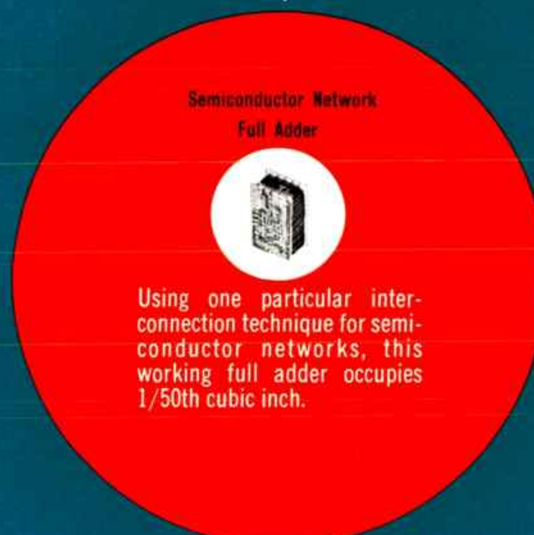
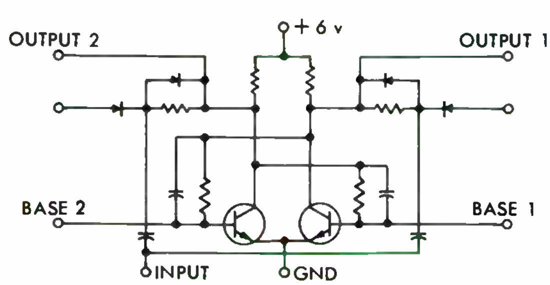
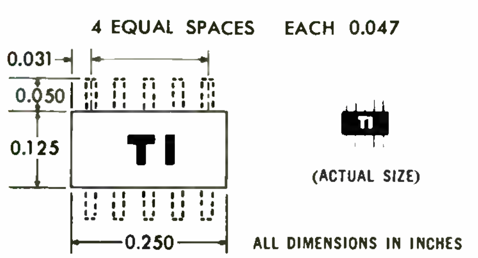


TI Type 502 silicon **SOLID CIRCUIT** network is intended for binary counter, flip-flop, or shift register applications. Maximum repetition rate of 200 kc. Hermetic sealed package. Write for bulletin TI type 502.





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You get improved reliability through the elimination of individual components . . . element interconnections are reduced as much as 80% . . . size and weight are drastically reduced and power consumption is minimized.

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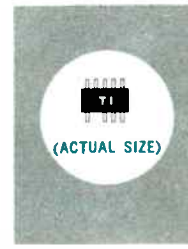
SOLID CIRCUIT*

DIFFUSED SILICON BISTABLE MULTIVIBRATOR NETWORK

A MICROELECTRONIC BINARY FLIP-FLOP



for application in

- **Counter Circuits**
- **Shift-Register Circuits**
- **Reset-Set Flip-Flop Circuits**

maximum ratings at 25°C ambient

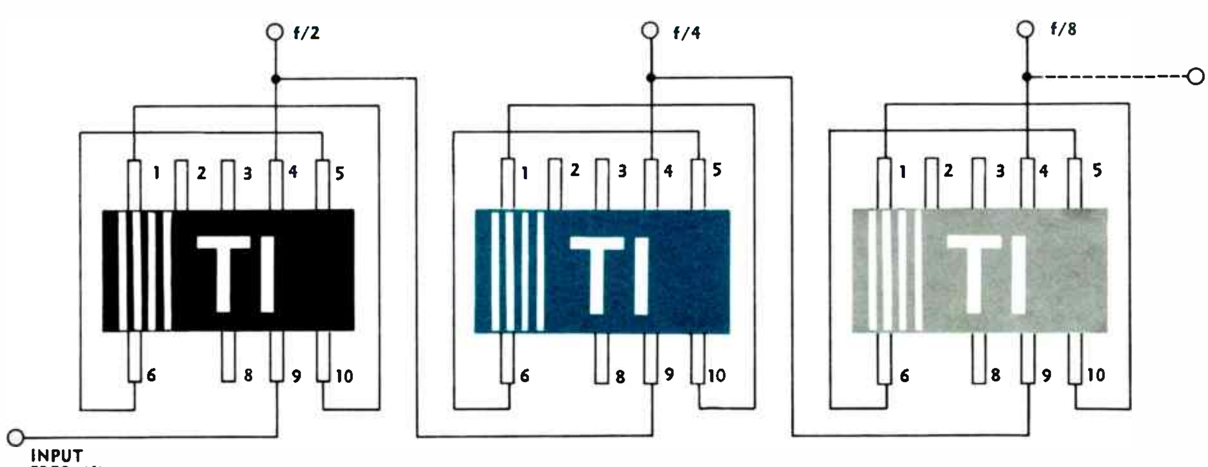
Maximum Supply Voltage	8 v
Maximum Trigger Voltage	8 v
Storage Temperature	— 55° to +125°C

design characteristics:

Design Supply Voltage	6.0 v ± 5%	
Input		
Minimum Pulse Width	1 μsec	
Minimum Trigger Voltage	2.5 volts	
Maximum Trigger Voltage	5.0 volts	
Maximum Trigger Frequency	200 kc	
Maximum Rise Time	1 μsec	
Output Voltages	ON	OFF
With 10K-ohm load to ground	0.7 v max	3.5 v min
With 10K-ohm load to V_s	0.8 v max	4.0 v min
Output Wave Form (typical)		
t_{ON} (90%)	1 μsec	
t_{OFF} (90%)	1 μsec	
Operating Temperature Range	— 40°C to +85°C	

ENGINEERING ASSISTANCE AVAILABLE FOR SOLID CIRCUIT NETWORK APPLICATIONS

TYPE 502 SOLID CIRCUIT* NETWORK CONNECTED AS A BINARY COUNTER



NOTE: SUPPLY VOLTAGE CONNECTIONS NOT INDICATED

* Trademark of TEXAS INSTRUMENTS INCORPORATED

(Continued from page 70A)

the Company's Electronics Laboratory and the Cathode Ray Tube Department.

He is an alumnus of Williams College, Williamstown, Mass., from which he received the B.A. degree in physics in 1944. Following naval service in the V-12 program and later as an electronics maintenance officer in the Pacific theater of operations, he enrolled at the Massachusetts Institute of Technology, Cambridge, for advanced studies.

While at MIT, he served for three years in the school's laboratory of electronics as a research associate and research assistant. He was awarded the M.S. degree from MIT in 1948 and the doctorate in 1949, both in the field of physics.

Dr. Maier is a member of the Research Society of America, American Physical Society, the scholastic honorary Phi Beta Kappa and the physics honorary Sigma Xi. In 1954, he served as program chairman, Syracuse area, of the IRE.

❖

Beryl L. McArdle (SM'50) has been appointed scientific advisor on the staff of Dr. Royal Weller, vice president for engineering of the Stromberg-Carlson Division of General Dynamics Corporation.

B. L. McArdle

He will engage in special studies in chemical and biological warfare, intelligence systems, operations research and countermeasures.

For three years prior to his new appointment, he was manager of the Intelligence Systems Department in Stromberg-Carlson's Electronics Division. He has been with Stromberg-Carlson since 1947, when he joined the company as a senior engineer in the Research Division. Subsequently, he became manager of the division's Data Systems Laboratory.

A native of Galien, Michigan, he received the bachelor's degree in electrical engineering with honor from the Michigan College of Mining and Technology, Houghton. During World War II, he served in the U. S. Signal Corps with the rank of lieutenant, and later was head of the research section on cryptographic communications in the Department of Defense.

Mr. McArdle holds nine patents, several of them classified, and is a member of the Committee on Military Electronic Systems of the Electronic Industries Association.

❖

The appointment of **Howard S. Moncton** (A'39–M'46) as manager of planning and review for the Amherst Laboratory of

(Continued on page 74A)

IRE People

(Continued from page 73A)

Sylvania Electric Products Inc., has been announced by Walter Serniuk, laboratory director.

H. S. MONCTON

He was previously business manager of the Buffalo Operations. Affiliated with Sylvania since 1939, he was assistant to the manager of the company's Physics Laboratory at Bayside, N. Y., and administrative engineer of Sylvania Home Electronics, another division of the company, prior to joining Sylvania Electronic Systems.

A graduate of Union College, Mr. Moncton is a member of Sigma Xi, honorary scientific fraternity, the Society for Advancement of Management, and The American Radio Relay League.

❖

The appointment of **William F. Palmer** (A'51) to the newly created post of manager of advanced circuit development and new product evaluation for the Semiconductor Division of Sylvania Electric Products Inc. has been announced by Dr. J. E. Thomas, director of research and engineering for the division. Sylvania is a subsidiary of General Telephone and Electronics Corporation.

In his new position, Mr. Palmer, who has headed the division's applications engineering section since 1957, will seek new applications for existing semiconductor products and for devices presently under development. He will continue to have his offices at the division's general engineering laboratory.

He joined Sylvania in 1952 as a device development engineer at Newton, Mass. The following year he was placed in charge of the division's advance communications application group. In 1955, he became head of the division's commercial engineering department.

Before joining Sylvania, he was a teaching fellow at Wesleyan University, Middletown, Conn., where he specialized in piezoelectric phenomena and received the Master of Arts degree in 1952.

A native of St. John, N. B., Canada, he attended Mount Allison University, Sackville, N. B., where he received the Bachelor of Science degree in Physics.

The author of more than a score of technical papers on solid state devices and applications which have appeared in trade and technical publications. Mr. Palmer has spoken extensively before professional societies and scientific groups. He is a member of the American Physical Society and the American Association of Physics Teachers. He has also served as chairman, vice chairman and secretary of the Boston Chapter of the IRE Professional Group on Electron Devices and on various IRE committees.

❖

(Continued on page 76A)

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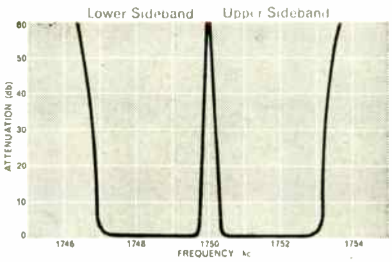
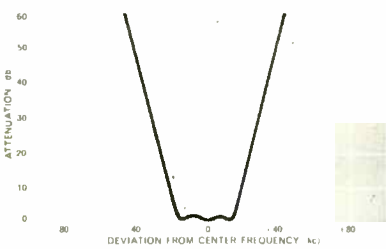


LOWER SINGLE SIDEBAND 1.75 Mc.

Specifications:

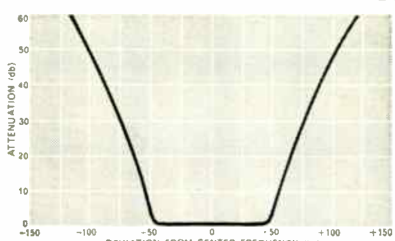
Passband width	2.7 Kc
Carrier rejection	50 db
Maximum ripple	±0.75 db
Impedance (in/out)	50/50 ohms
Maximum insertion loss	3 db
Size	8.5 cu. in.



BANDPASS 10 Mc.

Specifications:

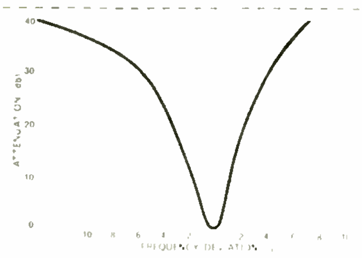
3 db bandwidth	40 Kc
Shape factor (60 db/3 db)	2.2 to 1
Maximum ripple	±0.75 db
Maximum insertion loss	6 db
Impedance (in/out)	1.5K/1.5K
Size	3.6 cu. in.



BANDPASS 30 Mc.

Specifications:

3 db bandwidth	108 Kc
Shape factor (60 db/3 db)	2.1 to 1
Maximum ripple	±1 db
Maximum insertion loss	8 db
Impedance (in/out)	2K/2K
Size	6 cu. in.



BANDPASS 100 Kc.

Specifications:

6 db bandwidth	2 cps max.
Shape factor (30 db/6 db)	5 to 1
Impedance (in/out)	1K/1K
Size	11.75 cu. in.

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HUGHES AIRCRAFT COMPANY

INDUSTRIAL SYSTEMS DIVISION

IRE People

(Continued from page 74A)

Charles J. Pence (A'52–M'57) has been promoted to engineering director of control systems at Remington Rand Univac, St. Paul, Minnesota, chief engineer N. T. Stone announced today.

C. J. PENCE

A graduate of Monroe high school, St. Paul, and the University of Minnesota, Minneapolis, Mr. Pence joined the Univac predecessor company, Engineering Research Associates, as an electrical engineer in 1950. For the past six months, he has been assistant department manager of the Univac advanced navy computer project.

❖

Election of **Herbert W. Pollack** (S'48–A'50–M'55–SM'58) as vice-chairman of the New York Section of the IRE has been announced.

H. W. POLLACK

He is manager of the general engineering department, Polarad Electronics Corporation, Long Island City, N. Y.

A graduate of City College of New York and New York University, he has previously served as chairman of the section's lecture committee and activities review committee, section treasurer, and secretary. The New York Section of the IRE is composed of 5500 engineers and scientists in the Metropolitan New York-New Jersey area.

Mr. Pollack, in addition to his activities with Polarad, is also a member of the electrical engineering graduate school staff at the Polytechnic Institute of Brooklyn.

❖

In furtherance of its recently announced plans for an integrated Electronics Division, Gladden Products Corporation, Glendale, has appointed **Robert R. Reaver** as Special Projects Manager, Electronics Division.

R. R. REAVER

In this newly created post, he is responsible for broadening the Company's electronic base by introducing new products and techniques.

Active in development work in the fields of radar, digital computers and RF

(Continued on page 78A)

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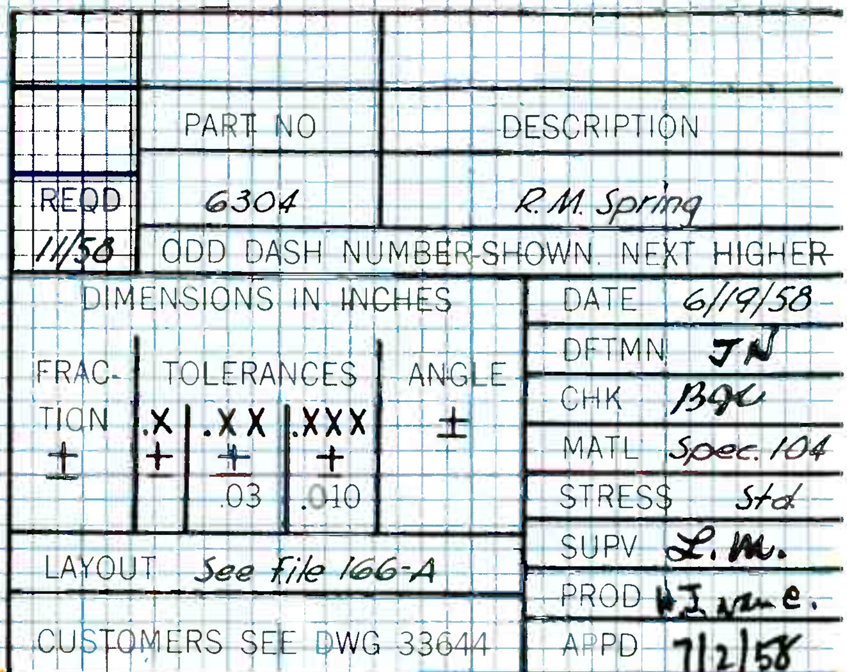

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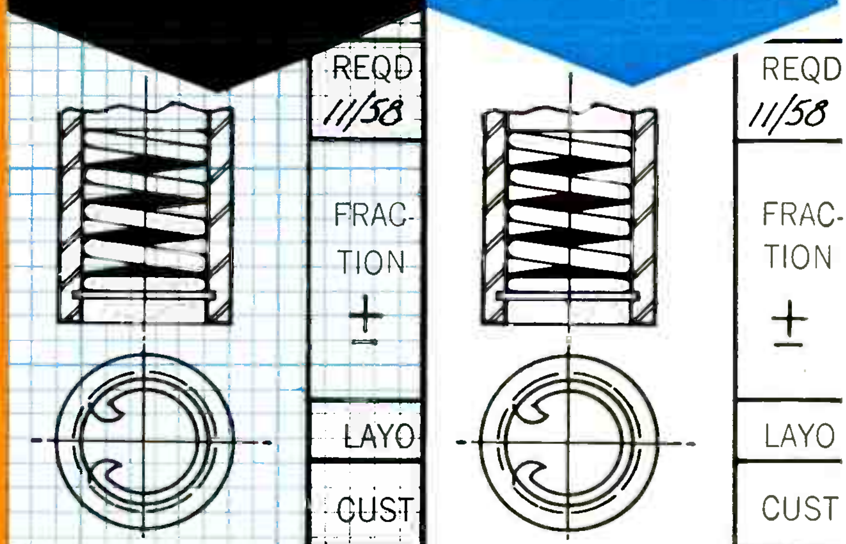

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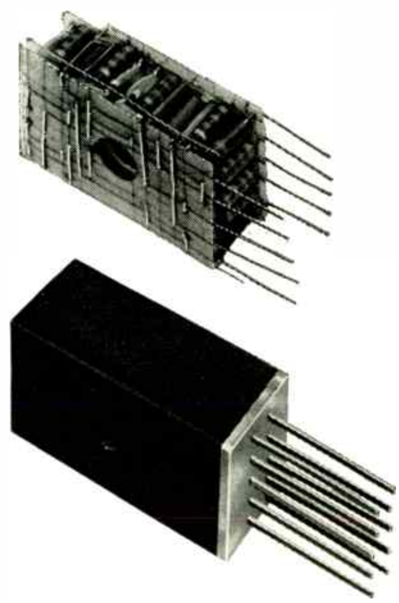


IRE People

(Continued from page 76A)

countermeasures, Reaver was Department Head, Radar-Beacons at Bendix Pacific Division before joining Gladden. He received the B.S.E.E. from Ohio State University in 1950.

❖

Norman J. Regnier (A'47)–M'57) has been named to the position of Program Manager for an advanced semiconductor reliability study being conducted by Motorola Semiconductor Products Division for Autonetics, a Division of North American Aviation, Inc., as part of the MINUTEMAN intercontinental ballistic missile program.

N. J. Regnier

He will have responsibility for over-all program direction and administration, including coordination of the efforts of Motorola's Engineering, Production, Quality Control and Marketing Departments.

Prior to joining Motorola, he was Western Regional Sales Manager and Director of Solar Cell Marketing for Hoffman Semiconductor Division.

A graduate of the University of the West, Los Angeles, Calif., Mr. Regnier received the B.S. degree in Electrical Engineering in 1957.

❖

Appointment of a new Radio Receptor Divisional Vice President has been announced by General Instrument Corporation, of which Radio Receptor Company is a subsidiary.

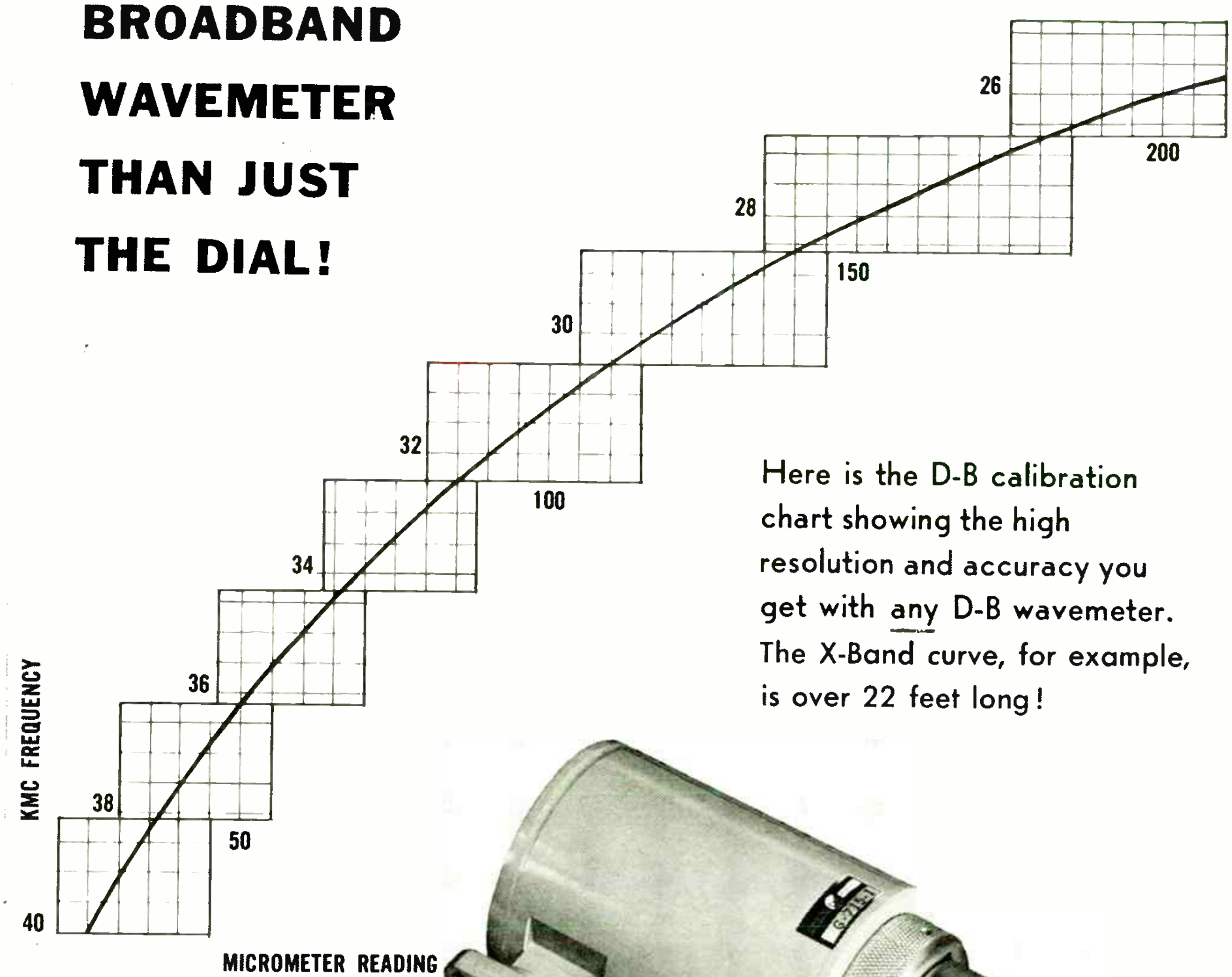
A. L. Rossoff

Arthur L. Rossoff (S'42–A'44–SM'54), until now Technical Director of the Company's Advanced Development Laboratory, has been named Divisional Vice President as well. The Laboratory, which is located in Westbury, Long Island, is part of General Instrument's Radio Receptor subsidiary.

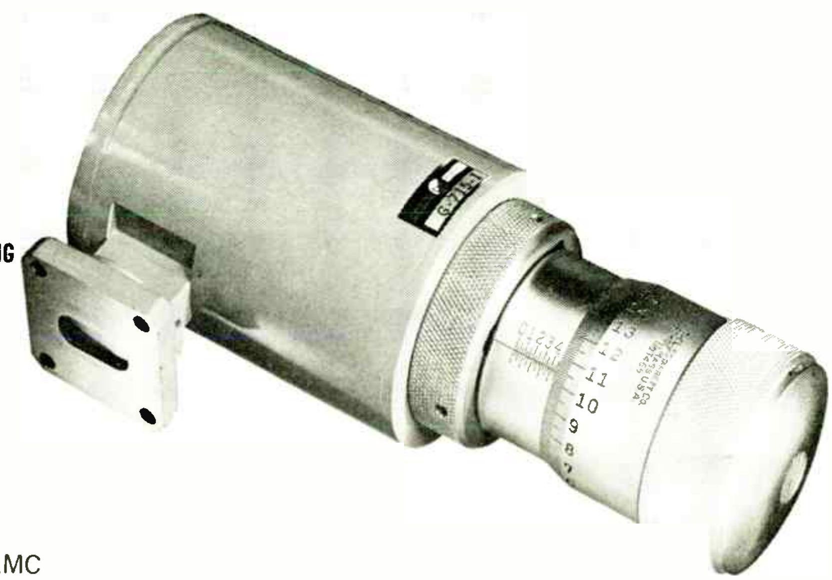
Since joining Radio Receptor's staff in 1950, he has held various posts, principally in product engineering and the Semiconductor Division of the Company. Prior to 1950, he was associated with Dorne and Margolin, Emerson Radio Company, and Universal Electronic Laboratories. He has also taught graduate studies at City College of New York, Polytechnic Institute of Brooklyn, and Columbia University. He is the author of the text book, "Tran-

(Continued on page 80A)

THERE'S MORE TO A D-B BROADBAND WAVEMETER THAN JUST THE DIAL!



Here is the D-B calibration chart showing the high resolution and accuracy you get with any D-B wavemeter. The X-Band curve, for example, is over 22 feet long!



Fail-Safe Design—
loss of gas pressure makes unit immediately inoperative.
Twelve models cover from 2.6 KMC to 140 KMC. Write for complete data in Bulletin D-B 715.

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IRE People

(Continued from page 78A)

sistor Electronics," and is a recognized authority in the field of semiconductor electronics.

Mr. Rossoff holds the B.E.E. degree from the City College of New York and the M.E.E. degree from Polytechnic Institute of Brooklyn.

❖

The appointment of **Dr. Harner Selvidge** (A'31–M'40–SM'43–F'58) formerly Western Corporate Representative of Bendix Aviation, as Vice President and General Manager of Meteorology Research, Inc., was announced by Dr. P. B. MacCready, Jr., President of this Altadena research and instrumentation company. The appointment is the first step in a plan for substantial expansion to meet increasing demands for meteorological research and instrumentation.

H. Selvidge

He has held a number of executive positions with Bendix during the past 15 years, specializing in missiles, instrumentation and new product development programs. Before joining Bendix, he was a staff member of the Applied Physics Laboratory of Johns Hopkins University, Baltimore, Md. Prior to the war, he taught at Harvard, Massachusetts Institute of Technology and Kansas State University, Lawrence. He holds advanced degrees from the Massachusetts Institute of Technology and Harvard University.

Dr. Selvidge is a member of the American Physical Society, the American Rocket Society and the Institute of the Aeronautical Sciences. For his work on proximity fuzes, he received the Naval Ordnance Development award in 1946. He is also President of the Soaring Society of America.

❖

Crosby-Teletronics Corporation announced the appointment of **John C. Simmons** (M'44) as assistant to the executive vice president. He will continue as contracts manager, a post he had held since 1957.

J. C. Simmons

Crosby-Teletronics Vice President, R.S. Marston, said that Mr. Simmons' promotion was designed to facilitate the coordination of sales, production and engineering departments of Crosby-Teletronics, which, along with its subsidiaries, is primarily a manufacturing, research and development company.

Mr. Simmons was contracts manager for Teletronics Laboratory, Inc., prior to its merger with Crosby Laboratories in 1959. From 1952 to 1957 he was in the engineering department of Teletronics and served as project engineer on Signal Corps and Navy Department development contracts.

Prior to joining Teletronics Laboratories, he spent 10 years with the Sperry Gyroscope Company. As project engineer with Sperry, he was responsible for the development and evaluation of air-to-air and air-to-ground microwave communication for the Bureau of Aeronautics, Navy Department. He also was responsible for the development of the Telemetering Signal Simulator for missile systems and worked on many other military communications developments.

❖

Robert L. Sell (S'47–A'49–M'55–SM'59), engineering executive for Telex, Inc. has been named assistant vice president of the Twin Cities electronics company, Arnold J. Ryden, president, has announced.

R. L. Sell

He will continue as director of engineering, Ryden said. In addition to his engineering duties, he is being given new responsibilities as Director of Sales and Marketing for the company's Components Group which includes Special Products, Communications Accessories Division, and Magnetics.

Since he joined Telex a year ago, he has directed development of a number of advances in hearing aid design, including its recently-introduced Telex Radiant, the first hearing aid to eliminate wire and mechanical connections by use of a radio transmission principle. Prior to joining Telex, he was chief engineer at Audio Development Company, Minneapolis, with ten years service there.

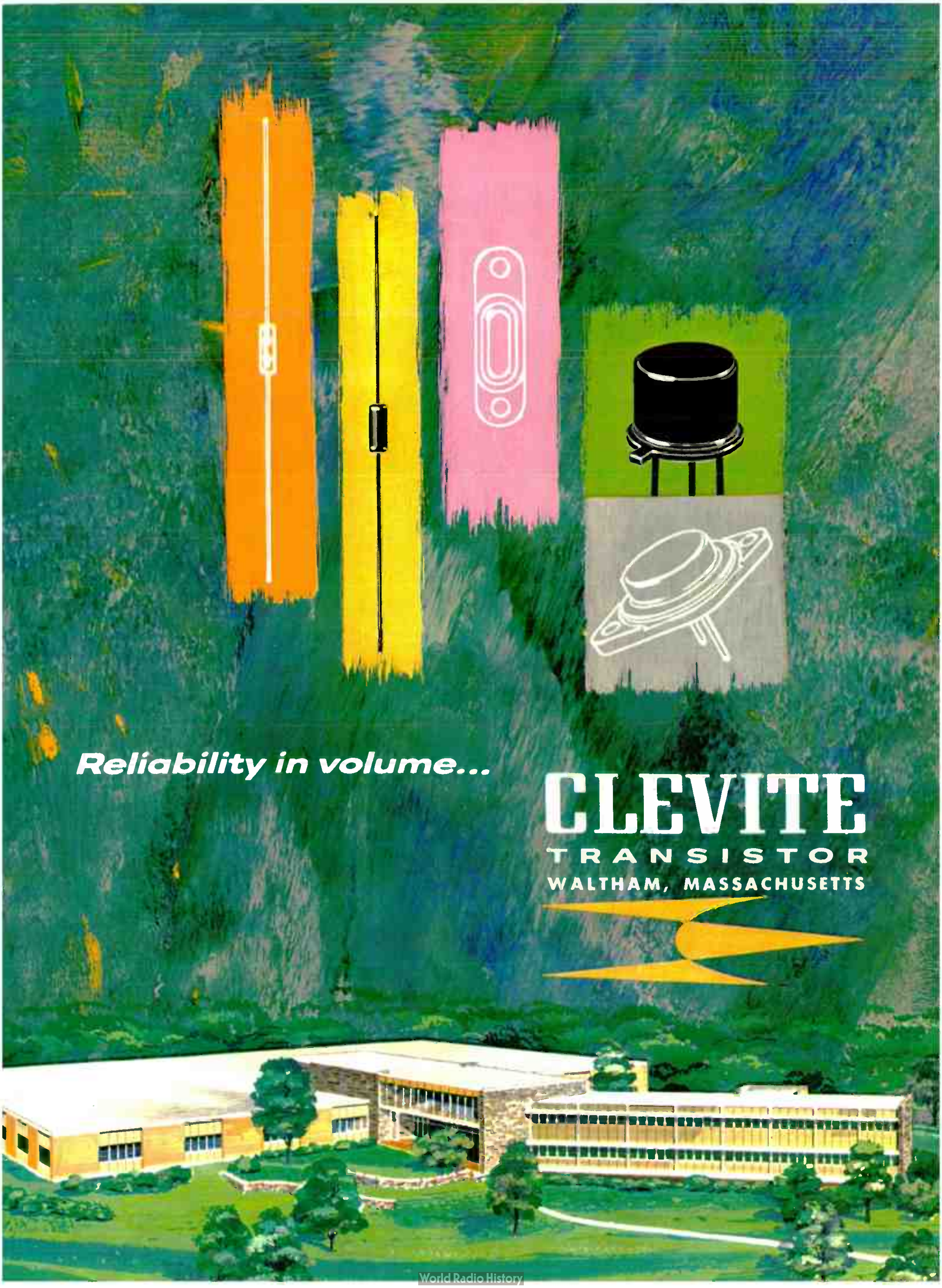
Mr. Sell has been active in the Institute of Radio Engineers, serving as chairman of the Twin Cities section of the Professional Group on Audio. He is also a member of the American Institute of Electrical Engineers and holds a committee chairmanship in the Electronics Industries Association.

❖

Emerson & Cuming, Inc., Canton, Massachusetts, announces the appointment of **Howard A. Smith** (A'54) as National Sales Manager. Mr. Smith joined the firm in 1958 as New England District Sales Manager. In his new position, he is responsible for all sales efforts for the company including its international representative coverage, regional sales offices, sales promo-

Howard A. Smith

(Continued on page 83A)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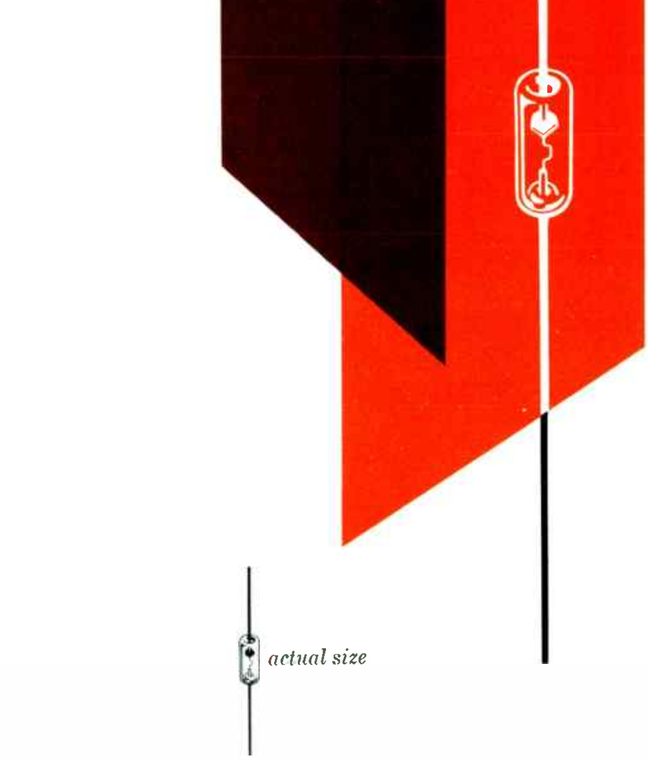

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(Continued from page 80A)

tion and advertising. Mr. Smith's background is extensive in the design and aplication engineering of dielectric material.

For two years previous to joining Emerson & Cuming, he was partner/owner of the electronics manufacturers' representative organization of Smith & Purdy Associates covering the New England Territory and specializing in microwave test equipment, components and materials.

Prior to that, his experience was primarily in the microwave instrumentation field, including antenna design and special test equipment design. He was associated with Gabriel Electronics for four years along this line.

Mr. Smith is at present the Editor of "The Reflector" magazine published by the Boston Section of the IRE.

❖

Robert C. Sprague (SM'53), chairman of the board of the Sprague Electric Company, has been elected a Fellow of the American Academy of Arts and Sciences.

The 180-year-old academy lists among its members about 1500 distinguished intellectual leaders in various fields. Mr. Sprague was one of the 115 new fellows elected this year.

Mr. Sprague, who founded Sprague Electric 34 years ago, is a graduate of the U. S. Naval Academy and holds the master's degree from Massachusetts Institute of Technology, Cambridge. He is chairman of the board of the Federal Reserve Bank of Boston.

His many services to the U. S. government include a tour as chairman of the so-called Gaither Committee to assess American defenses. The committee's 1957 report remains a closely guarded secret.

❖

H. Myrl Stearns (S'39–A'40–SM'48), president of Varian Associates, received an honorary Doctor of Science degree at special ceremonies at the 65th commencement exercises of the University of Idaho, Moscow, Idaho, June 5.

H. M. Stearns

He joined Russell and Sigurd Varian, the inventors of the klystron tube, and other scientists in founding Varian Associates in 1948. He served as executive vice president and general manager of the electronics firm from its founding until June, 1957 when he was named president.

He received his bachelor of science degree from the University of Idaho in 1937. After graduation he went to Stanford University and served as a teaching and research assistant while studying for his ad-

(Continued on page 84A)

IRE People

(Continued from page 83A)

vanced degree in electrical engineering, which he received in 1939.

In 1941 he joined Sperry Gyroscope and later headed that company's Doppler radar program. In 1946 he received a commendation from the Secretary of the Navy for work on radar in connection with submarine warfare and leadership in the war effort at Sperry.

Mr. Stearns holds patents in the fields of automatic frequency control and automatic ranging and is the author of technical research papers in the fields of radar, microwave tubes and engineering management. He is a past president of the Western Electronics Manufacturers Association, a Fellow of the American Association for the Advancement of Science and a member of Sigma Xi and Sigma Tau, scientific and engineering honorary societies.

❖

Telecomputing Corporation, Los Angeles, Calif., has announced the appointment of **Steve E. Strem** (M'57) to the newly created position of manager of its technical services department.

He will be responsible for preparation, graphic arts and production of proposals, publications, displays, and related services for the entire corporation.

He was formerly supervisor, project services section and a member of the technical staff of Hughes Aircraft Co., Culver City, Calif. Previously, he was technical publications manager of Associated Missile Products Corp., and before that, he was employed by Hycon Mfg. Co., Pasadena, Calif.

Educated in Paris, France, Mr. Strem is also a graduate of American Radio Institute. He is a past vice president and director, Technical Publishing Society national board, and presently a member of the Los Angeles Chapter.

❖

The Thurlow Navigation Award, the highest honor conferred for advancing the science of navigation, has been presented to **William J. Tull** (M'46–SM'55), Vice President, GPL Division-General Precision, Inc. for his work as the pioneer of Doppler navigation.

W. J. Tull

The award was presented by the Institute of Navigation during its annual meeting at the U. S. Air Force Academy.

The citation described Mr. Tull as "the father of Doppler navigation." It stated that he saw in 1945 the possibilities of applying Doppler principles to the problems of air navigation and that "his efforts resulted in development of the world's first practical automatic self-contained airborne navigation system." The citation further lauded him for his continuing role in refining Doppler systems for use in lighter aircraft.

He is one of the original M.I.T. scientists who founded GPL, and was largely responsible for the invention, design, and development of Doppler navigation systems.

Prior to the founding of GPL, he was a Staff Member of the M.I.T. Radiation Laboratory, working on missile guidance systems and the AN/APQ-24. He holds eleven patents on Doppler systems and computers, ferrite devices, missile guidance systems, and portions of the AN/APQ-24 Navigation and Bombing System.

He also made significant contributions to compass technology, having originated techniques for eliminating acceleration effects in gyro-magnetic compasses, and techniques for compass swinging. Microwave transmitters and antennas, magnetron phasing, Doppler theory, high duty cycle radars and magnetrons, and Doppler systems flight tests are other areas of his research efforts.

He has furthered the growth of GPL in administrative, as well as scientific, functions: as coordinator of military projects, and as supervisor of sales planning and future research, development and engineering.

He graduated from the University of Michigan, Ann Arbor, in 1942 with the B.S. degree in Electrical Engineering; he also did graduate study there.

The Thurlow Award dates back to 1945 when it was founded by Sherman Mills Fairchild to stimulate development of the science of navigation. The Award is given in memory of Colonel Thomas L. Thurlow, who, in the early thirties, was instrumental in establishing a navigation school for pilots, and, with Harold Gatty, laid the groundwork for future training in navigation. Over the years, it has been awarded to persons who have written new chapters in the history of navigation.

❖

David R. Weindorf (S'50–A'51–M'56) has been appointed Head, Production Engineering Section of the Rheem Semiconductor Corporation, it has been announced by Dr. E. M. Baldwin, Vice President and General Manager of the Mountain View, California silicon mesa transistor, diode, and special assemblies firm.

Prior to joining Rheem, he was employed as a Project Engineer for Sylvania Electric Products and Head of Preproduction Engineering for Hughes Aircraft Company and Fairchild Semiconductor Corporation.

Mr. Weindorf hold the Bachelor of Science degree in Electrical Engineering from Pennsylvania State University, University Park, and has patented a ceramic-metal fuse tube.

❖

David P. White (S'59–M'60) of Watertown, Massachusetts, a senior at Brown University, Providence, R. I., has been

(Continued on page 86A)

LOWEST NOISE* POWER KLYSTRONS FOR CW RADAR AND ILLUMINATORS

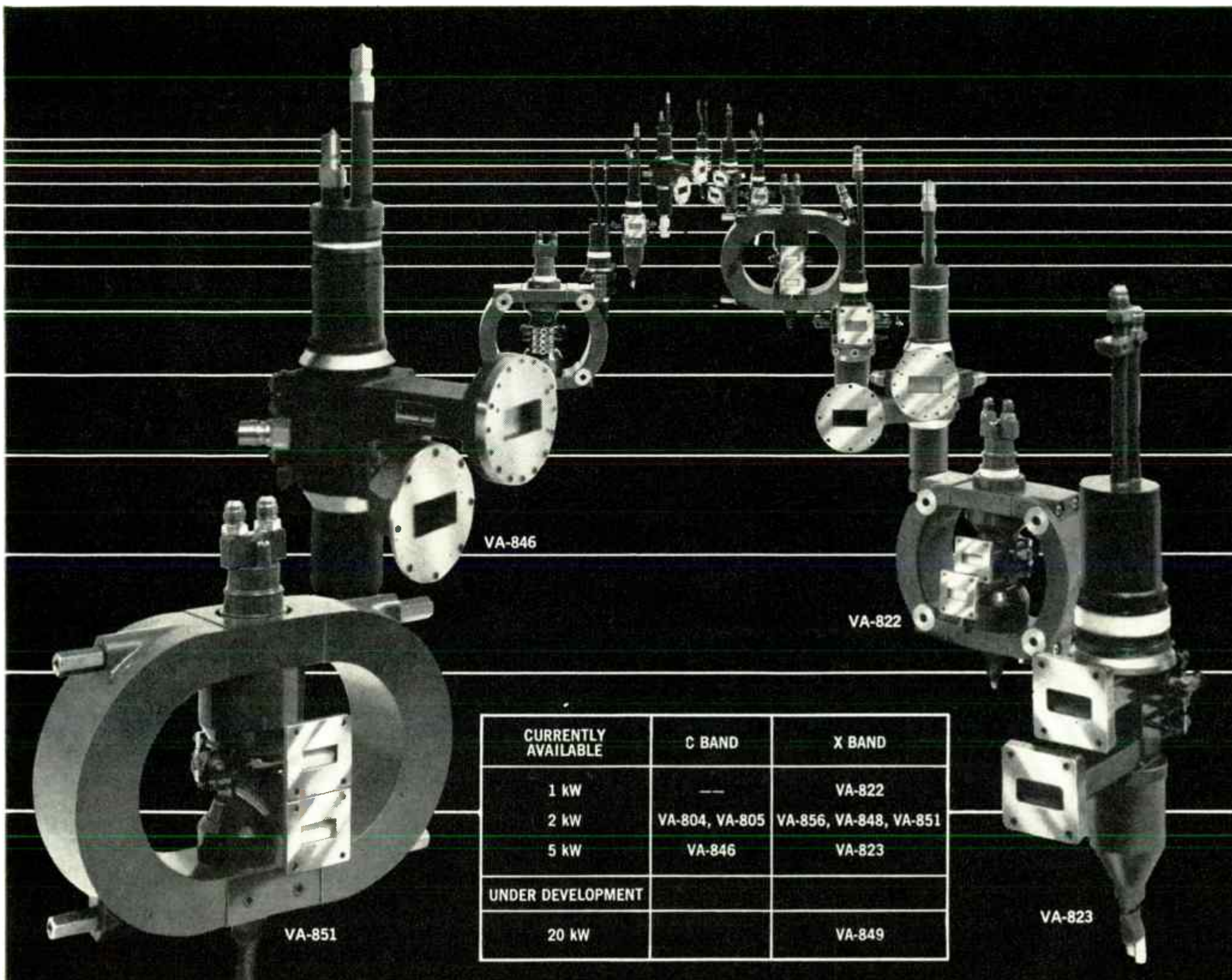
1 to 5 kilowatts CW power ● Noise 100 db below carrier*
50 db gain (most types) ● C and X band frequencies
Efficiency up to 45% ● Focusing by electromagnet or permanent magnet

*AM and FM noise is more than 100 db below carrier in any 1 kc channel more than 1 kc removed from the carrier.

Varian is delivering many high-power low-noise CW klystrons for CW radar and illuminators; eight types are unclassified and are listed below. A 20 kilowatt X band type, soon to be added, will be another extension of the state-of-the-art by Varian.

Because of their extremely low noise close to the carrier, these types will enable you to design radar and illuminator systems of great sensitivity and precision. They also afford new design opportunities in radio astronomy and communications systems.

May we send full specifications or discuss development of special types for your requirements?



CURRENTLY AVAILABLE	C BAND	X BAND
1 kW	—	VA-822
2 kW	VA-804, VA-805	VA-856, VA-848, VA-851
5 kW	VA-846	VA-823
UNDER DEVELOPMENT		
20 kW		VA-849

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IRE People

(*Continued from page 84A*)

awarded the coveted Hannibal C. Ford Fellowship for study at the Graduate School of Engineering of Cornell University, it has been announced.


D. P. WHITE

The fellowship, awarded by the Ford Instrument Company division of Sperry Rand Corporation, commemorates the memory of Hannibal C. Ford, Cornell '03, inventor, scientist, designer, and electromechanical genius, one of the nation's pioneers in the development of ordnance and navigational controls and computers, and founder of the company which bears his name.

He attended Watertown High School and entered Brown University in 1956. He has had a distinguished record at the university and held a four year scholarship. At graduate school he plans to concentrate in the field of electromagnetic wave propagation.

At Brown, Mr. White is a member of Tau Beta Pi, Sigma Xi, the Brown Engineering Society, and the Brown Engineering Student Council.

❖

Dr. Alfred K. Wright (A'37–SM'45), vice president of operations, Tung-Sol Electric Inc., has been elected to the company's board of directors, Louis Rieben, chairman, has announced.


A. K. WRIGHT

He joined the 56-year-old electronics and automotive components manufacturing firm in 1937, and has held a succession of managerial posts with the company.

In 1938, as a sales engineer, he helped develop the firm's then new branch of engineering activity which aided radio and other electronics set manufacturers in solving advanced technical problems. This activity was an important factor in the growth of Tung-Sol as a major supplier of original equipment electron tubes.

In 1951, he was elected vice president in charge of engineering. In 1958, he was given the broadened responsibilities of his present position.

He is a member of the Joint Electron Tube Engineering Council, the American Society of Quality Control, and the Society of Automotive Engineers. He is a graduate of Northeastern University, Boston, and Harvard University, Cambridge. He received the Doctor of Science degree from Harvard.

On April 21, 1960 Dr. Wright received a "Citation for Distinguished Attainment" from Northeastern during the exercises commemorating the 50th anniversary of the College of Engineering and its cooperative engineering program.

❖

Directors of Hazeltine Corporation have announced the election of **Dr. Victor J. Young** (SM'52) as vice president of Hazeltine Electronics Division. He is in charge of the Electrical Engineering Department of the 36-year-old electronics firm. He had been an assistant vice president of the electronics division since 1956.


V. J. YOUNG

He was graduated cum laude from Albion College in 1935, and received the Master's degree in Physics from the University of Iowa, Iowa City, while on a teaching fellowship there. After earning the Ph.D. degree in Physics from the University of Iowa in 1940, he was a physics instructor at New York University where he also was active in nuclear and radar research programs. Prior to joining Hazeltine in 1949, he was in charge of radar, undersea detection and recorder projects for Melpar, Inc. and the Sperry Gyroscope Company.

The new engineering vice president is the author of "Understanding Microwaves," published by John Rider in 1945, and many scientific papers on physics and electronics problems. He is a member of the American Physics Society, New York Academy of Science, American Ordnance Association, Alpha Tau Omega, Sigma Xi, Delta Sigma Rho and Kappa Mu Epsilon.

❖

The appointment of **Ray A. Zuck** (A'55) as vice-president of Atronic Products, Inc., Bala-Cynwyd, Pa., electronic equipment manufacturer, has been announced by **George J. Laurent** (A'47–SM'52) company president. Mr. Zuck will direct activities in design and production of Material Handling Systems.

R. A. Zuck

Since joining Atronics, he has been active in work on automatic test equipment for production testing of transistor circuits; in transistorized recognition circuits for detection of specialized signals; and in electromechanical control equipment for both military and industrial applications.

During the last two years, he has been responsible for the development of Atronics' line of industrial products. This includes a series of code-reading devices for control of packages on conveyor lines, and automatic material handling systems for warehouses.

Born in Ladysmith, Wis., he received the Bachelor of Science degree in electrical engineering from the University of Wisconsin, Madison. While at the university, he worked at the Wisconsin Alumni Research Institute on computers for electronic weather instruments, and, in the Navy, he served as an Aviation Electronics Technician.

In his previous experience at Philco Corporation and Electronic Tube Corporation, he led engineering groups concerned with airborne radar circuit techniques, and did development work on several industrial products, including a hearing aid and a personal paging system. He designed and developed new instrumentation, including a stabilized DC amplifier and a neuro-electronic stimulator and recorder.

During this period, Mr. Zuck served on the faculty at Drexel Institute of Technology, Philadelphia, Pa., for four years, teaching courses in transmission lines, communication circuit analysis, and transistor circuits. He has a number of patents pending as a result of working in these fields, and he is the author of several technical articles.

❖

Use your

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It's valuable!

Multi-Phase

Voltage Calibrators

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Industrial Engineering Notes*

Governmental and Legislative

Statistics for standard metro areas published—The Census Bureau's preliminary General Statistics for Selected Standard Metropolitan Statistical Areas by Major Industry Groups, drawn from the 1958 census of manufactures, has just been published. Order MC(P)-4 (Preliminary) from the Bureau of the Census, Washington 25, D. C. Price 25 cents.

Engineering

Expanded "phosphor book" published by JEDEC—A greatly expanded edition of "Optical Characteristics of Cathode Ray Tube Screens," a compilation of available phosphor data related to cathode ray tube applications, has been published by the JEDEC Electron Tube Council. The 106-page "phosphor book" contains a wealth of information on phosphors, including spectral energy distribution data and curves and persistence curves for all registered phosphors up to P29. The revised publication also contains sections on the measurement of color, persistence and screen brightness, and describes standards for use in making the measurements. Also included is information on phosphor screen defects and a bibliography of phosphor literature. The book was prepared by the JEDEC JT-6.3 Subcommittee on Phosphors and Optical Characteristics of Cathode Ray Tubes. It is identified as JEDEC Publication No. 16 and may be ordered from the EIA Engineering Office, Room 2260, 11 West 42nd Street, New York 36, N. Y. Price is $5.00 a copy.

Industry Marketing Data

Expanded Japanese market seen for automated machines—An increasingly larger market in Japan for U.S.-made electronic and electromechanical business and industrial machines was forecast last week by the Department of Commerce week by the Department of Commerce's Bureau of Foreign Commerce. Reporting the observations of the sixth and latest U. S. Trade Mission to Japan, BFC said automated machines have been effectively used in Japan and that a need will develop for many more. "The larger industrial and public utility organizations, and the more progressive business concerns, have accepted and applied electronic and electromechanical equipment very effectively," according to the trade mission's observations. The mission found that Japanese business lags far behind the United States in using modern processes and equipment. This is apparently the fault of management attitudes, which appear to hold that only very large firms can use automated equipment. "There seems to be little realization that in the United States many firms, large, medium size, and small, utilize and improve management and administrative functions through use of modern data-processing equipment, accounting machines, and other modern office equipment." The mission observed that a large part of office operations are being done by hand. "Realization is growing, however, among progressive business leaders that introduction of modern office machines can add to management efficiency," the mission found. "Prospects for U.S. manufacturers of such equipment to participate in this business to a greater degree appear very bright," the mission concluded. . . . **March transistor sales up 2½ million units —The number of transistors sold at the factory in March took an impressive two and one half million jump over total units sold during previous month, and revenue from factory sales rose nearly $4 million.**

The month-end statistics compiled by the Marketing Data Department:

	Factory Sales Units
March	12,021,506
February	9,527,662
January	9,606,603
Year-to-date '60	31,155,798
Year-to-date '59	16,898,980

Transistor sales drop during April, but remain healthy—Total factory sales of transistors dropped to 9.891 million units in April after hitting a record high of 12 million in March, according to figures just published by the EIA Marketing Data Department.

April's total sales, however, were more than one million units higher than the best monthly total for 1959—8.652 units sold in September. Year-to-date totals for 1960 remained significantly ahead of those for the same period last year.

	Factory Sales Units
April	9,891,236
March	12,021,506
February	9,527,662
January	9,606,630
Year-to-date '60	41,047,034
Year-to-date '59	22,805,716

* The data on which these NOTES are based were selected by permission from *Weekly Reports*, issues of June 6, 13 and 20, published by the Electronic Industries Association whose helpfulness is gratefully acknowledged.

Professional Group Meetings

Aeronautical and Navigational Electronics

Oklahoma City—April 26

"FAA Doppler VOR," C. Bibberstine, FAA, Oklahoma City; Movie: "Eyes of Flight" (in color), furnished by Ryan Electronics Co., San Diego, Calif.

Philadelphia—March 29

"An Automatic Air Traffic Control System," A. W. Muoio and C. H. Taylor, RCA, Camden, N. J.

Antennas and Propagation

Boston—April 18

"Phase Modulated Traveling Wave Antennas," A. S. Thomas, A. S. Thomas Inc., Westwood, Mass.

"The Design of a Lens Corrector for a Spherical Reflector," H. J. Rowland and J. Seavey, D. S. Kennedy & Co., Cohasset, Mass.

Los Angeles—February 24
Orange Belt

"Conical Helix Feed For Parabolic Antenna," R. F. H. Yang, Andrew Corp., Chicago, Ill.

Los Angeles—May 12

"A Pillbox Feed System for Monopulse Operation," P. A. Jensen, Hughes Aircraft Co., Fullerton, Calif.

San Francisco—May 11

"A Slot Antenna with Variable Coupling and its Application to a Linear Array," R. Tang, Hughes Aircraft Co., Fullerton, Calif.

Antennas and Propagation
Communications Systems

Washington, D. C.—January 25

"Communications Systems for Project Mercury Space Capsule," W. Benner, McDonnell Aircraft Corp., Robertson, Mo.

Orange Belt—May 25

"An Axial Mode Bifilar Helical Antenna," A. G. Holtum, Jr., Andrew California Corp., Claremont, Calif.

Antennas and Propagation
Microwave Theory and Techniques

Columbus—January 11, 1960

"Present Trends in Antenna Research," T. E. Tice, Ohio State Univ., Columbus.

Columbus—February 2

"Present Trends in Radar Scattering Theory," K. M. Siegel, Univ. of Michigan, Ann Arbor.

Columbus—May 31

"Present Developments in Propagation Phenomena," Prof. E. C. Jordon, Univ. of Illinois, Urbana.

Audio

Cleveland—May 19

"The Wood Panel that Talks," A. B. Cohen, Advanced Acoustics, Inc., Nutley, N. J.

Bio-Medical Electronics

Philadelphia—October 14

"Modern Analytical Methods for Evaluating Cardio-Vascular Function," L. H. Peterson, Univ. of Pa. Medical School.

Philadelphia—January 13

"Dosimetry in X-Ray Therapy," Dr. G. Henny, Temple Univ. Medical School, Philadelphia, Pa.

Philadelphia—March 9

"A Portable Artificial Cardiac Pacemaker," 1) S. Furman, Montefiore Hospital, New York, N. Y.; 2) D. G. Kilpatrick, Atronic Products, Philadelphia, Pa.; 3) O. Muller, Philadelphia General Hospital; 4) L. D. Sher, Moore School of EE, Univ. of Pennsylvania, Philadelphia.

Philadelphia—May 19

"Medical Tests with the Panex X-Ray Amplifier," 1) E. W. Godfrey, Princeton Hospital, Princeton, N. J.; 2) C. P. Hadley, RCA, Lancaster, Pa.

Bio-Medical Electronics
Instrumentation
Space Electronics and Telemetry

Washington—March 14

"Biological Instrumentation for Space Vehicles," R. M. Adams, Aerospace Medical Center, San Antonio, Tex.

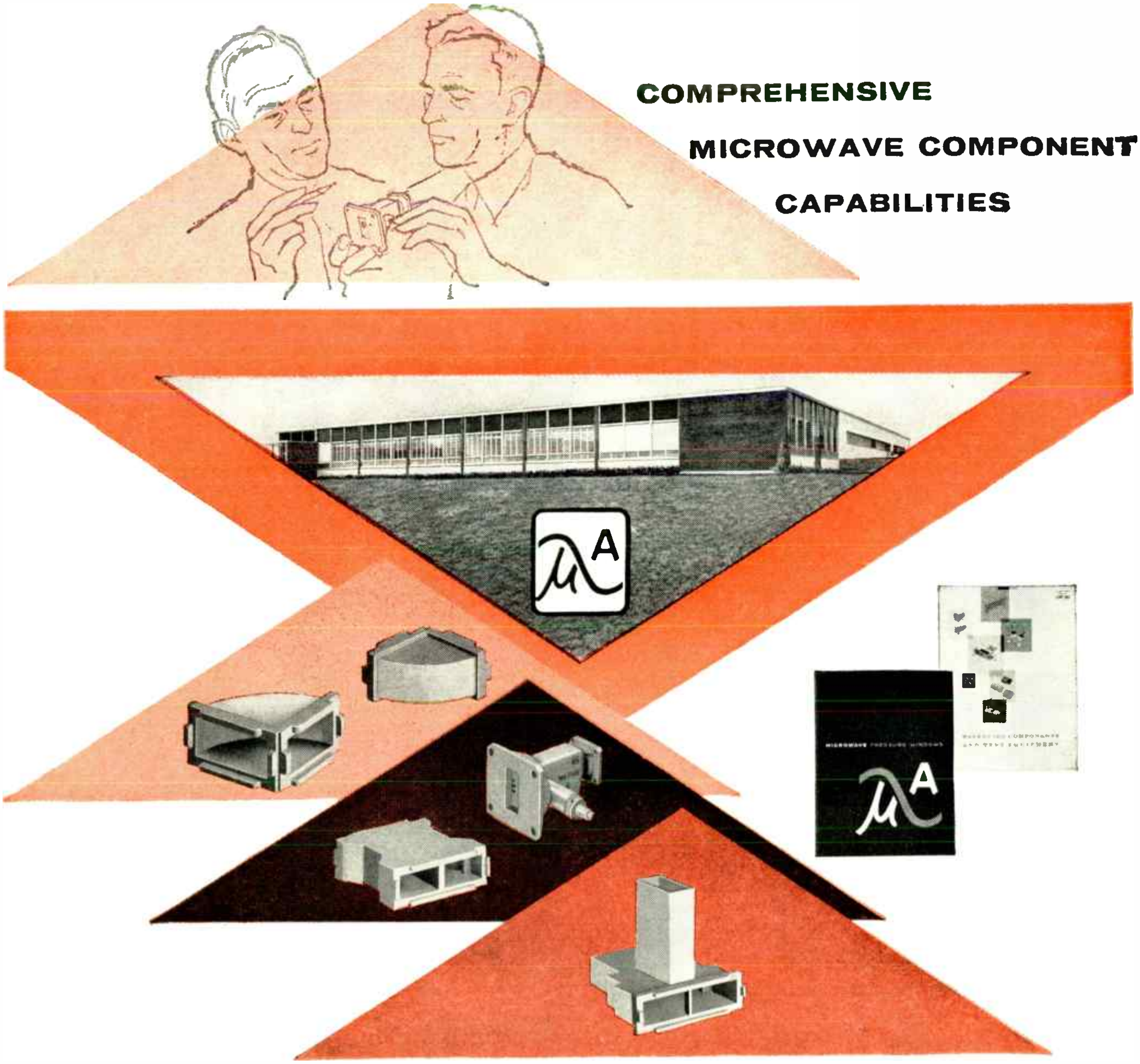
Broadcasting

Houston—April 12

"Directional Antennas and Phasing Networks," R. S. Bush, Gates Radio Co., Quincy, Ill.

(Continued on page 92A)

COMPREHENSIVE MICROWAVE COMPONENT CAPABILITIES



This modern three-quarter-acre plant has significantly expanded the services and production capabilities of Microwave Associates' experienced Waveguide Components Division. This new research and production facility is one of the most completely equipped on the east coast. A large 3′ x 2′ capacity dip-brazing unit as well as complete plating and other shop facilities are now handling both large volume and custom-engineered orders. Components are precision-machined and produced in beryllium-copper, cast and fabricated aluminum, and cast magnesium.

Over 500 microwave components for applications from 1.12 to 90.0 kMc/s are standard items. Our Sales Engineers will gladly discuss current work in sophisticated components and RF packaging with you.

A FEW OF THE MANY COMPONENTS MANUFACTURED HERE

New High-Power Varactor Harmonic Generators — excellent suppression of unwanted harmonics and record power levels are available from these solid-state harmonic generators.

New Cast Bends — Zero bend radius — 90° E and H plane bends in S through Ka bands... Each bend is compensated to a VSWR of 1.05 over its entire waveguide band.

Sidewall Hybrid Couplers (3db) and H-Plane Folded Hybrid Tees — Cast in aluminum and beryllium-copper are available in S through Ka-band models.

Two New Catalogs — Waveguide Components Short-form Catalog (CSF-60) gives data on over 500 items of waveguide components and test equipment.

Pressure Window Catalog (12 pages) contains electrical and mechanical data on a complete line of glass-kovar, mica, and special pressure windows plus valuable installation and testing tips.

MICROWAVE ASSOCIATES, INC. Burlington, Mass. • Western Union FAX • TWX Burlington, Mass 942 • BRowning 2 3000
Export Sales, Microwave International Corporation, 36 West 44th St., New York, New York

Professional Group Meetings

(Continued from page 90A)

Philadelphia—April 14

"TC Automation," R. D. Houck, RCA, Camden, N. J.

Communications Systems

Oklahoma City—March 24

"Underwater Television used in Skate Submarine in Recent Expedition Under the Polar Ice Cap," B. A. Bang, Bendix-Friez Instrument Div. (This meeting was open to the public. A film of the installation and expedition was shown.)

Oklahoma City—May 24

"Tour of Telephone Facilities with Question and Answer Session," G. W. Holt, S. W. Bell Telephone Co., Oklahoma City.

Rome-Utica—April 13

"The RCA Microminiaturization Program," D. Mackey, Radio Corp. of America.

Rome-Utica—May 5

"Managers Born or Developed," R. F. Risley, Cornell University, Ithaca, N. Y.

"Marketing & Marketing Planning in the Defense Market," J. H. Richardson, Hughes Aircraft Corp., Los Angeles, Calif.

"Planning—A Tool of Engineering Management," R. J. Rattell, Westinghouse Air Arm Div.

San Francisco—April 26

"Non-Great Circle Communications," R. Wolfram, Stanford Research Institute, Menlo Park, Calif.

Communications Systems Vehicular Communications

Omaha-Lincoln—January 27

"Automobile Transistor Voltage Regulators," R. Beard, Delco Radio.

"Diode AC Generators for Cars and Trucks," P. Benedict, Delco-Remy.

Automotive Electronics Films, commentary by R. Enders.

Tour of General Motors Training Center.

Syracuse—January 12

"Communications—Electronics in Aerospace Strategy," J. B. Bestic, United States Air Force.

Component Parts

Baltimore—May 17

"Micro-Circuitry: Recent Developments and Reliability Aspects," Dr. J. A. Bohrer, International Resistance Co., Philadelphia, Pa.

Los Angeles—February 8

"Thermoelectrics in Electronics," W. E. Bulman, Ohio Semiconductors, Inc., Columbus, Ohio.

Los Angeles—June 1

"The Micro-Component Approach," E. Madden, Pacific Semiconductors, Inc.

"Circuit Design for Microcircuitry," H. J. Weber, Servomechanisms, Inc.

"High Density Component Interconnections," J. Richardson, Hughes Aircraft Co.

Electron Devices

Los Angeles—February 17

"Joint Symposium on Tunnel Diodes," R. M. Hall & others, General Electric Co.

Los Angeles—April 27

Debate: "Vacuum Tubes vs. Solid State Devices for Microwave Power Generators," Dr. L. M. Field, Hughes Aircraft Co., Culver City, and Dr. H. Kroemer, Varian Associates, Palo Alto.

Los Angeles—June 1

"Microminiaturization in Electronics," E. Madden, Pacific Semiconductors, Inc.; H. J. Weber, Servomechanisms Inc., Hawthorne, Calif.; J. E. Richardson, Hughes Aircraft Co., Los Angeles, Calif.

(Continued on page 94A)

ANTENNA TOWERS & STRUCTURES

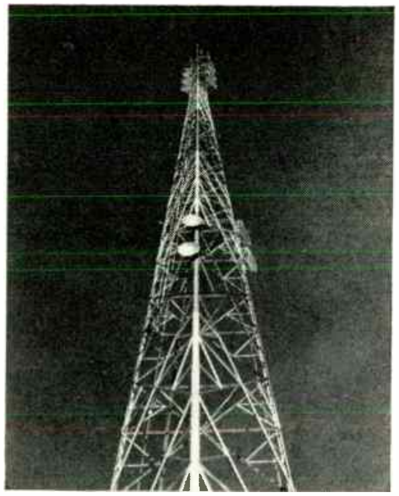
Dresser-Ideco Designs, Fabricates, Erects A Wide Variety of Large Land-Based Antenna Structures

Whatever the antenna tower or structural problem, Dresser-Ideco can handle the job. A large technical service staff, more than a quarter-million square feet of well-equipped production area, extensive research and development facilities, and many years experience designing and erecting steel and aluminum structures qualify Dresser-Ideco as a major contractor for any type of antenna tower . . . from 100 to 2000 feet. Below are some of the many types of Dresser-Ideco towers serving the nation's communications and military electronics facilities.

A.

B.

C.



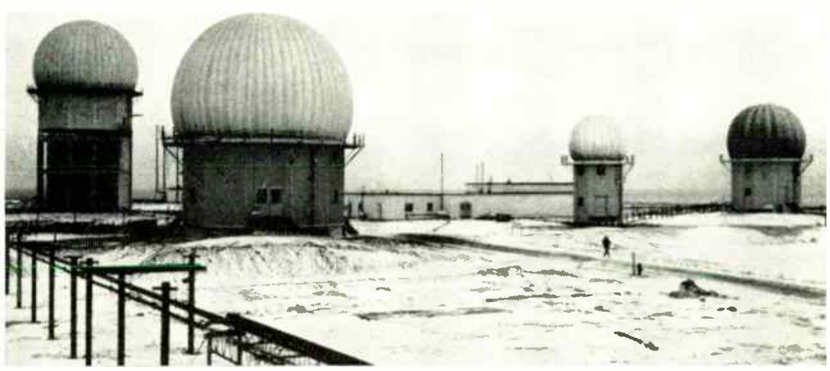
A. Multiple antenna tower. This big 729′ tower in Baltimore supports three television antennas on a 105′ wide platform at the top.

B. Antenna test range tower for height finding radar.

C. Microwave antenna tower typical of those used in systems built for the Ohio Turnpike and the Illinois Toll Road.

D. Surveillance radar towers on the DEW line. Designed for hurricane force winds and heavy ice and snow loads. These are some of more than 1,000 radar towers built by Dresser-Ideco for the nation's early warning systems.

D.



Dresser-Ideco offers a complete tower and structural service—design, application, field installation. Call or write for consultation on your structural problems at no obligation. Send for booklet "Facilities For Defense Production". Dresser-Ideco Company, A Division of Dresser Industries, Inc., 875 Michigan Avenue, Columbus 15, Ohio. Branch: 8909 So. Vermont Avenue, Los Angeles 44, Calif.

DRESSER-IDECO COMPANY

DRESSER IDECO®

DRESSER INDUSTRIES INC.

ELECTRONIC • INDUSTRIAL
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Professional Group Meetings

(Continued from page 92A)

San Francisco—April 26

"Ultra Microwave Generators and Amplifiers with Maserlike Systems," J. R. Singer, Univ. of California.

San Francisco—May 25

"Localized Imperfection in *p-n* Junctions," Dr. A. Goestzberger, Shockley Transistor Corp., Mt. View, Calif.

Washington, D. C.—May 23

"Survey of Low-Noise Microwave Amplifiers," Dr. G. Wade, Raytheon Mfg. Co., Burlington, Mass.

Electronic Computers

Houston—April 12

"Transistorized Data Process Recorder," H. B. Patterson, Southwestern Industrial Electronics, Houston, Tex.

Omaha-Lincoln—February 17

"Applications of an Analog Computer in Support of Development Engineering," L. R. Abkemeier, McDonnell Aircraft, Robertson, Mo.

San Francisco—March 22

1) "An Evaporated Random Access Memory";
2) "A Thin Film Magnetic Shift Register," K. D. Broadbent, Hughes Res. Labs., Culver City, Calif.

San Francisco—April 26

"Parametric Phase-Locked Oscillators and Esaki Diodes as Switching Elements," Dr. A. W. Lo, IBM Corp., Poughkeepsie, N. Y.

San Francisco—May 24

"The Neuristor," H. D. Crane, Stanford Res. Inst., Menlo Park, Calif.

Engineering Management

Boston—May 10

Group Participation in Playing "The Business Game" with a Programmed Digital Computer.

Dayton—February 18

"PACE—A New Industrial Engineering Technique," D. N. Petersen, Northrop Corp.

Metropolitan New York—May 19

"The Role of Venture Capital in Financing," W. H. Shepard, Payson and Trask, New York, N. Y.

Engineering Writing and Speech

Boston—May 18

"What the Air Force Looks for in Engineering Reports," A. McCalmont, AFCRC, Bedford, Mass.

Instrumentation

Los Angeles—March 2

"Vibration and Its Measurement," W. Hancock and B. Munster, Endevco Corp., Pasadena, Calif.

Los Angeles—June 1

"How to Save Your Life," Rear Adm. F. Dunbar, Ret., Los Angeles County Civil Defense.
"How to Save Your Life: Nuclear Instrumentation," E. Kaufman, Litton Industries.

San Francisco—June 6

"Management for Creativity and Productivity in Electronic Instrumentation," J. M. Cage, Hewlett-Packard Co., Palo Alto, Calif.

Medical Electronics

Buffalo-Niagara—January 22

"Implantable Cardiac Pacemakers," W. Greatbach, Electronics Consultant.

Microwave Theory and Techniques

Baltimore—May 25

"Infrared Systems," R. Grove, Martin Company.

Long Island—November 24

"Multiple Feed Antennas," J. Blass, W. L. Maxson, New York, N. Y.

Long Island—January 19

"Practical Uses of Transmission Line Mapping," Dr. E. Fubini, Airborne Instruments Lab., Melville, N. Y.

Long Island—April 5

"Tutorial Discussion on Infrared Instrumentation," F. Harjes, Servo Corp. of America, Plainview, N. Y.

Long Island—May 24

"Traveling Wave Resonators," S. J. Miller, MIT Lincoln Lab.
"Coaxial Rotary Joints," C. E. Meule, Jr., MIT Lincoln Lab.

Military Electronics

Northern New Jersey—February 24

"Crossed-Field Microwave Tube," Dr. J. Feinstein, S-F-D Labs., Union, N. J.

Northern New Jersey—April 20

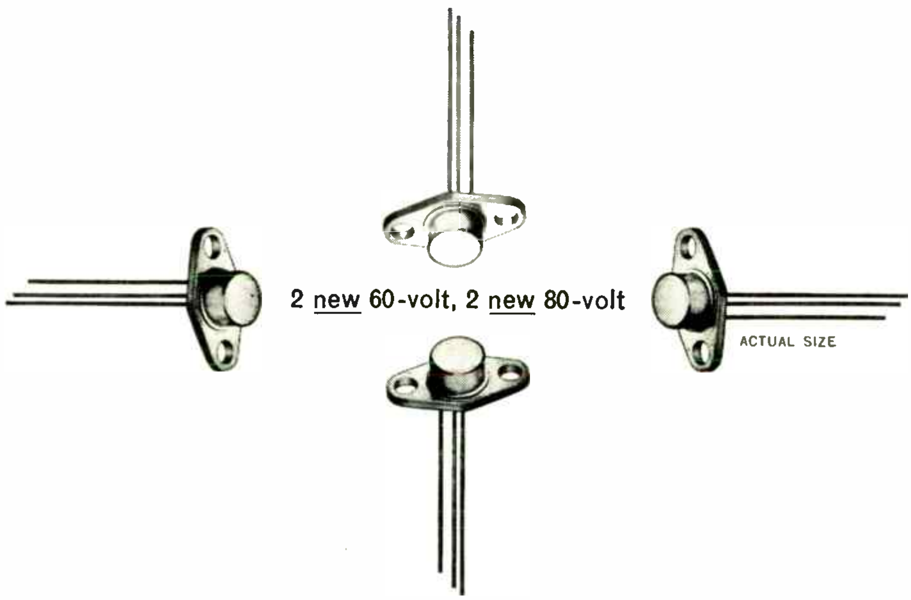
"New Applications of Parametric Devices," Dr. H. Seidel, Bell Telephone Labs., Murray Hill, N. J.

New York—October 7

"Biological Effects of Microwave Energy," Panel Discussion: Col. G. M. Kauf, Air Force Missile Test Center, Patrick AFB, Fla., Dr. J. H. Vogelman, Dynamic Electronics, N. Y., and Dr. T. S. Ely, U. S. Atomic Energy Commission.

(Continued on page 96A)

4 new miniature DELCO POWER TRANSISTORS



NOW, FROM DELCO RADIO, A COMPLETE LINE OF SMALL, HIGH-POWER TRANSISTORS!

	2N1172	2N1611	2N1612	2N1609	2N1610
V_{CB}	40	60	60	80	80
V_{EBO}	20	20	20	40	40
V_{CEO}	30	40	40	60	60
I_C	1.5 A	1.5 A	1.5 A	1.5 A	1.5 A
I_{CO}	200 μ a	100 μ a	100 μ a	100 μ a	100 μ a
H_{FE}	30/90	30/75	50/125	30/75	50/125
$V/_{Sat}$	1.0 V	1.0 V	0.6 V	1.0 V	0.6 V

The four new Delco transistors, plus the 2N1172 40-volt model, offer highly reliable operation in a new range of applications where space and weight are restricting factors.

Designed primarily for driver applications, Delco's versatile new transistors are also excellent for amplifiers, voltage regulators, Servo amplifiers, miniature power supplies, ultra-low frequency communications, citizens' radio equipment and other uses where substantial power output in a small package (TO 37) is required.

Special Features of Delco's Four New Transistors: Two gain ranges. Can be used on systems up to 24 volts. Can be mounted with the leads up or down with the same low thermal resistance of 10° C/W. Dissipation up to 2 watts at a mounting base temperature of 75°C.

Available in volume production. Write for full engineering data.

DELCO RADIO
DEPENDABILITY
RELIABILITY

Newark, New Jersey
1180 Raymond Boulevard
Tel.: Mitchell 2-6165

Santa Monica, California
726 Santa Monica Boulevard
Tel.: Exbrook 3-1465

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

Detroit, Michigan
57 Harper Avenue
Tel.: Trinity 3-6560

Division of General Motors • Kokomo, Indiana

See you at the WESCON Show, Booths 2345 & 2347

Professional Group Meetings

(Continued from page 94A)

New York—November 12

"Multimode Waveguide Discontinuities," Dr. L. B. Felsen, Microwave Research Institute, Brooklyn, N. Y.

New York—December 10

"Diode Switching for Microwaves," R. V. Garver, Diamond Ordnance Fuze Labs.

Omaha-Lincoln—February 25

"Parametric Amplifiers—Principles and Application," H. J. Peppiatt and R. A. Powell, General Electric Co.

San Francisco—April 19

"Significance of the Maser in a Comparison of Classical and Quantum Electromagnetic Theory," E. T. Jaynes, Stanford University, Stanford, Calif.

Military Electronics

Northwest Florida—February 10

"The Use of Microwave Energy to Support Space Platforms," R. L. McFarlan, President, IRE.

Military Electronics
Communications Systems
Vehicular Communications

Philadelphia—May 17

"Radio Coverage—Area Survey—Instrumentation Research," R. E. Lacy, U. S. Army Signal Corps, Ft. Monmouth, N. J.

Nuclear Science

Oak Ridge—May 19

"Cathode Ray Oscilloscope Manufacture," C. L. Bouffiou, Tektronix, Inc., Atlanta, Ga.

Production Techniques

Boston—May 24

"Translating Specifications into Finished Products," Panel Discussion: E. R. Rowlands, L. J. McCarthy, F. Seekell, C. Peterson, Cannon Electric Co. Plant tour of Cannon Electric, Salem Division.

Los Angeles—May 18

"Multi-Layer Etched Laminates in High Density Electronic Equipment," N. J. Schuster, Litton Industries, Beverly Hills, Calif.

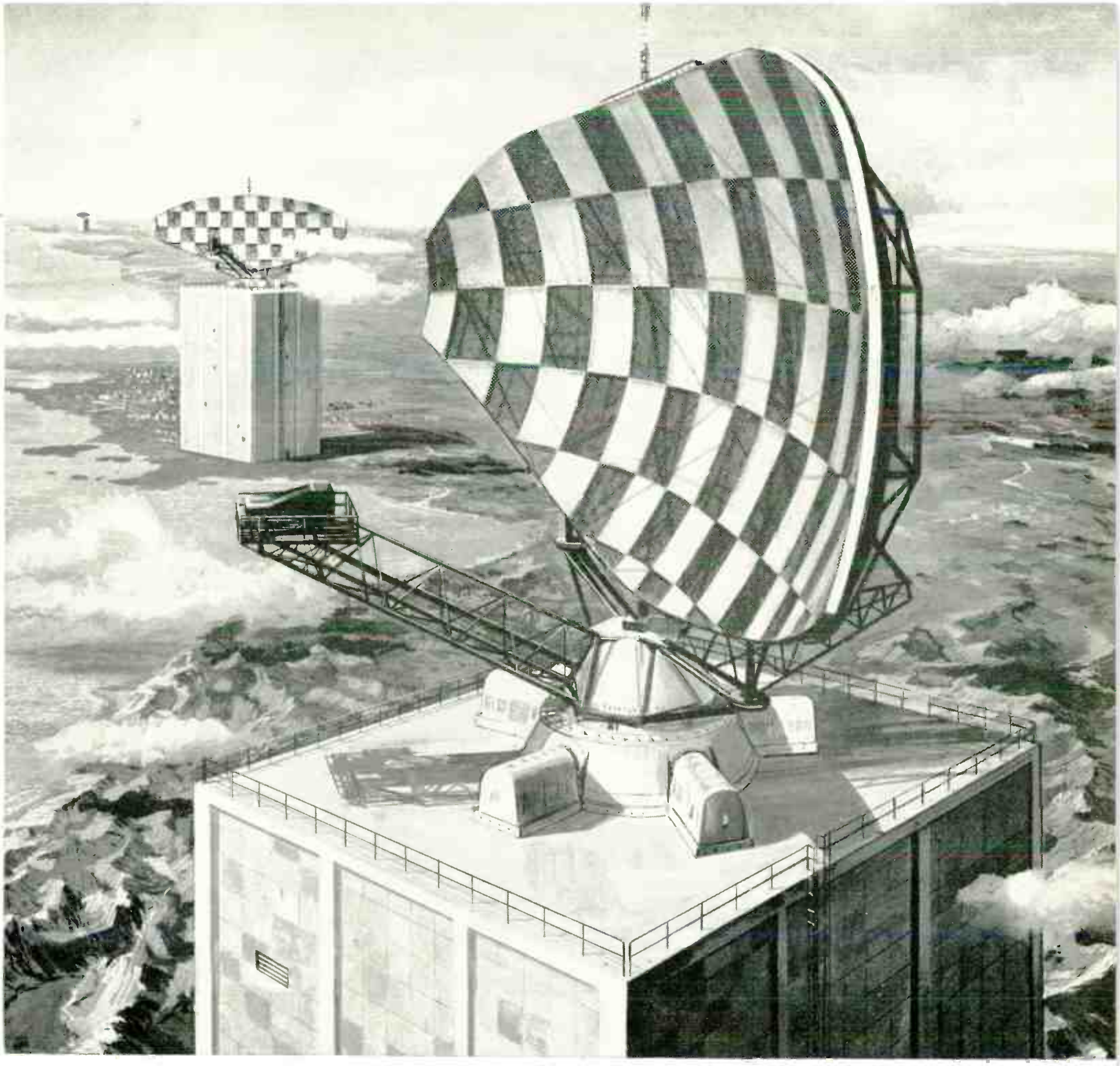
San Francisco—November 24

United Airlines Maintenance Base Tour.

San Francisco—February 2

"Production of Thin Metallic Films,"

(Continued on page 98A)



Professional Group Meetings

(Continued from page 96A)

J. R. Jennings, Spectracoat Inc., Belmont, Calif.

"Construction of Microminiature Photocells," C. Reise, Hewlett-Packard, Palo Alto, Calif.

San Francisco—February 9

"Welded Assemblies and the Use of Computers to Develop Optimum Wiring Sequence," W. H. Ayers, Sippican Corp., Marion, Mass.

San Francisco—February 23

"Microminiaturization Work Being Done at Lockheed," D. Fuller, Lockheed MSD, Sunnyvale, Calif.

San Francisco—March 1

"Panel Discussion on Future Possibilities of Microminiaturization," Dr. C. A. Rosen, SRI, Menlo Park; Maj. O. R. Hill, ARDC, Los Altos; Dr. J. R. Nall, Fairchild Semiconductor, Palo Alto.

San Francisco—April 26

Plant Tour—W. Bain, Assistant to the Factory Manager, Friden Calculator Co., San Leandro, Calif.

Reliability and Quality Control

Boston—June 2

"Malfunction Reporting," all day session.

Los Angeles—May 16

"Poly-Vironmental Effects and Measurement," R. Ratcliffe, Rototest Labs., Lynwood, Calif.

"Acoustics as a Laboratory Environment," J. Fromkin, Rototest Labs., Lynwood, Calif.

"Nondestructive Testing and Analysis by Means of X-Ray Motion Pictures," Dr. I. Rehman, Ph.D., Rototest Labs.

Space Electronics and Telemetry

Washington—May 17

"Tiros Satellite Systems," J. Lehman, RCA, Princeton, N. J.

Vehicular Communications

Detroit—May 27

"Tower Effect on Antenna Patterns," Dr. R. Yang, Andrew Corp. Chicago, Ill.

Houston—April 12

"Directional Antennas and Phasing Networks," R. S. Bush, Gates Radio Co., Quincy, Ill.

Los Angeles—May 19

"Microwave as Applied to Two Way Systems," S. Combs, RCA, Los Angeles, Calif.

Vehicular Communications Communications Systems

Philadelphia—February 9

"An Integrated Traffic Control System," C. J. Schultz, Motorola, Inc., Scottsdale, Ariz.

Section Meetings

Albuquerque-Los Alamos

"The Los Alamos Space Research Program," John Northrup, Los Alamos Scientific Lab. 5/20/60.

Annual Picnic—Installation of Officers; 6/4/60.

Anchorage

"Electro-Physiology of the Nervous System," P. A. Mead. 5/2/60.

Atlanta

"Electronic Engineers in Management," D. J. Morrissey, Lockheed Aircraft Corp. 5/27/60.

Baltimore

"Use of High Speed Computers in Predicting Hospital Needs of a Community," John Moss, Johns Hopkins Univ.; Election of Officers 6/13/60.

Bay of Quinte

"IRE Today and Tomorrow," J. N. Dyer, IRE Vice-President 6/16/60.

Binghamton

"Theoretical Techniques for Adaptive Control Systems Design," G. W. Johnson, IBM. 5/23/60.

"An Objective Appraisal of Essential National Strengths," M. J. Kelly; Election of Officers. 6/16/60.

Central Florida

"PERCOS Performance Coding System," Ernest Keller, Motorola, Inc. 5/19/60.

Central Pennsylvania

"State of the IRE and its Position in the Scientific Community," John Dyer, IRE Vice-President; "Satellite Studies of the Ionosphere," W. J. Ross, Pennsylvania State Univ. 5/17/60.

"Electronic Switching in the Bell System," L. A. Moretz, Bell Telephone Co. 5/19/60.

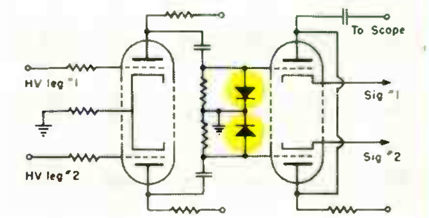
Columbus

"A Look at Russia Through the Eyes of an Engineer," E. A. Harty, Power Equipment Co.; Installation of new Officers. 6/7/60.

(Continued on page 100A)

GENERAL INSTRUMENT SEMICONDUCTOR REPORT

Design Notes...

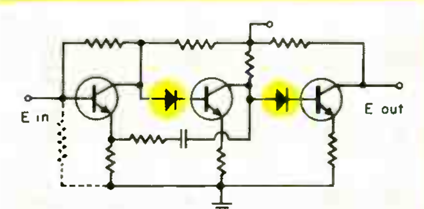


DIODE CLAMPS PREVENT DRIFT IN HIGH SPEED ELECTRONIC SWITCHES The problem of drift in high speed electronic switches has been solved through the use of clamp diodes. This application is especially useful for stabilizing the operation of go, no-go oscilloscopic testing of dynamic parameters in a variety of electronic components.

In the circuit, General Instrument MP-300 silicon diode/rectifiers may be used because of their superior stability and low reverse leakage (only .05 μa @ 25°C).

Changing from vacuum tube to silicon clamp diodes minimizes problems associated with varying contact potentials. Equipment reliability is improved since total thermal dissipation is reduced. Further, equipment does not have to be reset in case of power line failure.

The small physical size of General Instrument diode/rectifiers is important where a large number of switches are to be used in a single piece of equipment.



NOVEL CIRCUIT USES DIODES FOR AUDIO COUPLING There are many benefits to be gained through *diode coupling* of audio amplifiers. The simplified three-stage transistorized audio amplifier shown above uses General Instrument IN645 subminiature silicon diodes.

Since the diodes are forward biased, ac is virtually direct coupled—resulting in a flat frequency response limited only by transistor parameters. Need for large coupling capacitors is eliminated. Virtually lossless ac coupling is obtained. And, temperature stability is improved because of low external base resistance.

Complete schematics of above circuits are available upon request.

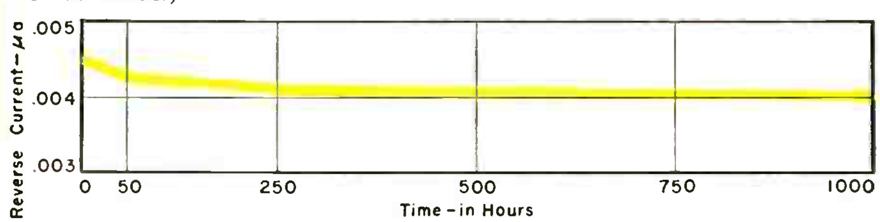
Proved Reliability: ZERO FAILURES after 11,000 hours operation at 150° C!

IN 645-IN 649 DIODE/RECTIFIERS
AVAILABLE IN PRODUCTION QUANTITIES... EXCEED USAF STANDARDS

General Instrument 1N645 through 1N649 subminiature rectifiers are ideally suited for applications requiring small size and very high reliability. These hermetically sealed glass units are designed to operate over an ambient range from −65° to 150°C . . . pass MIL-E-1/1143 specfications for breakdown voltage . . . offer superior life test performance. This series covers the range of 225 to 600 PIV, with maximum average rectified current of 400 ma @ 25°C. Maximum reverse current @ PIV is only 0.2 μa.

These diode/rectifiers are subjected to 100% environmental testing and dynamic oscilloscopic tests to assure high electrical and mechanical uniformity, surpassing the most stringent military specifications.

■ **LIFE TESTS** indicate outstanding stability of the General Instrument 1N645 series subminiature rectifiers under load. Graph shows results of a 1,000-hour test of 231 units from a normal production run. (Conditions: V_{RMS} 160 V ac; Io 400 ma dc.)



NEW MP SERIES DESIGNED FOR 200°C OPERATION!

General Instrument has achieved an outstanding power-to-size relationship in the high quality MP silicon diode/rectifier series. Parameters for these subminiature glass units are suitable for a wide range of applications under high-temperature conditions:

TYPE	PIV	DC OUTPUT CURRENT (Ma) 25°C	DC OUTPUT CURRENT (Ma) 200°C	REVERSE LEAKAGE (μa) @ PIV 25°C	REVERSE LEAKAGE (μa) @ PIV 200°C	FORWARD DROP @ 400 Ma @ 25°C
MP 100	100	400	50	.05	75	1.0
MP 225	225	400	50	.05	75	1.0
MP 300	300	400	40	.05	75	1.0
MP 400	400	400	35	.05	75	1.0
MP 500	500	400	25	.05	75	1.0
MP 600	600	400	20	.05	75	1.0

CALL ON GENERAL INSTRUMENT for technical data and applications assistance on the complete line of G.I. high reliability silicon diodes.

SEMICONDUCTOR DIVISION
GENERAL INSTRUMENT CORPORATION
65 Gouverneur Street, Newark 4, New Jersey

4448

Section Meetings

(Continued from page 98A)

Dallas

"Bioinstrumentation for Space Vehicles," Robert Adams, School of Aviation Medicine, 6/7/60.

"Alternating Current Spectroscopy of Biomatter; Basic and Applied Work," Herman Schwan, University of Pa. 6/14/60.

Detroit

"Island Holiday," Nate Reiss; Election of Officers. 5/27/60.

Elmira-Corning

"Presidential Address," Dr. R. L. McFarlan, President of the IRE. 5/18/60.

El Paso

Group Discussion on Programs & Meeting Attendance. 4/28/60.

"After Demobulation—What Then?" J. H. Maury, Mt. States Tel. & Tel. Co. 5/26/60.

Evansville-Owensboro

Annual Picnic Meeting. 6/12/60.

Florida West Coast

"Test Equipment," George Frederick, Hewlett-Packard Co. 5/18/60.

Fort Huachuca

"Satellite Navigation," Dr. Marner, Collins Radio Co. 5/23/60.

"Tropo-Scatter Communications," J. C. Domingue, USAEPG; Election of Officers. 6/13/60.

Indianapolis

"Thin Film Micro-Electronics," E. N. Skomal, Motorola, Inc. 5/26/60.

Israel

"The Emergency Radio-Telephone Network in Border Zones," A. Bar-El, Israel Army. 11/3/59.

Annual Meeting—Election of Officers. 12/15/59.

Ithaca

"Some Impressions of Poland," Mark Kac, Cornell Univ.; Election of Officers. 5/26/60.

Las Vegas

"Filtering & Equalization in Optics," Barney Oliver, Hewlett-Packard Co. 5/31/60.

Long Island

"Inertial Navigation Equipment," Mr. Plevin & Representatives of Arma, American Bosch Arma Corp.; Tour through Arma plant. 6/2/60.

Los Angeles

"The Role of Industry in the Space Age," W. H. Pickering, Caltech, Jet Propulsion Lab. 4/21/60.

"Challenging New Horizons for Space Age Engineers," W. J. Scarpino, Point Arguello; "The Navy in the Space Age," J. P. Monroe, Pacific Missile Range; Tour of Point Arguello. 5/13/60.

Lubbock

"National Defense," Glenn Scott, Southwestern Bell Tel. Co. 5/17/60.

Milwaukee

"Communication Boundaries on Electronic Engineering in a Social Environment," Panel—Walter Richter, Consulting Engineer; R. J. Jones, Mich. Mining Tech.; J. W. Nelson, General Elec. X-ray; P. Mundie, Industrial Psychologist; Annual IRE Student Award Night. 5/11/60.

Montreal

"The IRE and Its Role in Medical Electronics," Dr. R. L. McFarlan, President of the IRE. 6/8/60.

New Orleans

"Digital Transmission on the DDD Telephone Network," J. H. Felker, AT&T—Joint meeting with AIEE. 6/9/60.

North Carolina

"A Survey of the Hi-Fi Components Field Today," W. G. Trayer, Dalton-Hege Radio Supply Co.; "Multiplexing Applications to Stereophonic Systems," Robert Linz, GE Co. 5/20/60.

Northwest Florida

"PERCOS Performance Coding System of Methods & Devices," Ernest A. Keller, Motorola Inc. 5/17/60.

Oklahoma City

"Communications Satellite," A. T. Mayle, Jr., Farnsworth; Joint Meeting with AIEE & AFCEA. 4/21/60.

Film: Laying of the Transatlantic Cable Furnished by Bell Tel. Co., George Holt, SW Bell Tel. Co.; Election of Officers. 5/10/60.

Annual IRE-AIEE Joint Student Banquet & Student Paper Contest Awards Meeting. 5/20/60.

Orlando

"The Project System vs. The Technical Section," A. R. Gray, Martin Co.; Election of Officers. 5/18/60.

Philadelphia

"Determining the Vulnerability of Military Equipment to Electronic Countermeasures," R. H. Sugarman, U. S. Army Signal Research & Dev. Labs. 5/4/60.

Phoenix

"Filtering & Equalization in Optics," B. M. Oliver, Hewlett-Packard, Inc. 5/17/60.

Portland

"Attendance of American Exhibition in Moscow," Joe Roizan, Ampex. 5/17/60.

"The Contribution of the IRE to Medical Electronics," Dr. R. L. McFarlan, President of the IRE. 5/25/60.

Rochester

"Electronic Music," D. W. Martin, Baldwin Piano Co. 5/19/60.

St. Louis

"Project Eclipse," Michael Witunski, McDonnell Aircraft Corp. 12/8/59.

"Radio Astronomy," Dennis Walsh, Observatory, Univ. of Michigan. 1/19/60.

"Your IRE," C. E. Harp, Univ. of Oklahoma. 2/16/60.

"Practical Antenna Applications of Microwave Components," T. N. Anderson, Airtron, Inc. 4/12/60.

"Basic Logic Circuits," W. J. Murphy, St. Louis Univ.; "A Synchronized Television Signal Analyzer," G. R. Couranz, J. L. Wyland, Washington Univ.; "Generation of Periodic Wave Forms Using Fourier Synthesis," N. Koening, Univ. of Mo. 5/18/60.

San Antonio-Austin

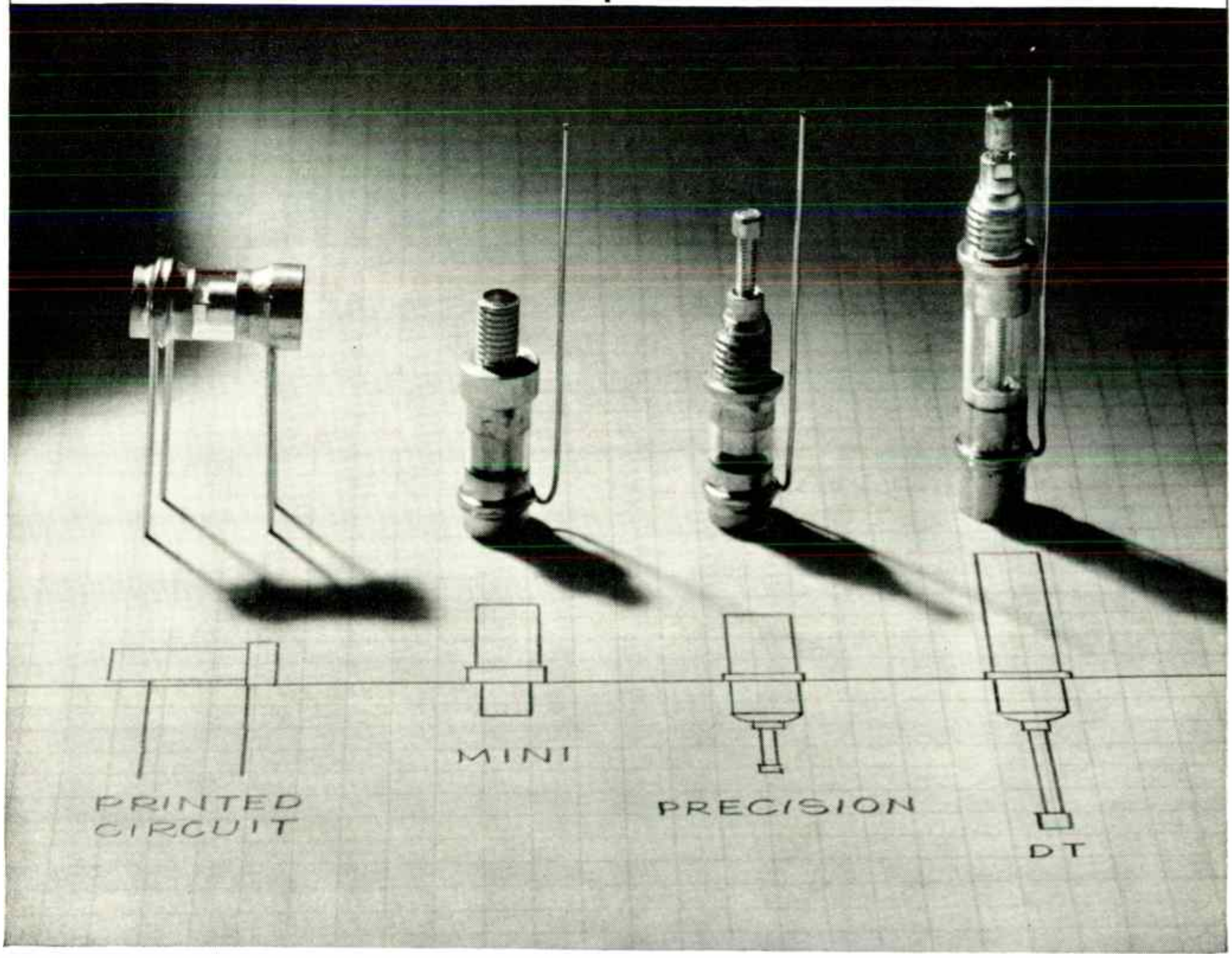
"Semiconductor Electronics," Ernest Wuerst, Texas Instruments, Inc. 3/7/60.

"Diode and Transistor Action," B. M. Williams, Texas Instruments, Inc. 3/14/60.

(Continued on page 102A)

TYPICAL VALUES FOR DIRECT TRAVERSE TRIMMERS	
DC Volts	1000
Dielectric Strength	1500
Megohms IR	10^6
Q Factor @ 50 MC	500
TC	±50 to ±100

	Range	Mini-Trimmers	Precision DT	Standard DT
SIZES	0.5-3	—	—	X
	1-4	X	X	—
	1-8	X	X	X
	1-12	X	X	X
	1-18	X	X	—
	1-30	—	X	—



FOR BACKLASHLESS LINEAR TUNING
FOUR DIRECT TRAVERSE TRIMMERS

Some years ago we came up with the concept of a *direct traverse* trimmer capacitor . . . a unit in which the core slides in and out without rotating.

This motion results in a completely linear tuning curve, utterly devoid of capacitance reversals or backlash.

The mechanism compensates for wear automatically, so torque is maintained at all positions. Cores cannot work loose and become microphonic. It's impossible for capacitance to change, even under shock and vibration.

There is no tuning breakage.

We added to this direct traverse motion the many values of glass. No other material combines such high reliability with such low TC. Or such precision at low cost. Specs and performance speak for themselves . . . eloquently.

Just recently we added three new models to our standard direct traverse line:

THREE NEW DIRECT TRAVERSE TRIMMERS

NEW MINI-TRIMMERS Where space is tight both in front and behind the panel. Fixed cavity tuning keeps screw enclosed at all times. Only 0.40 *uuf* change per turn.

PRINTED CIRCUIT MINIS Same specs as mini-trimmers, but designed for compacting printed circuits.

NEW PRECISION DT TRIMMERS Even shorter behind the panel, slightly longer hardware. 0.50 *uuf* change per turn.

STANDARD DT TRIMMER When you have more space to play with than money, the best buy in reliable trimmers is still our standard direct traverse model. 0.60 *uuf* change per turn.

OFF-THE-SHELF DELIVERY For quantities of less than 1000, contact a local distributor serviced by Erie Distributor Division.

For specs, Corning Glass Works, 542 High St., Bradford, Pa.

SEE US AT WESCON . . . Booths 853-854

CORNING ELECTRONIC COMPONENTS

CORNING GLASS WORKS, BRADFORD, PA.

Section Meetings

(Continued from page 100A)

"Radar Moon Reflection Studies," A. W. Straiton, Univ. of Texas. 3/18/60.

"Transistor Internal Parameters & Device Design," R. S. Stewart, Texas Instruments, Inc. 3/21/60.

"Useful Parameters & Basic Transistor Circuits," W. T. Jones, Texas Instruments, Inc. 3/28/60.

"Oscillators & R.F. Circuitry," G. E. Penisten, Texas Instruments, Inc. 4/18/60.

"The IRE and its Role in Medical Electronics," Dr. R. L. McFarlan, President of the IRE. 4/27/60.

"Network Design & Feedback Amplifiers," W. E. Crisman, Texas Instruments, Inc. 4/4/60.

"Digital & Pulse Circuits," Harvey Cragnon, Texas Instruments, Inc. 4/11/60.

San Diego

"Geodetic SECOR for World Mapping," R. V. Werner, Cubic. 4/6/60.

Schenectady

"Measuring Radio Interference," Joseph Lorch, Empire Devices, Inc. 5/24/60.

Shreveport

"Digital Data Transmission," J. H. Felker, AT&T.; Election of Officers. 6/7/60.

Virginia

"Microminiaturization," Mr. Doctor, Diamond Ordnance Fuze Labs.; Election of Officers. 4/8/60.

"High Resolution Television Viewing System," J. F. Fisher, Philco Corp. 4/22/60.

"New Concept in Stereo," Paul Weathers, Weathers Industries. 5/6/60.

"The Electronic Speech Analyzer & Comparator," W. H. Darnell, III, Student V.P.I.

"RMS Volt Meter Design," Marvin Perlman, University of Virginia; Annual Student Competition Award. 5/13/60.

Washington

"Space Instrumentation," R. B. Kershner, Johns Hopkins Univ. 4/4/60.

"Molecular Electronics," Charles Feldman, Melpar, Inc. 5/2/60.

Western Michigan

"Electronics in Air Navigation," W. A. Hawkins, Federal Aviation Agency. 11/11/59.

"Radiation Biology," David Scobie, Hope College. 12/9/59.

"Electronics in Mechanized Inspection," M. J. Applegate, Vari-Tech Co. 1/13/60.

"Electronics in Chemistry," James Dye, Michigan State Univ. 2/10/60.

"Nuvistors," David Lovcik, RCA. 3/9/60.

"A New Microphone," Robert Ramsey, Electro-Voice Co. 4/13/60.

"Compatible Color TV," J. W. Wentworth, RCA. 5/11/60.

Winnipeg

"Data Transmission on Micro-wave Systems," Monte Bramhall, Automatic Electric Co. 4/29/60.

Field Visit to Canadian Pacific Telegraph's Communications Installations. 5/18/60.

SUBSECTIONS

Buenaventura

"Space Age Tracking," Carl Nielsen, Hallamore Electronics; "Phase Lock Application at 'X' Band," Richard Hartenstien, Hallamore Electronics; Student Award. 4/13/60.

Lancaster

"Satellite Communication," C. C. Cutler, Bell Telephone Lab. 4/21/60.

Lehigh Valley

"Evaluating High Fidelity from the Consumer's Point of View," D. M. Berk, Consumer's Research, Inc.; Election of Officers. 4/27/60.

Merrimack Valley

"A New Technique for Mapping the Moon Using Radar," G. H. Pettingill, MIT Lincoln Lab.; Election of Officers. 4/18/60.

Mid-Hudson

"Parametric Amplifiers," J. C. Greene, Airborne Instruments Lab.; Election of Officers. 4/20/60.

Monmouth

"Principles & Applications of the Esaki Diodes," W. W. Anderson, Bell Telephone Labs., Inc.; Election of Officers. 5/18/60.

Northern Vermont

"Fuel Gaging, Injection Pumps & Other Products," Harrison Edwards, Simmonds Accessories Inc. 5/23/60.

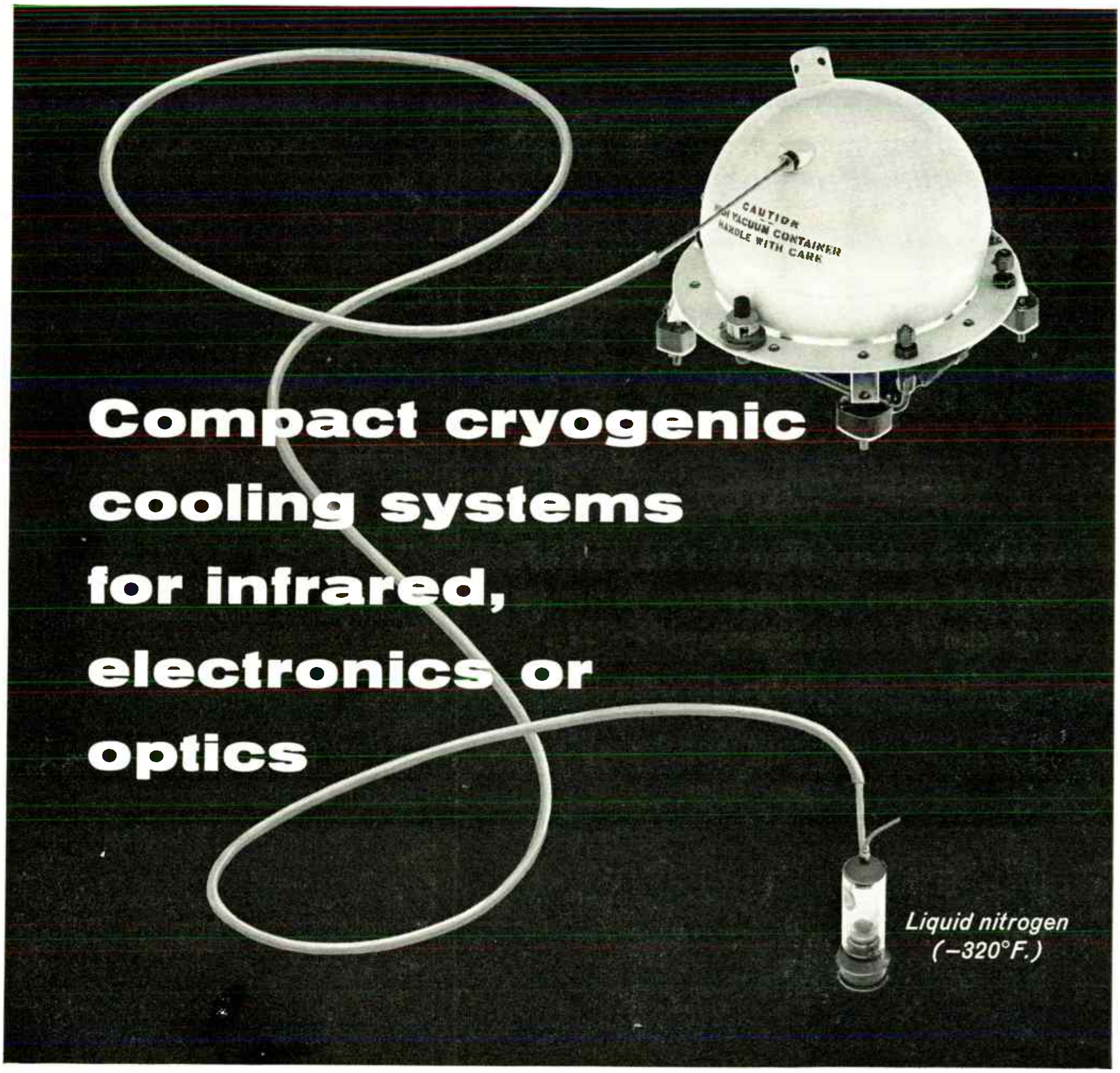
Orange Belt

"Hello Around the World," W. R. Simmons and Kenneth Lee, Pacific Telephone Co. 2/9/60.

"Problems of Space Travel," Harry Lass, Jet Propulsion Lab.; "Earth Movement," Hugo Benioff, Cal. Tech. Joint Meeting with Pasadena Subsection. 3/15/60.

"Evolution of Microminiaturization," W. D. Fuller, Lockheed Missile & Space Div.; Tour of Lockheed Aircraft Trainer Div. 5/10/60.

(Continued on page 104A)



New AiResearch system delivers nitrogen in liquid form from storage system to cooling area

Now units requiring cryogenic cooling no longer need be designed with allowances made for bulky expanders or adjacent storage tanks.

The new AiResearch system transfers the coolant *in liquid form* to a point of use *25 feet or more away*. The liquefied gas passes through an uninsulated, small, flexible tube which can be bent over and around obstructions. Because the storage system can be placed anywhere, space limitations are overcome and vehicle installation problems are simplified.

The complete system includes the cryogenic liquid container, pressure and flow controls, the liquid transfer tube and cooling adapter. The system can be operated without external power. It can be used with missile, aircraft, space or ground based units and can be converted to a closed-cycle system with the addition of a small gas liquefier.

AiResearch has pioneered many new developments in the cryogenic field. It is presently engaged in work on systems utilizing helium, hydrogen or neon as coolants, and cryogenic systems for zero G operation.

• *Please direct inquiries to Los Angeles Division.*

THE GARRETT CORPORATION

AiResearch Manufacturing Divisions

Los Angeles 45, California • Phoenix, Arizona

Systems and Components for: AIRCRAFT, MISSILE, SPACECRAFT, ELECTRONIC, NUCLEAR AND INDUSTRIAL APPLICATIONS

Section Meetings

(*Continued from page 102A*)

PASADENA

"Hello Around the World," W. R. Simmons and Kenneth Lee, Pacific Telephone Co. 2/9/60.

"Problems of Space Travel," Harry Lass, Jet Propulsion Lab.; "Earth Movement," Hugo Benioff, Cal. Tech. Joint Meeting with Orange Belt Subsection, 3/15/60.

RICHLAND

"The Future of Nuclear Business," Lyman Fink, GE Co. 2/24/60.

"The Evaluation & Testing of the PRTR Controller, Utilizing an Analog Computer," William Cameron, GE Co. 3/30/60.

"The Measurement of Radioactivity in People," W. C. Roesch, GE Co.; Election of Officers, 5/25/60.

SANTA ANA

"X-15 Research Aircraft & B-70 Valkyrie," A. White, North American Aviation, Inc. 5/17/60.

WESTCHESTER

"Semiconductor Properties at Very Low Temperatures & Some Device Applications," S. H. Koenig, IBM Watson Lab. 5/18/60.

Membership

The following transfers and admissions have been approved and are now effective:

Transfer to Senior Member

Abjanich, G. S., Hicksville, L. I., N. Y.
Franklin, G. F., Stanford, Calif.
George, N., Pasadena, Calif.
Gutton, H., Suresnes, Seine, France
Hooper, J. R., Jr., Cleveland Heights, Ohio
Hunicke, R. L., Ambler, Pa.
Knowles, E. R., Bisbee, Ariz.
Leinkram, C. Z., Asbury Park, N. J.
Lielbriedis, K., Jamaica, L. I., N. Y.
Ludvigson, M. T., Cedar Rapids, Iowa
Machlis, J., Granada Hills, Calif.
Martin, L. H., Concord, Mass.
Meyers, R. E., San Gabriel, Calif.
Moore, H. S., Alexandria, Va.
Okada, R. H., Perkiomenville, Pa.
Parsons, R. L., Dallas, Tex.
Pearson, H., New York, N. Y.
Roberts, W. W., Charlottesville, Va.
Rosenblatt, H., Baltimore, Md.
Sick, J. R., Dayton, Ohio
West, J. P., McLean, Va.
Widell, C. D., Encino, Calif.
Worthen, K. J., Lynchburg, Va.
Yarnell, W. M., Bethpage, N. Y.

Admission to Senior Member

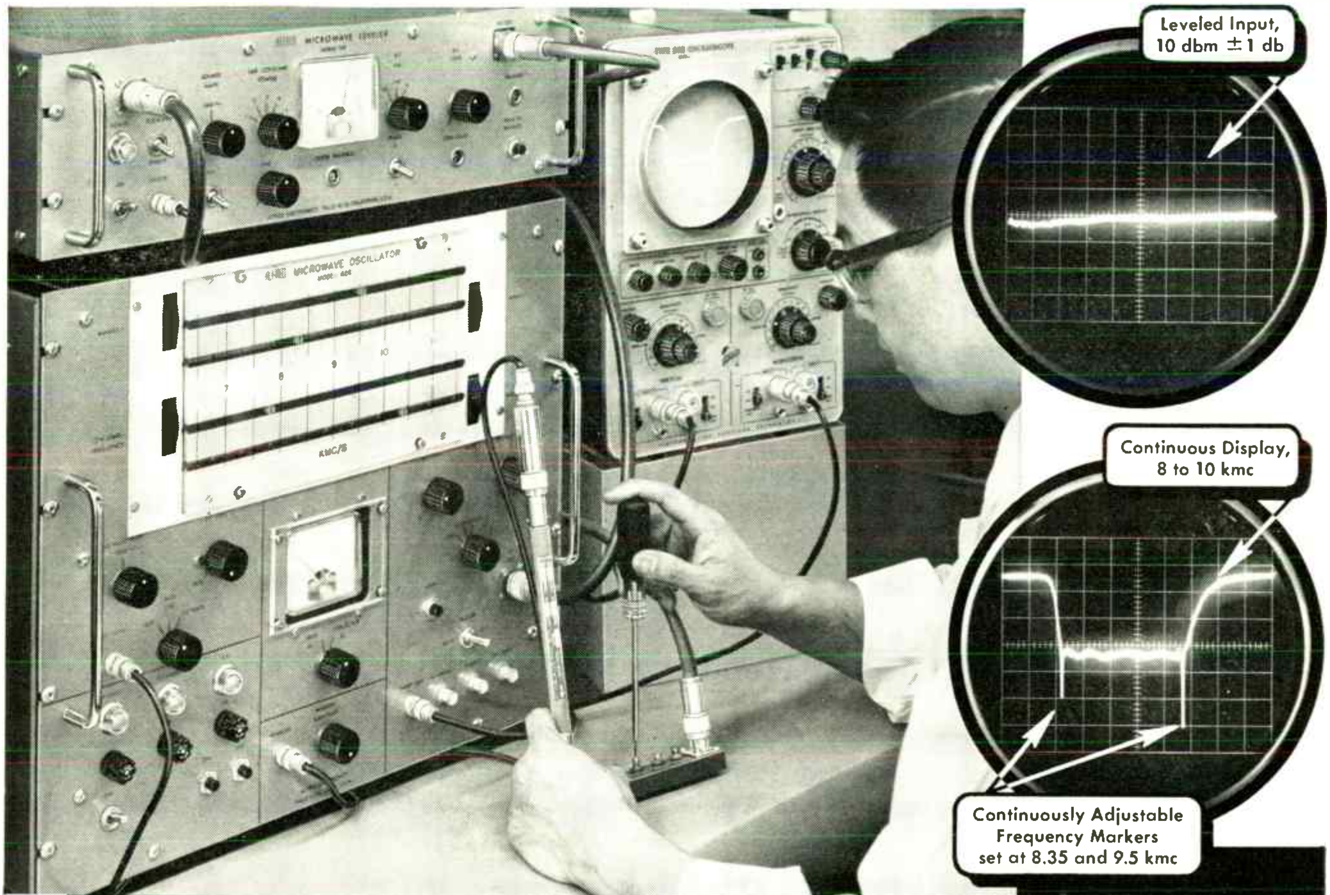
Adam, R. P., San Pedro, Calif.
Amara, D., Sherman Oaks, Calif.
Arifin, Z., Djakarta, Indonesia
Beddoes, M. P., Vancouver, B. C., Canada
Boxma, Y., Den Haag, Netherlands
Bradfield, H. J., Plainfield, N. J.
Brown, C. M., Syracuse, N. Y.
Cokas, G., Redondo Beach, Calif.
De Mott, E. G., Chatham, N. J.
Frost, R. F., Alexandria, Va.
Harrison, F. H., Jr., Dallas, Tex.
Hattersley, F. L., Los Angeles, Calif.
Hodara, H., Morton Grove, Ill.
Holscher, D. J., Playa del Rey, Calif.
Ibison, R. A., St. Petersburg, Fla.
Liss, S., Fair Lawn, N. J.
Miller, F. N., Fort Huachuca, Ariz.
Osborn, W. C., Glendale, Calif.
Paynter, H. M., Cambridge, Mass.
Piket, J., Den Haag, Netherlands
Robertson, J. O., Ann Arbor, Mich.
Rollosson, G. W., Albuquerque, N. M.
Shank, H. C., Glendale, Calif.
Smith, D. S., La Mirada, Calif.
Stanz, P. F., Pacoima, Calif.
Tanner, W. P., Jr., Ann Arbor, Mich.
Weiman, L., Pasadena, Calif.
Wolfe, I. R., Palo Alto, Calif.

Transfer to Member

Abourezk, W., Hawthorne, Calif.
Abrams, S. J., Brooklyn, N. Y.
Alonso, R. L., Cambridge, Mass.
Andros, L. R., Moorestown, N. J.
Bell, J. H., Florissant, Mo.
Berg, R. S., Syosset, L. I., N. Y.
Bissell, H. M., Jr., Murray Hill, N. J.
Blaski, M. F., Whittier, Calif.
Bowles, K. L., Boulder, Colo.
Braunlin, R. J., Indianapolis, Ind.
Brown, E. A., Los Alamos, N. M.
Butala, E. F., North Linthicum, Md.
Caffrey, T. P., Syracuse, N. Y.
Calhoun, S. H., Tonawanda, N. Y.
Close, J. N., Utica, N. Y.
Constantinides, J. S., Stanford, Calif.
Cole, C. F., Jr., Ruston, La.
Cox, H. N., Cambridge, Mass.
De Pasquale, N. D., Winter Park, Fla.
De Pew, R. T., Albuquerque, N. M.
DuWane, W. H., Klamath Falls, Ore.
Duerig, W. H., Nutley, N. J.
Dukes, J. R., Columbus, Ohio
Dunn, J. R., Rolling Meadows, Ill.
Falconer, S., Manchester, Mass.
Ferguson, R. C., Bellevue, Wash.
Forest, O. J., Atlantic Highlands, N. J.
Gardner, R., Brooklyn, N. Y.
Hackenberger, R. B., Concord, Mass.
Hamburger, T., Baltimore, Md.
Handy, R. M., Evanston, Ill.
Hart, C. E., Fairview Park, Ohio
Hart, D. M., Northampton, Mass.
Howard, P. H., Apalachin, N. Y.
Hupp, C. E., Evanston, Ill.
Jodon, F. R., Bethesda, Md.
Kaiser, H. F., Washington, D. C.
Kincaid, N. L., Hawthorne, Calif.
Klimberg, J., Sunnyvale, Calif.
Knouse, R. H., Hatboro, Pa.
Laughlin, C. R., Jr., Silver Spring, Md.
Leslie, R. E., East Northport, L. I., N. Y.
Levy, A. J., Orlando Fla.
Lorenz, R., Culver City, Calif.
Meckfessel, E. F., Los Angeles, Calif.
Mielziner, W., Denver, Colo.
Moran, W. R., Jr., Gaithersburg, Md.
Moreira, N. M., South Farmingdale, L. I., N. Y
Neveroski, F., North Vancouver, B. C., Canada

(*Continued on page 106A*)

Testing insertion characteristics of X-band filter with Alfred Swept Generator. It consists of Alfred Microwave Oscillator and Alfred Microwave Leveler. This combination electronically sweeps frequency linearly with respect to time while maintaining RF power virtually constant across the band.



Save Test Time, Assure Test Accuracy

with ALFRED'S new SWEPT Microwave Generator...........

The scope patterns tell the story. Top pattern shows constant power input from Alfred Swept Generator to component (filter) under test. With known input, variation in output is due to filter characteristics. Lower pattern is especially significant, showing continuous, flicker-free display, 8 to 10 kmc. Any changes in stubs or irises are immediately reflected. Measurement accuracy is assured *at every frequency*, not just at selected points.

THIS TECHNIQUE CAN BE USED FOR MOST PRESENTLY KNOWN MICROWAVE TESTING APPLICATIONS. HERE'S WHY IT'S FASTER THAN CONVENTIONAL SIGNAL GENERATORS:

* *Continuous Display* allows immediate measurements—no plotting needed. Trace can be recorded if desired.

* *Sweep Technique* eliminates time-consuming "point-to-point" frequency and power setting methods of conventional signal generators. Sweep range is continuously adjustable with 1% accurate Direct Reading Slide Rule Dial.

* *"Quick Look Readout"* eliminates calculations in setting sweep range.

* *Adjustable Frequency Markers* allow rapid, broadband calibration of scope or recorder trace.

SOME MORE FACTS YOU SHOULD KNOW

* *Frequency Ranges.* The Swept Generator is available in five ranges to 12.4 kmc—1 to 2, 2 to 4, 4 to 8, 7 to 11, 8.2 to 12.4.

* *Stability.* At any single frequency, stability of the Swept Generator equals that of a conventional signal generator. Spurious modulation is low.

* *Power Output.* Greater than a signal generator: 10 milliwatts as compared to 1 milliwatt.

FUNCTION OF THE LEVELER It holds power output constant to ±1 db over standard frequency ranges, and better than ±1 db over narrower ranges. The Leveler serves as a broadband attenuator with up to 20 db dynamic range control, providing constant output over a wide range. It can be used as a general purpose instrument for a wide variety of oscillators and amplifiers.

For more details on the Alfred Swept Generator—please contact your Alfred sales engineering representative, or write direct.

Key specifications for Signal Generators available for coverage from 1 to 12.4 kmc

Frequency—Controls: Continuously adjustable with direct calibrated dial. Calibration accuracy: 1%. Stability: ±0.02%/hr. Residual FM: ±0.0025%. **Power Output** (Minimum): 10 mw ±1 db. Continuously adjustable from zero to maximum. Attenuation Range: Up to 20 db. **Sweep**—Selector: Recurrent Sweep, Single Sweep, Single Frequency, and External on panel switch. Time: 100 to .01 seconds, continuously adjustable. **Monitor Output**—Sweep Out: Positive linear sawtooth, 45 volts peak. Panel BNC connector. **Amplitude Modulation**—Internal Square Wave: RF output is alternately 0 and unmodulated CW value. Frequency 800 to 1200 cps, adjustable by panel control.

See us at Wescon, Booth 630

7

ALFRED ELECTRONICS
897 COMMERCIAL STREET
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(Continued from page 104A)

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Bechert, T. E., Paoli, Pa.
Boynton, P. M., Seattle, Wash.
Bridges, W. A., Paris, France
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Brown, R. L., Jr., Richland, Wash.
Buckmaster, H. A., Calgary, Alberta, Canada
Buetow, R. C., Mount Prespect, Ill.
Buckvar, B., Lexington, Mass.
Byrne, E. L., Northridge, Calif.
Campaigne, H. H., Jessup, Md.
Cantwell, T., Lexington, Mass.
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Casey, J. P., Amityville, N. Y.
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Chantrain, N., Antwerp, Belgium
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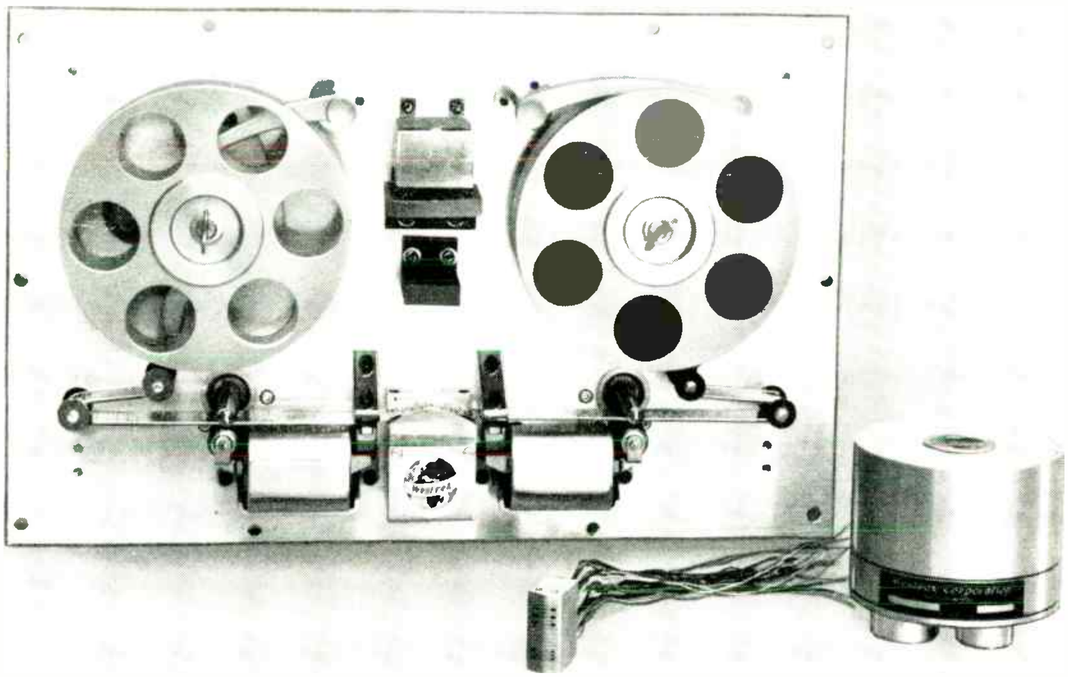
Flaherty, J. E., Dayton, Ohio
Flemming, K. F., Linthicum Heights, Md.

(Continued on page 109A)

A WESTREX BRIEFING...

To your specifications for recording, storing, and recovering data, Westrex brings more than a quarter of a century of experience. Our major disciplines are (1) electronics, (2) mechanics, as needed in mechanical design for precise tape-pulling mechanisms, and (3) optics. Here, briefed, are descriptions of some of our new products...



PERFORATED TAPE READERS A new Westrex perforated tape reader handles 8 levels of information; reads in both directions at continuous speeds up to 1400 characters per second; stops on character at 200 characters per second. Starting time 22 milliseconds, stopping time 1 millisecond. Remote control operation and logic level output to meet your specific needs. Solid state electronics, with miniaturized etched circuit, plug-in modules. End of tape sensing by logic read out.

MAGNETIC HEADS These include multiple section instrumentation heads; memory drum heads; and erase-record-reproduce assemblies for applications that range from sound systems to missiles. Catalog items or custom-built units to your requirements. Our experience assures proper utilization of design factors that most economically meet your needs. Consideration of your special requirements, such as high crosstalk rejection, stability under extreme environmental conditions, and precise mechanical tolerances, are a part of our service to customers. What are your needs?

MINIATURE AIRBORNE TAPE RECORDERS Designed to withstand impacts of 1500 G's, a new Westrex miniature recorder can simultaneously record and monitor 14 tracks of information. With 14 tracks to the inch, unique shielding provides a crosstalk ratio of over 40 db at 5000 c.p.s. Precise gap alignment, obtained by optical lapping methods, maintains gap scatter within plus or minus 50 microinches. The positively-driven tape-pulling mechanism, and virtually continuously supported tape, are features which reflect our unique and proprietary knowledge in this field. The entire hermetically-sealed recording unit is contained in a single cylinder 3 inches high and 4 inches in diameter.

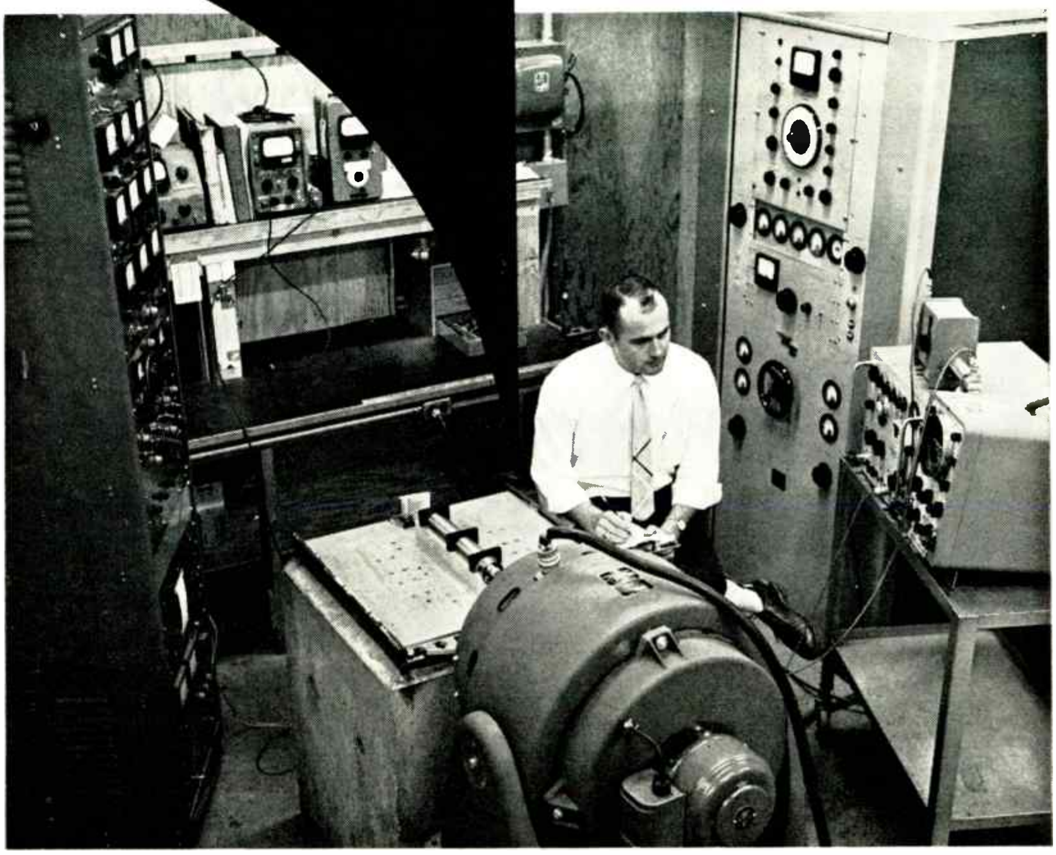
For information on this and other Westrex products, address your inquiry to Mr. L. A. Call, Westrex Corporation, Recording Department, 6601 Romaine Street, Hollywood 28, California.

Westrex Corporation

A DIVISION OF LITTON INDUSTRIES

RUGGEDIZATION

Huggins HA 20 undergoes Vibration tests



TRAVELING WAVE TUBES

Tests prove that Huggins TWT's can be provided to operate at 30G shock; 90,000 ft. altitude; 10G vibration to 2000cps with 0.5DB max. amplitude modulation; −55°C to +71°C with ±2DB variation in gain.

A world leader in research and production of TWT's, the Huggins line is applicable to all phases of military equipment . . . ground, shipboard, and airborne.

Whatever your project, if it involves TWT's, call on Huggins for complete assistance.

Visit us at Wescon, Booth 820-821

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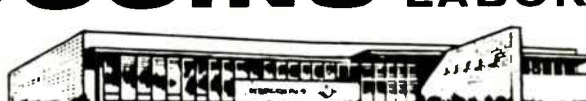

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(Continued from page 106A)

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Miller, G. E., Westbury, L. I., N. Y.
Moldow, B. D., Silver Spring, Md.
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Morrisroe, D. W., Lodi, N. J.
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Parsons, P. L., Broadview, Ill.
Perry, J. R., North Syracuse, N. Y.
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Provost, A. N., Dallas, Tex.
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Reninger, N. W., Columbus, Ohio
Rheinfelder, W. A., Phoenix, Ariz.
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Roller, R. F., Columbus, Ohio
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Smith, W., Freeport, Ill.
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Spacek, C. G., Milwaukee, Wis.
Sweet, W. L., Mountain View, Calif.
Szot, H. J., Montreal, Que., Canada
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Temes, R. J., Paguannock, N. J.
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Teoste, R., Cincinnati, Ohio
Turner, J. W., Baltimore, Md.
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Vonbun, F. O., Hyattsville, Md.
Wallene, G., Chicago, Ill.
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Welch, J. P., Glendora, Calif.
Wezner, F. S., Haddonfield, N. J.
Widmer, A. N., Poughkeepsie, N. Y.
Wyels, P. J., Long Beach, Calif.
Yao, F. C., Pensauken, N. J.
Yengst, W. C., Palso Verdes Estates, Calif.
Zechmeister, J. F., Sunnyvale, Calif.
Zorio, L. F., Dedham, Mass.

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Avery, L. R., Fairborn, Ohio
Bachman, R. G., Jr., New York, N. Y.
Baker, H. M., Glen Burnie, Md.
Barrett, W. S., Paramount, Calif.
Becker, H. J., Palo Alto, Calif.
Bishop, J. E., Wellfleet, Mass.
Boren, J. T., Dallas, Tex.
Bragassa, E. B., Honolulu, Hawaii
Bright, J. R., Mt. Clemens, Mich.
Brooks, H. A., Jackson, Mich.
Byrne, B. H. L., Pasadena, Tex.
Carter, I. P. V., Yorktown Heights, N. Y.
Carter, J. K., Oklahoma City, Okla.
Chapin, G. H., Jr., Duxbury, Mass.
Cohen, A. Z., St. Albans, New York
Congoule, R. L., Redwood City, Calif.
Cowan, J. E., Los Angeles, Calif.
Cronin, P. R., East Williston, L. I., N. Y.
Dailey, L. D., Washington, D. C.
Day, A., Chicago, Ill.
Donsky, Philadelphia, Pa.
Dorrell, R. F., Des Plaines, Ill.
Evans, T., Montreal, Que., Canada
Feely, M. J., Garden City, L. I., N. Y.
Freniere, J. J., Natick, Mass.
Frink, R. D., Lawrence, Mass.
Fujiwara, T., Yuri-gun, Akita-pref., Japan
Gaddis, M. T., Claremont, Calif.
Grimm, J. H., Port Canaveral, Fla.
Harper, J. D., Beaverton, Ore.
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Hokes, F. R., Garfield Hts., Ohio
Horvath, S. M., Philadelphia, Pa.
Jensen, G. N., New Orleans, La.
Joyce, C. F., Dorchester, Mass.
Kalic, D. D., Belgrad, Yugoslavia
Kalnins, M., Cambridge, Mass.
Kibbler, T. H., Wayland, Mass.
Koester, J. C., Santa Monica, Calif.
Kupnowicki, M., Edmonton, Alta., Canada
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Libenson, L. H., Saxonville, Mass.
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Moore, J. V., State College, Pa.
Muscat, E. A., New York, N. Y.
Nutman, A. J., Bayside, L. I., N. Y.

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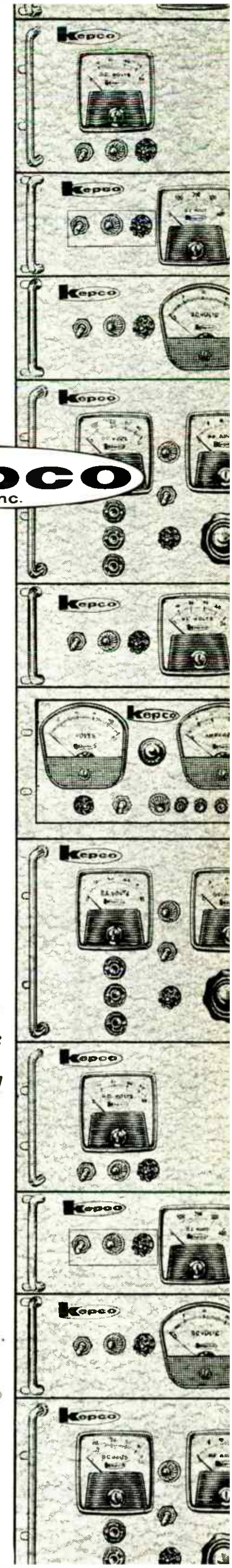

Membership

(Continued from page 109A)

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Palevsky, M., Los Angeles, Calif.
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NEWS New Products

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

(Continued from page 98A)

Pressure Switch

Development of the A-21 pressure switch, has been announced by **Astromics, Div. Mitchell Camera Corp.,** 611 W. Harvard St., Glendale, Calif.

Primary application of the pressure switch is anticipated in the aviation and missile fields, however, it should also be found useful in the varied phases of general industry.

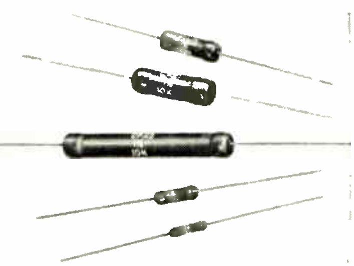
The A-21 pressure switch attains its sensitivity by means of a stretched diaphragm pressurized by dry nitrogen. Earlier approaches to high sensitivity have employed various configurations of solid-state switches. This design utilizes solid Paliney contacts.

Potential applications should encompass aircraft cabin pressurization protection; missile arming; initiating events at pre-set or programmed altitudes, and many industrial control system uses.

Accuracies of plus or minus five feet at 10,000 feet can be obtained in barometric applications.

Solid Resistor Coating

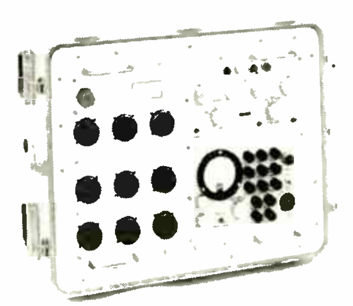
An exclusive "Aeroglaze" conformal coating applied as a 100% solid is claimed to give superior performance and greater mechanical protection over conventional painted carbon deposited resistors by the **Hi-Q Div. Aerovox Corp.,** Olean, N. Y. This new feature is offered on Hi-Q Carbofilm resistors at no increase in price.



Applying the coating as a 100% solid eliminates weakened coatings through use of solvents and provides uniform sizes and improved appearance. "Aeroglaze" coating will meet all the requirements of MIL-R-10509B. Mechanical protection is said to be superior to any other method and special tubes, sleeves and jackets are eliminated. "Aeroglaze" resistors are available in $\frac{1}{8}$, $\frac{1}{4}$, $\frac{1}{2}$, 1 and 2 watt sizes to meet MIL designations RN10, RN20, RN25 and RN30. All ratings from 5 ohms to 50 megohms are available for immediate delivery in production quantitites.

For complete information write to the firm.

Cable Harness Analyzer



The Model 196 Militarized Cable Harness Analyzer allows simple and complex branch circuits to be high-potted and measured for leakage to all other circuits. The manufacturer, **California Technical Industries Div. Textron, Inc.,** 1421 Old Country Rd., Belmont, Calif., states that simultaneously with the leakage test, continuity (conductor resistance) is measured. In automatic operation the wires under test are checked at a maximum rate

(Continued on page 112A)

Professional Group on Aeronautical and Navigational Electronics

Although radio communications early became an essential factor in aircraft development, perhaps one of the most dramatic and specialized fields of present-day radio-electronic progress has been airborne electronics.

Partly due to the war-time secrecy necessary, the Professional Group on Airborne Electronics was the eleventh formed by IRE members. But, it is a particularly clean-cut example of the need for Professional Groups, and the value the groups can render to members. The problems presented called for a high degree of specialization—yet they were essentially radio techniques—from servomechanisms to radar, and from communications to computers and guided missiles.

The Dayton IRE Section introduced its own specialized conference on Airborne Electronics in 1946, which has now become the National Conference of the Group. This Conference publishes a very complete and printed "Proceedings" annually, in book form. The Group also publishes four Transactions a year in cooperation with the IRE Editorial Department, and it sponsors full technical sessions at other Conferences, and notably at the IRE National Convention and WESCON, the papers from which are available to Group members as a part of the Convention Record at a reduced rate.

The group has grown rapidly, and now has members in all parts of the country. It has local Chapters operating in eight cities in the United States.

One very favorable outcome of the development of this Professional Group is that it is tending to accelerate the coordination between air frame design and electronic design in this field—to the obvious benefit of both. It has also drawn together into one Group, through practical engineering necessity, engineers working in sometimes widely separated fields: propagation theory, navigational aids, servomechanisms, etc.

Ernst Weber

Chairman, Professional Groups Committee

NEWS New Products

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

(Continued from page 110A)

of five wires per second. When a fault occurs the tester stops and indicates the circuit and type of fault.

The tester will check 150 simple circuits, 75 main circuits with any combination or number of branch circuits to a total of 75 branches, or any intermediate combination of main and branches up to a total of 150.

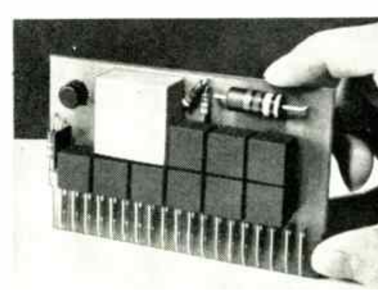
Measurements are made on precision bridges which combine accuracy with fail-safe operation. Continuity testing can be made from 0.3 to 10.0 ohms at 0.5 to 3 amperes. High-pot testing can be made up to 1,000 volts dc. For high-potting, a continuously variable control gives dwell times from 0.2 to 10 seconds. A built-in ohmmeter for manual checking measures leakage resistance from 0 to 1,000 megohms.

Designed to meet the requirements of MIL-T-945A the equipment has been rigorously type-tested and in keeping with its intended function is housed in a moisture and airtight instrument case.

Among the accessory units are: an aluminum transit case and extension switching units which can be used to increase the basic capacity of this unit.

Magnetic Shift Registers

Magnetics Research Co., 255 Grove St., White Plains, N. Y., announces the availability of a group of unusually flexible magnetic shift registers with these features: The ability to achieve non-destructive readout, reversible data-flow, reversible counting, code generation, and "corner-turning" using one core per bit. The use of "wide-width" bit-coupling, which permits controllable-width output pulses, operation essentially independent of shift-pulse width, and minimum shift-pulse power, both peak and average.

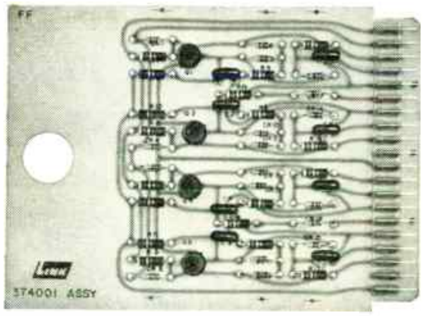


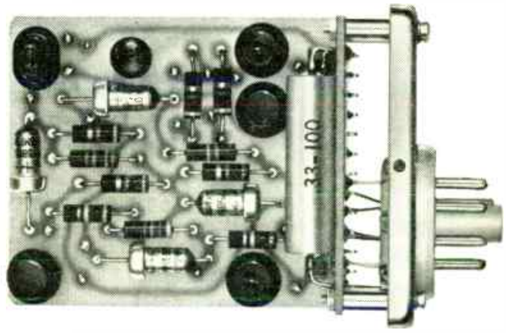
These registers are available in 5-bit, 9-bit, and 10-bit length/module, and may be combined without limit, since each module contains its own shift drive, wide-width gate, and hold-gate circuitry. A wide variety of pulse amplitude, power supply, and operating frequency ratings is available in this series.

Illustrated is the Model 9TDWW-150/U, mounted on a standard ELCO

(Continued on page 116A)

FOR SYSTEM DESIGN





DIGITAL TECHNIQUES

ANALOG TECHNIQUES

Combined by Link to give you Dialog* Systems Utilizing Available Components and Building Blocks

Link's proved analog and digital capabilities are available to help you with your system design requirements. By offering a wide variety of analog and digital modules, Link provides the building blocks that facilitate system design and reduce costs.

Listed below are Dialog system components for control, computation and simulation.

DIGITAL CIRCUIT MODULES

Logic Circuits

- Counter Flip-Flop
- Power Latch
- Power Inverter
- Multiple Input "AND" Circuits
- Multiple Input "OR" Circuits
- Delay Line

Pulse Generation and Amplification Circuits

- Power Driver
- Indicator Driver
- Clocking Pulse Generator

Data Handling Circuits

- Write Amplifier
- Read Amplifier
- Read Head Selector
- Read Head Diode Assembly
- Core Register Assembly
- Core Driver
- Decoding Matrix

Housing

- Card Cage With Locking Bar

ANALOG BUILDING BLOCKS AND SERVO COMPONENTS

Servo Systems

- Rectilinear Servo Motor With Drive Amplifier
- Standard Servo Assemblies, AC or DC

Operational Amplifiers

- Transistorized Standard Operational Amplifier
- Transistorized Low Drift Operational Amplifier
- Miniaturized Electronic Tube Operational Amplifier

Buffer Amplifiers

- Transistorized Standard Buffer Amplifier
- Transistorized 25 Volt Operational Amplifier

Computation Amplifiers

- Transistorized Resolver Driver Amplifier
- Summing Amplifier, Model 301
- Linear Phase Detector
- Phase Detector, Model 303

Servo Amplifiers

- Transistorized 400 cps Servo Amplifier—20 Watt
- Universal Electronic Tube Servo Amplifier, Model 300

Electronic Multipliers

- Time Division Multiplier
- Sine-Cosine Generator

Write to Dept. PI, Industrial Sales Department
for data and specifications on DIALOG components and building blocks

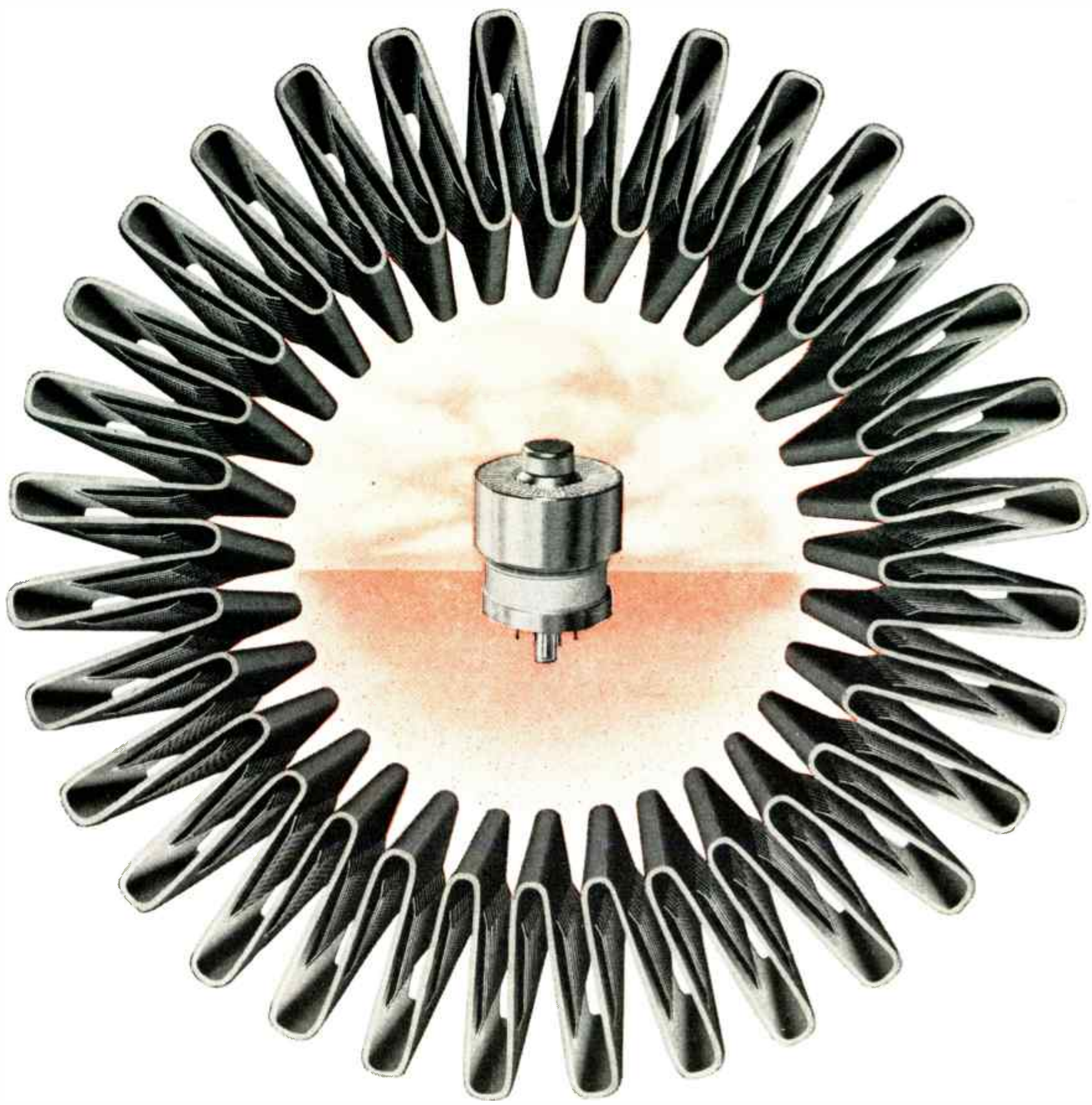
*DIALOG *(Link Digital-Analog Systems, Components and Building Blocks)*

See Dialog components at the WESCON Show August 23-26—Booths 641 and 642

Another example of Link / Ability

LINK DIVISION
Binghamton, New York

GENERAL PRECISION, INC.
Other Divisions: GPL, Kearfott, Librascope,



RCA radiator efficiency allows you to cut blower horsepower in half

Resulting space and weight savings permit more compact packaging, greater design versatility

Recently a leading manufacturer of communications equipment ran a comprehensive series of tests comparing the heat dissipating capabilities of RCA-7203/4CX-250B, with integral radiator, to a standard JAN/4X-250B. The results: the JAN/4X-250B required twice as much blower horsepower.

The integral radiator on RCA-7203 is an exclusive RCA development. It is used on many types of RCA tubes for air-cooled operation. Its remarkably efficient louvered construction is a logical development in radiator design . . . but only RCA has mastered the complex manufacturing problems involved.

This high-efficiency integral radiator is one of the things that makes RCA power tubes so popular in designs where space and weight are critical considerations. For further information about this and other pace-setting features of RCA power tubes, including the new Cermolox* line, write:

The Marketing Manager, RCA Electron Tube Division, Industrial Tube Products Department, Lancaster, Pa.

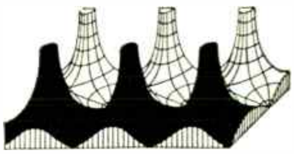
*RCA's line of coaxial, precision-aligned grid, beam power tubes of ceramic and metal construction. See RCA's forthcoming advertisement on this complete line.

RCA ELECTRON TUBE DIVISION— FIELD OFFICES . . . Government Sales: Harrison, N. J., 415 S. 5th Street, HUmboldt 5-3900 • Dayton 2, Ohio, 224 N. Wilkinson St., BAldwin 6-2366 • Washington 7, D.C., 1725 "K" St., N.W., FEderal 7-8500 • Industrial Products Sales: Detroit 2, Michigan, 714 New Center Building, TRinity 5-5600 • Newark 2, N. J., 744 Broad St., HUmboldt 5-3900 • Chicago 54, Illinois, Suite 1154, Merchandise Mart Plaza, WHitehall 4-2900 • Los Angeles 22, Calif., 6355 E. Washington Blvd., RAymond 3-8361 • Burlingame, Calif., 1838 El Camino Real, OXford 7-1620

The Most Trusted Name in Electronics
RADIO CORPORATION OF AMERICA

August, 1960
Vol. 48 No. 8
Proceedings of the IRE

Poles and Zeros



SAC. At the invitation of the Strategic Air Command, your Editor and Managing Editor visited SAC Headquarters, in Omaha, Nebraska, to attend a communications briefing. The visit was made by a group of communications and electronic editors of national magazines and periodicals. We were afforded the opportunity to see and examine the extensive communications facilities established by SAC as a vital aspect of their task of war prevention.

It is gratifying and exciting to learn of the outstanding communications development of the Strategic Air Command. The ability of SAC to achieve rapid communications between its Headquarters, its air bases, and its aircraft was vividly demonstrated. Perhaps most impressive was an actual contact using single-sideband voice communication. Headquarters in Omaha called to and received acknowledgment from Guam, Tokyo, Okinawa, Fairbanks, Hawaii, Thule, Goose Bay, Bermuda, Puerto Rico, Tripoli, England, Turkey, Madrid, Dhahran, and Westover, March, and Barksdale bases in the United States; these seventeen contacts required less than two minutes of elapsed time. Unquestionably the ability to to achieve this performance is in part due to the wisdom of the Air Force in choosing single sideband as the basic communication technique. One wonders whether the Special Issue of the PROCEEDINGS in December, 1956, might have played a role in this instance.

The much publicized "red telephone," the SAC hot-line for alerting SAC's entire force, was also demonstrated. In this case, over seventy locations, both in the United States and overseas, were contacted simultaneously and acknowledgment was received in less than ten seconds. This was a remarkable demonstration of the communication engineer's dream come true.

From the point of view of engineering, in addition to that of the radio and communication engineer, the command post itself is a fascinating achievement. This structure, of three floors, extends 45 feet below ground level. It can be sealed off and become self sustaining. It is provided with its own air conditioning system, an independent water supply, an emergency electrical power system, complete communications as described above but including teletype, and a supply of emergency rations.

As usual, the group of editors was occasionally skeptical of some of the claims made. They asked many questions about the facility for combating the many possibilities of failure that are inherent in any complex system. SAC was always ready with an answer and the evidence of careful detailed planning was clear. Even the back-ups have back-ups almost ad infinitum.

While arguments will continue as to the relative role of manned aircraft and missiles in the country's armament posture, this Editor returned from the briefing feeling proud of the demonstrated significance of the contribution of communications engineering to the defense plans of the United States.

Hands Across the Border. Not only hands, but also whole bodies traveled across the border (northern) last April. An automobile caravan of 65 IRE members from the Ottawa Section, led by Past President John Henderson, paid a weekend visit to Syracuse, and were guests at a joint dinner meeting of the Syracuse and Ottawa IRE Sections. Wives were included in the visit, and shopped in the Syracuse stores while their husbands participated in a plant tour. The dinner meeting featured a talk by James E. Keister, "How Much Space." The Ottawa to Syracuse visit was so successful that plans are being made for a Syracuse to Ottawa visit next spring. This interesting aspect of IRE "internationalism" should be emulated elsewhere; joint meetings of adjacent sections, even within the United States, should be stimulating.

Abstracts and References. The monthly feature of the PROCEEDINGS, Abstracts and References, was emphasized in June with the publication of the yearly index, as Part II of the June issue. The significance of this contribution is reenforced by a study of the magnitude of the task performed by the compiler. The Radio Research Organization of the Department of Scientific and Industrial Research, England, scans 190 publications from 20 countries in assembling the Abstracts and References which IRE publishes by arrangement with the Department and with *Electronic Technology*.

The coverage offered by these abstracts is not always appreciated. One's appreciation may be enhanced by the following tabulation of the countries and the number of journals covered in each country. The list is presented in alphabetical order.

Argentina	1	England	49	Netherlands	7
Australia	3	France	15	South Africa	1
Austria	2	Germany	18	Spain	2
Belgium	2	India	3	Sweden	6
Canada	2	Italy	6	Switzerland	6
East Germany	5	Japan	7	Turkey	1
USA	40			USSR	14

You curious readers may now refer to the annual index to assuage your curiosity about the titles of the journals represented by the tabulation.

Section Number 107. The Executive Committee, in May, 1960, approved the change in status of the Kitchener-Waterloo Subsection to full Section status. The new Section also becomes the fourteenth Section in Region 8. We extend a hearty welcome!—F. H., Jr.

B. J. Dasher

Director, 1960–1961

Benjamin J. Dasher (A'41–M'55–SM'57–F'59) was born in Macon, Ga., on December 27, 1912. He received the B.S. in E.E. degree in 1935 and the M.S. in E.E. in 1945, both from Georgia Institute of Technology, Atlanta, Ga. He received the Sc.D. in E.E. from Massachusetts Institute of Technology, Cambridge, Mass., in 1952.

As an undergraduate student, he enrolled in the co-operative program, and during the last two years spent his work periods as an operator for radio station WMAZ in Macon, Ga. After graduation he spent four years as a partner in a radio and electrical service business in Macon and then returned to Georgia Tech as a graduate assistant in Mechanical Engineering. In 1940, he was appointed Instructor in Electrical Engineering and remained at Georgia Tech during World War II.

In 1946 he went to Massachusetts Institute of Technology as an Instructor in Electrical Engineering where he was engaged in teaching, research, and graduate study. He was associated with the Telemetering and Instrumentation Group of Project Meteor, Research Laboratory of Electronics, M.I.T. from 1950–1951.

In the fall of 1951 he returned again to Georgia Tech and was appointed Director of the School of Electrical Engineering in 1954.

Dr. Dasher is the author or co-author of a number of technical papers, most of them in the field of circuit theory dealing with the synthesis of RC networks. He is co-author of the Chapter on alternating-current measurements in "Radio Engineering Handbook," edited by Keith Henney, and has contributed articles on electrical engineering to World Book Encyclopedia.

He served as Student Branch coordinator and member of the Executive Committee of the Atlanta Section, IRE, for several years and he has been an editorial reviewer for the PROCEEDINGS since 1954. He has served as General Chairman of the IRE Instrumentation Conference, sponsored by the Atlanta Section and the Professional Group on Instrumentation, and also as a member of the Administrative Committee of PGI.

Dr. Dasher is a member of the AIEE, the American Society for Engineering Education, Eta Kappa Nu, Sigma Xi, and Phi Kappa Phi.

Scanning the Issue

Can the Social Sciences Be Made Exact (Berkner, p. 1376)—Last March, at the Opening Meeting of the IRE International Convention, a Director of the IRE called attention to some interesting similarities between the behavior of circuits and of humans. These comparisons were presented with the thought that they might provide social scientists with some much needed help in developing a more precise and quantitative understanding of the behavior of the human mind. Such an understanding would benefit the electronics engineer, too, for he must match many of his communication systems to human terminals about which he presently has inadequate knowledge. It is not the usual practice of the PROCEEDINGS to print speeches, but then this is not a usual speech.

A Review of Panel Type Display Devices (Josephs, p. 1380)—The science of communications and electronics has traditionally been concerned with methods of generating, transmitting, and receiving information. With the advent of television, radar, and the computer there has been added a need for methods of displaying information. During the last five years this need has led to the development of a surprising variety of panel display devices, not only of the solid type but evacuated, gas, liquid and mechanical devices as well. This paper presents a timely review of the substantial activity in this field.

An Improved Film Cryotron and Its Application to Digital Computers (Newhouse, *et al.*, p. 1395)—Several laboratories have been working intensively on methods of depositing thin films of superconducting and insulating materials, less than a micron in thickness, to form extremely compact cryotron circuits for computers. As a result, film cryotrons have in recent months become one of the most important and rapidly moving developments in the computer field. Film cryotrons constitute the first advance over transistors with respect to computer miniaturization. Much larger numbers of cryotrons can be used in a small volume without encountering the heat dissipation problems of transistors. Packing densities of a million elements per cubic foot are now a distinct probability. Moreover, they are easy to assemble; thousands of film cryotrons and all their interconnecting circuitry can be deposited at one time in a few simple steps. This is one of the first few papers on superconductive film devices (one appeared here last month) and is believed to be the first to describe the crossed-film cryotron.

Gallium Arsenide Tunnel Diodes (Holonyak and Lesk, p. 1405)—Most of the work on tunnel diodes that has been reported to date has involved diodes fabricated from germanium and silicon. This paper concerns the fabrication and properties of tunnel diodes made from gallium arsenide, an intermetallic compound which offers many attractive properties. Those properties give gallium arsenide tunnel diodes a number of advantages over silicon and germanium varieties, especially in the larger number of applications for which they are well suited. This paper will, therefore, be of great current interest to practically everyone in the solid-state field.

The Reflected-Beam Kinescope (Law and Ramberg, p. 1409)—The authors have made some radical changes to a conventional picture tube, resulting in a substantially shorter device, less than a foot long, especially suited for picture sizes of 21 inches or more. The phosphor viewing screen has been moved back a few inches from the tube face, while the gun assembly has been moved forward, telescoping the neck of the tube. The electron beam goes through the phosphor screen, which is perforated. As it approaches the tube face, the beam is reflected by a transparent conductive coating and falls back onto the phosphor screen to form the picture. This unusual construction offers a novel approach to thin cathode-ray-tube design.

Shot Noise in Tunnel Diode Amplifiers (Tiemann, p. 1418)—This paper develops some basic theory and calculations on the noise properties of tunnel diodes, providing a valuable reference for future work in this area. The widespread interest in this subject, both on the part of device physicists and designers and on the part of circuit designers who may want to employ tunnel diodes, will make this a widely read paper.

A Parametric Device as a Nonreciprocal Element (Kamal, p. 1424)—The author proposes connecting an up-converter parametric amplifier to a down-converter parametric amplifier to obtain a nonreciprocal phase shifting device. He demonstrates that the phase of a signal can be shifted $+90°$ when it is converted up and then down and, conversely, $-90°$ when it is converted down and then up, thus giving a two-way phase shift of 180°. This proposal is of considerable interest because it offers a way of providing isolators and circulators at frequencies below the range where ferrite devices can operate.

The Compatibility Problems in Single-Sideband Transmission (Powers, p. 1431)—An analysis of single-sideband signals has been made which yields results of fundamental importance in determining the conditions under which SSB transmissions can be made compatible with standard AM receivers. The first important result of this analysis is that conventional SSB cannot be made fully compatible without a sacrifice in spectral economy. The second important result is that a proposed square-law SSB system, in which the message is conveyed in the square of the envelope of the wave rather than in the envelope itself, requires no sacrifice in spectral economy. In this case, a square-law envelope detector would be required in the receiver for distortionless detection of the signal. However, a conventional AM receiver could receive speech transmission without serious loss of intelligibility, thus preserving a measure of compatibility. Readers will also be interested in the comments which appear on page 1503.

On the Resolving Time and Flipping Time of Magnetoresistive Flip-Flops (Aharoni and Frei, p. 1436)—Magnetoresistance is a well-known phenomenon which has been used for studying the properties of elements and alloys but which has not yet been applied to circuits or devices. This paper discusses an application, probably the first in the electronics field, which may be potentially important in the switching and data processing fields. Although the results are inferior to existing art, the theory presented shows no inherent limitations. It is felt that publication of this paper may stimulate others to find improved magnetoresistive materials.

IRE Standards on Nuclear Techniques: Definitions for the Scintillation Counter Field (p. 1449)—Definitions are presented for terms related to scintillation counters, ranging from Accelerating Electrode to Light Pipe to Wavelength Shifter, and encompassing 83 other terms in between.

Noise in Oscillators (Edson, p. 1454)

Background Noise in Nonlinear Oscillators (Mullen, p. 1467)

Monochromaticity and Noise in a Regenerative Electrical Oscillator (Golay, p. 1473)—It will be noted that the above three papers are addressed to the identical problem. The viewpoints, however, differ substantially. Indeed, the authors arrive at results which are not equivalent in every respect. The problem they have tackled is an extremely difficult one because it involves nonlinear circuits, and because of this has received relatively little attention. The problem is also an important one, for noise affects the starting characteristics of oscillators, alters the amplitude and phase of the oscillations, and contaminates the spectral purity. It is felt that publication of three different approaches will do much to help solve a complex and basic problem.

Scanning the Transactions appears on page 1509.

Can the Social Sciences Be Made Exact?*

L. V. BERKNER†, FELLOW, IRE

Summary—The social sciences are in great need of a more precise understanding of how an idea is conveyed from one mind to another. The difficulty seems to stem from an inability to find the fundamental and independent parameters from which the behavior of the brain and the consequent human response can be constructed in a mathematical sense.

This problem concerns the electronics scientist as well. He must provide the communication links for conveying information from individual to individual, but he has inadequate knowledge of the human terminals to which he must match his systems.

It is suggested that certain strong similarities between the behavior of circuits and of man may provide analogs which would reveal characteristics of the mind that are fundamental and quantitative. A number of such analogs are proposed and discussed.

THE SCIENCE of communications which underlies the whole thinking of the IRE is ultimately concerned with the transference of information from individual to individual. Our whole science of radio has grown up because man has felt it necessary to convey communications without regard to distance between communicants in the most rapid and effective way. The very primitive means of communications such as smoke signals or tom-toms, or even the eventual telegraph were not fast enough, so man now carries the voice, and indeed the picture of the whole surroundings of that voice, to make communications more complete. Gradually the distance of effective communication has lengthened until reliable voice communication to any part of the earth has become feasible. In the past few weeks coded messages have been received from millions of miles in space to extend our range of transmission to a substantial part of the planetary system. Where sufficient power is not available to transmit a very sophisticated signal, we are prepared to slow down the rate of transmission and then to resynthesize to bring it back to normal picture speed, thus retaining the level of detail that we choose.

Out of this business of conveying signals of one character or another from one place to another in an intelligible and useful way has grown the whole field of electronics. Over the past three decades, starting with the original formulations of Hartley and Heising, we have learned how to measure precisely the amount of communication that can be conveyed between two points in terms of the strength of transmitted signal, the intervening loss, the noise level at the receiving source, and the bandwidth required to contain the whole transmission. Using Shannon's modern formulation of information theory, this whole matter of conveying information in an acceptable form from one point to another has become a well-defined science, and the whole performance of any system can be predicted almost unambiguously. With the elaboration of Wiener's cybernetics, and the intricacies of stochastic theory we can dissect the signal in almost any way we choose, transmit it using very advanced systems of modulation, and then resynthesize it in forms that, employing the advantages of redundancy, can sometimes be better than the original. Thus we can dissect a photograph into appropriate bits of information and resynthesize this photograph at a distance out of those same bits to reconstruct the photograph with a definable measure of deterioration. But, even more important, we can surpass the human capability for assimilating such information very considerably by scanning a whole series of rather poor photographs of the same thing and reproducing a single photograph out of all of this information which is very considerably and definably better than any of the foggy originals simply by statistically suppressing the random noise in the final reproduction. Of course, there are a variety of phenomena affecting the transmission and reception about which we need to know a great deal more, and so we have electronics and computer and transmission research. But as a whole the last two decades have brought the whole problem of the conveyance of information from point to point into a manageable form, well defined in terms of understandable and quantitatively definable parameters.

Information, however, when communicated from one point to another, even from mouth to ear, is altogether useless unless it can be formulated by the sender in understandable form and unless it is received and assimilated in a way that produces something similar to the image which the sender intended to convey. Now this question of how to formulate a message in the brain of the sender to contain the minimum of information necessary to convey an image with definable precision and without employment of unnecessary redundancy, and the corresponding problem of the extent to which the message will create an adequate image in the mind of the receiver of the message, now lies pretty much in the realm of the social sciences. Indeed, psychology has grown up over the past century to deal with the problem of stimuli (the messages) and responses (which are the recipient's reactions to the images produced by the message).

At the turn of this century Ivan Petrovitch Pavlov published his original researches on cerebral activity and the theory of reflexes. His experiments on "condi-

* Received by the IRE, March 25, 1960. Presented at the Opening Meeting, 1960 IRE International Convention, New York, N. Y.; March 21, 1960.

† President, Associated Universities, Inc., New York, N. Y.

tioned reflexes" set the tone for modern behaviorist psychologists. Since that time psychology, using Bridgman's concepts of experimental measurement to provide precise and repeatable measurements of response to defined stimuli, has come a long way in increasing our confidence that *under defined conditions of learning and stimuli life will behave within a statistically predictable pattern.* As the psychologist says it, given a motivation, the individual will respond after a given and definable response time in accordance with a reaction pattern determined by the preparation or training of the individual. For instance, the strength of hunger of a rat after a known period of food deprivation can be measured objectively by the strength of a shock from a grill lying on the floor of a passageway that will deter the rat from crossing the grill in order to obtain food. I remember that in the Antarctic I was impressed with the similar reactions on the part of humans who had been out on the ice for a long time and were willing to undergo certain measures of exposure or hazard simply with the expectation of getting a drink. More recently, rather complex psychological experiments have been conducted among groups of individuals to determine their statistical behavior in problem-solving when placed in any one of a variety of positions in the group with respect to the availability of information and the redundance and order of that information. Alex Bavelas has shown that the same group will exhibit very different capabilities for dealing effectively with a given problem depending upon their lines of communications and their accessibility to the judgments and reactions of others. Interestingly enough, an excess of communication interconnection may actually reduce the problem-solving capability of the group.

But such experiments in psychology are a far cry from providing quantitative understanding of how an idea is conveyed from one mind to another. Just what is the information content of an idea? How fast can this information be assimilated in terms of parameters of the senses and of the brain? What pre-existing memory units and preformed programs (to borrow the words from computer lingo) must a brain contain to generate the image that the idea contains? How can we measure the precision of transfer in terms of the quality of the image and of the ultimate response reproduced? While psychological experiments of today produce numbers, these numbers in themselves are not directly applicable to the basic problem of the formulation and reception of ideas of given complexity in an efficient and defined way. The problem in social science is to find elementary, fundamental, and independent concepts or parameters, whose coefficients can be determined numerically, and which combined in suitable mathematical formulations could predict analytically something about the ultimate capacities of the individual.

The basic problem of teaching and learning perhaps best points up the difficulties faced in social science. Among all teachers, I am sure we would agree that some are known as good teachers (and perhaps we should not admit that there are any bad teachers). Nevertheless, there can be no doubt that some men can formulate and convey information more effectively than others. But indeed no one would contend that there is anything like an objective measure of the capability of a teacher. On the other hand, we can consider the student—his ability to learn, his rate of learning, the minimum requirements for the content of the information to permit him to learn in the light of his past experience, and all of the other questions that are involved in receiving information and reproducing and retaining the desired images with definable precision. This too is a purely subjective matter with which social scientists have not gotten very far. Indeed, one need only read the statement from the discussion of the Trustees of the Carnegie Foundation for Advancement of Teaching to realize how subjective these matters are at the present time. In discussing the selection of academic talent for special attention in teaching, the report of this discussion says:

> . . . appraisals of academic talent must be based on many kinds of evidence. Scholastic aptitude tests provide one significant kind of evidence. Achievement tests provide another. School grades offer a third kind of evidence. The judgments of teachers expressed in rating scales or in written comments are important. The views of counselors, deans, and others who have dealt with the student may be significant. There is no faultless measuring instrument for identifying the academically talented. Each type of measurement has its advantages and its limitations. The important thing is to make use of all available measures as intelligently as possible.

Unfortunately, the report does not define how this information is to be employed to meet the criteria of *intelligent* use. There is no formula by which the information about the student can indeed define his learning ability. I would defy anyone to assess in even a slightly objective way how the appraisal of a student's learning ability can be made unless your knowledge of him is so redundant as to require an unacceptable level of experience with his capabilities.

One wonders, in fact, whether the really *talented* student can be recognized by the *average* teacher, since that teacher, however well trained, cannot by definition extend his vision to the image capabilities of the *talented* student. In such a case, "intelligent" use of information in judging the student is meaningless.

I recall that one imaginative university professor began to wonder why his students were all so very average and untalented. He studied the State Board examinations used to select students—examinations that were of the true-false and check box variety. They contained such questions as:

"Vegetables are destroyed by:
moths
worms
butterflies
snails
slugs."

The student was to check one of these. Now if the student were lazy and unimaginative, he would have studied only one general science textbook—the one prescribed by the State Board of Education—and unhesitatingly would check snails, thereby receiving full credit. But if the student were imaginative, talented and creative, he would have read several books on natural science and biology, and having learned that any one of these pests could destroy vegetables, he had only one chance in five of getting credit. After reviewing the whole examination structure, the professor concluded that the State Board examinations were drawn hastily from specified and incomplete texts by persons of no better than average talent and learning. Therefore, the examinations were selecting persons just as stupid as the persons preparing the examinations and, worse luck, rejecting the brighter and more promising ones.

All of us have had the experience in IQ testing of finding true-false questions that were completely ambiguous in light of our special training. In the face of such questions, the really bright person will be smart enough to ask himself, "How would I answer this if I were very stupid?," thereby recognizing the correct answer without delay, and justifying his high IQ. But I doubt that the examiners planned it that way! I think it is evident that if average teachers are to identify talented students, they need more objective measures than those provided by their own limited estimate and appraisal of the student.

Certainly, we in electronics and communications are vitally concerned with the analytical process of conveyance of an idea from one brain to another with a minimum definable level of information. It is perhaps the most fundamental problem of the social sciences. For it involves the matching of the means of communication from the originator to the receptor. One cannot help speculating whether at this stage in the development of our science of information theory, its formulation cannot be adapted to apply further to analysis of the minds of the individual (and then perhaps to the defined statistical behavior of a group). We, as communicators, need some more reliable and analytical way of understanding how ideas are formulated, conveyed, acquired, retained, and modified.

Since no one else has succeeded in dealing with this problem at the present time, we can be excused if we entertain ourselves with it a little. And so, we ask the question: Is it possible that the fundamental and independent properties of learning and of formulation of ideas might be those same fundamental properties which we find in the electronic circuits that we deal with every day? A lot of people have thought about this in a general way, but not many psychological experiments have been designed to produce coefficients in a form that will fit our every-day circuit parameters.

Suppose we postulate some possible *fundamentally independent* functions in the human brain, reasoning from circuit analogs. Of course, this process is not new—many workers have studied the similarities between the computer and the brain. But perhaps it is worth trying again.

Many people have studied the obvious analog between memory in the brain and in a variety of electronic circuits. Memory seems to be packaged individually in a very large number of independent "circuit" units, each of which has a content determined, at least in part, by the time order of acquisition of the information.

Associated with problems of memory, are problems of retention of memory with time. Certainly, this has obvious circuit analogs involving leakage of charge or of physical or chemical deterioration of circuit properties. Possibly a whole memory package may be lost with a slight modification of only one element of the package, and restoration of memory storage may require restoration of only one single element of the package. Certainly the test of memory as a pure matter, without resort to its renewal by association, might be quantitatively tested. Even more important would be determination of what elements of a lost memory package must be renewed to revitalize the whole package without necessity of its entire replacement. Incidentally, one might remark that some memory packages are supplied at birth—packages acquired by heredity or prenatal social experience. These may be substantially different in circuit characteristics from those formed from experience.

Given a supply of memory packages, the mind has perhaps a number of programs which permit association of the memory packages in different ways. Thus, combinations of memory packages can be set up in a program to solve problems. While some programs are obviously available at birth, most are acquired by experience and training. Certainly, programs of thinking are not very useful until perfected by disciplined and repetitive experience. But I would emphasize in this hypothesis the probable complete independence of the memory process from the acquisition and perfection of independent programs—programs that employ that memory in problem-solving, by successive association of memory items, with the objective of optimizing the solution.

Both the number of memory packages accessible, and the number of independent and well-disciplined program procedures available, will provide limits of creative problem-solving. Perhaps, then, we can say that an idea can be specified as a number of relevant memory packages when scanned in proper succession in accordance with a well-defined program. According to this hypothesis, conveyance of an idea involves specification of the necessary memory packages and of the appropriate program for their utilization.

Now in the conveyance of ideas we must be concerned with acceptable data rates and noise levels, as well as with the speed with which the receptors program will function in reproducing the desired image. It seems probable than an individual brain, or at least the sensing device stimulating the brain, has a bandwidth which restricts the rate at which information can be assimilated into the memory. This bandwidth may vary a good deal among individuals, and certainly differs in the same individual from one means of sensing to another. Whether it is the bandwidth of the brain or of the sensing device that limits the information rate in all situations is not clear; probably there are at least three independent "bandwidths" involving sensing, speed of access to memory storage, and functioning of the program itself.

Moreover, the highly trained brain may sometimes function efficiently with a very mininum of input by stimulation of a program that creates an analogous image from pre-existing memory packages. Certainly, in this case, sufficient redundancy in the message is required to check the precision of the analogous image that has been generated. Consequently, perhaps a brain with a narrow bandwidth (that is, a slow learner) may acquire (through long and patient study with carefully tailored rates of training) very efficient, though sometimes inaccurate, means of image formation and problem-solving. Indeed, in the face of high noise levels, the brain with the narrow bandwidth might be superior on some occasions.

Noise certainly appears at the input in the form of irrelevant information or, even worse, misinformation in the desired circuit; or it may involve an overpowering external signal or stimulus from an unwanted channel. Perhaps the brain or sensors can adjust their parameters and narrow down the bandwidth to minimize the noise, but certainly the information rate is correspondingly reduced. In addition, the mind must deal with internal noise in the form of fears and prejudice which distort the image that the message is designed to convey. In severe cases of prejudice or emotional disturbance, or after "brain washing," most or all programs seem permanently connected to cavities or resonators that are excited by almost any input to amplify the same old fixed idea whose noise level over-rides any hope of reproducing a recognizable image. I suppose the "conditioned response" involves the manufacture and installation of such noise amplifiers, some of which can be very useful, when they respond to a coherent signal. But they also tend to fog and distort the beauty of complex images and thus diminish the effectiveness of otherwise efficient programs of creative thought.

In this hypothesis, the speed and effectiveness of response to a message depends on an adequate availability of suitably designed and prestimulated memory units, on rate of information supply with acceptable redundancy, on rate of information assimilation, on both internal and external noise levels, and on availability of a well-functioning and suitable program to organize the material at hand into a proper and useful sequence which forms the image and depicts the decision. Assimilation of information may involve both supply to additional memory units and selection of the most suitable or applicable program. The response is tactical in nature, since speed implies a time limitation. Therefore, effective tactical thinking implies preorganization of a sufficient number of programs that will be needed for the variety of tactical situations that may be encountered. In addition, judgment represents development of a well-designed "super-program" to select the program or program combination most applicable to a given situation.

Now, the problem of generating strategic thinking is quite different. Here we must install additional pre-excited memory units—that is, acquire all applicable information in sufficient detail. But, more important, we must develop new programs to associate the memory packages in new ways, some of which will permit generation of new, creative, and useful "ideas." This takes time, study, and mental discipline. It requires familiarity with libraries containing information too involved for simple storage in memory units, and complex and rarely used programs for their association. Consequently, I am led to suspect that the length of the university semester is no accident. I have noticed in the several "study" projects in which I have participated that new creative ideas take several months to incubate and the process leading to them involves a discipline of study and thinking that is well-defined. Moreover, the presence of several differently trained brains, with resultant group access to a wider variety of programs, produces a composite image that more clearly defines a new idea. The high noise of conditioned response of each individual, which is random among a group with diverse training, can be sufficiently suppressed in the composite to provide a greater insight previously obscured in any one of them by his own internal noise level. Therefore, strategic thinking may be more effective in group study because of enforced thinking through group discipline and because of suppression of noise in generation of clear insight.

In conclusion, some interesting and amusing interpretations can be drawn from our elementary hypothesis. Man lives primarily as a tactician with a limited supply of memory units and a few well-used programs to associate those units in a stereotyped and unimaginative pattern. Perhaps this represents the so-called "one-track" mind. A few who develop strategic capabilities by installing additional memory units and devising new programs to employ them are known as scholars, but more often as "egg-heads." Conventional courses of education and examinations tend to emphasize tactical ability and tell little about bandwidth, internal noise level, unusual memory capacity or storage, discipline in learning, or consequent capability for devising novel programs for creative thinking. Conventional education may actually tend to suppress the progress of those with creative capability.

Now you may very well say that all of this is hypothetical and unproved by experiment and therefore simply a flight of fancy. I would agree. But the image I am trying to convey is this. The social scientist is in trouble. He cannot define objectively man's capability to employ information. We as communicators are in trouble with him, because as conveyors of that information, we have no objective or definable knowledge of the "source" or the "sink" to which we can match the terminals of our intermediate links. The trouble seems to stem from an inability to find the fundamental and independent parameters from which the behavior of the brain and the consequent human response can be constructed in a mathematical sense. We *have* developed very precise means of understanding information capacity in terms of circuit theory. There appear to be strong similarities between the behavior of circuits, and of man, with respect to the independent parameters leading to images and decisions of defined reliability and quality. Perhaps, through experiments and analysis derived from these analogs, we can ascertain the fundamental quantities and their coefficients from which the properties of man's brain and his consequent mental capabilities can be synthesized and mathematically defined.

A Review of Panel-Type Display Devices*

JESS J. JOSEPHS†

***Summary*—A review of the activities of various governmental, industrial, and academic institutions is made in the field of panel display devices. These devices are thin sheets which can display luminescent information. Insofar as possible the present state of the art is indicated. The displays are grouped as: evacuated, solid, gas, liquid and mechanical.**

Among the evacuated types are listed the thin cathode ray tubes and the image intensifier.

The solid-type displays employ the electroluminescent panel in conjunction with an *xy* matrix of electrodes. Recent progress in panels of the above type with memory are included as well as a brief discussion of solid-state image intensifiers.

Descriptions of proposed gaseous displays as well as of liquid and mechanical displays are included.

A brief discussion of switching circuits is included.

Actual as well as possible applications of these devices are listed.

I. Introduction

CONSIDERABLE interest has been shown of late in the class of devices known as panel displays. Industry has always been interested in the ultimate "color-picture-on-the-wall"-type television, while the military has been looking for large flat display boards for combat information centers, air-traffic control boards, radar plotting displays, and displays used for readout of computer information. In many of these applications, a fast writing, bright display with holding or storage is sought. In addition, a selective erase feature would be desirable for some applications.

The panel-type display is a flat device (flat in the sense that the depth is considerably less than any area dimension) that acts as a transducer to convert an optical or electrical input into an optical output primarily adapted for human observation. This eliminates from consideration those devices which may act as inputs for electronic or other type systems.

Some of the earliest displays have been found in advertising and include lighted billboards with animated displays. One of the best known of these is the *New York Times* news bulletin display [1], which can be viewed from the Times Square area in New York and will be described briefly below.

Many of the more recent displays utilize electroluminescent (EL) panels modified for added versatility.

II. Types of Panel Displays

Advances in solid-state physics and improvements in the preparatory arts of other fields have given added impetus to research and development of these newer displays.

The approaches have become so diversified that one may find displays using liquids, solids, and gases as the display materials.

The sizes of these displays may vary from 1 inch × 1 inch to 5 feet × 5 feet. Larger sizes are also being planned.

In the descriptions that follow, displays will be grouped into: evacuated, solid, liquid, gas, mechanical and miscellaneous.

* Received by the IRE, August 15, 1958; revised manuscript received, January 18, 1960. The work reported in this paper was performed by Lincoln Lab., Massachusetts Institute of Technology, Lexington, Mass., with the joint support of the U. S. Army, Navy, and Air Force.

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A. Evacuated Displays

The evacuated-type panel displays fall into two groups: 1) conventional CRT's redesigned to give the desired geometry, 2) image intensifier displays. In the second class electromagnetic radiation of the desired spectral distribution forms an image on an appropriate photocathode. The photocurrent is increased by striking anode surfaces liberating additional electrons. In various ways these electrons find their way to a phosphor screen where they produce light of the desired color.

1) Evacuated Cathode Ray Tubes of "Flat" Design [2–8]: The geometries of these tubes are fairly similar for the different types. Deflection is usually obtained by electrostatic means, although one design uses magnetic deflection in addition to electrostatic deflection. In a few of the schemes provision for color is included. Fig. 1 shows a pictorial representation of a particular flat tube [2]. Electrons are emitted from a gun in a cylindrical stream along the x axis in the space bounded by a set of horizontal deflection plates in the xz plane, and the lowermost vertical deflection plate in the xy plane (see Fig. 1). Since this space is initially field-free, the electrons travel in straight lines. To deflect the beam upward, the potential of one of the horizontal deflection plates is lowered. In this way the deflection can be made to occur anywhere along the x axis.

The beam, traveling in the y direction, has as a cross section a thin ellipse with its long axis in the z direction. This is due to the electron optics involved.

The upward deflected beam now enters another field-free region bounded on one side by the glass plate with the phosphor screen and on the other side by another glass plate to which are attached the vertical deflection plates lying in the xy plane.

To deflect the beam outward in the positive z direction and onto the phosphor, the potential on one of the vertical deflection plates is lowered. The position at which the deflection occurs can be changed continuously by a corresponding variation in the potentials of the adjacent vertical deflection plates. The beam, which originally had a thin ellipse for a cross section, now has a small circular cross section.

If one chooses an appropriate sequence of potential variations over the horizontal and vertical deflection plates, then the spot at which the beam strikes the phosphor can be made to sweep out a raster.

A particular sample displayed a television picture which could be viewed from either side, and another sample employed a transparent phosphor.

In this and similarly designed tubes a switching problem is presented which is absent in conventional cathode ray tubes. In the latter, horizontal and vertical deflection voltages at the neck of the tube cause the electron beam to sweep out the raster, while, in the former, a succession of decreasing potentials need to be applied to both the horizontal and vertical deflection plates to accomplish the same effect. Details may be obtained in the references cited.

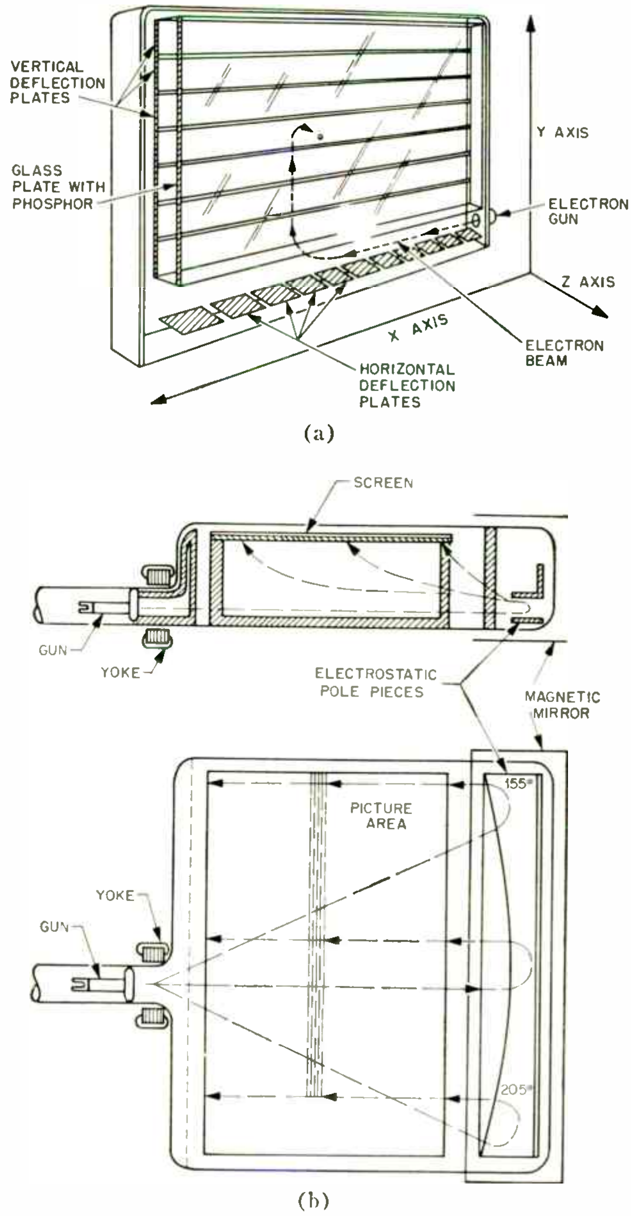


Fig. 1—(a) Pictorial representation of a flat tube. (b) Schematic representation of a magnetic deflection flat tube.

Various switching schemes have been proposed. These include the use of beam switching tubes and delay lines as well as various solid-state and electronic devices. They are discussed in Section III.

In another design [4] of a "flat" tube, electrostatic and magnetic deflection are used in such a way as to obviate the need for switching circuits. See Fig. 1(a).

The tube, as before, is a rectangular parallelepiped of small depth with an electron gun in one of the upright sides. The electrons leaving the gun receive line frequency deflection by means of a conventional yoke forming a plane and arrive at the opposite upright side of the tube where they may be deflected through more than 180° in the same plane by a novel magnetic electron mirror. The return beam always travels in a straight line toward the electron gun in a plane perpendicular to the sides of the tube (the xy plane). One surface (in the xy

plane) of the tube is coated with a phosphor and all that is needed for the return beam to play upon this phosphor coated face plate is that the electron beam be deflected through a small angle in the z direction toward the face plate. This is done by providing a suitable electrostatic field at the site of the magnetic mirror. A sawtooth voltage at the electrostatic pole pieces which is properly synchronized with the line frequency will cause the electron beam to sweep out a raster.

The geometry of the tube and the position of the magnetic electron mirror make it possible to view the tube from both sides as in the design previously mentioned. The use of transparent phosphors in such tubes makes it possible to see through the tube and at the same time paint a luminous picture thereon.

The tube mentioned above, while not in final form, is in operating condition.

Another design [5] proposes the use of wide-angle deflection of two cathode ray tubes, joined in such a way as to provide a single tube of dimensions which allow it to be classified as a flat device. This tube also makes it unnecessary to have special switching circuits.

Provision for color in these tubes is in the developmental stage. In one design [6], while the tube planned is fairly large and bulky, color is obtained by using three electron guns, each having its electrons deflected in the manner described for the tube shown in Fig. 1.

2) Image Intensifier-Type Tubes: The image intensifier type of panel display holds great promise for meeting many of the requirements deemed necessary for a flat display mentioned earlier. One design still in the preliminary stage [9] achieves a flat display in the following manner: Electromagnetic radiation forms an image on a photocathode which covers one face of an evacuated flat tube. The opposing face, which may be an inch or so away, is coated with a phosphor. The space between these two surfaces is filled with a system of parallel secondary-emission dynode surfaces whose function is to cause multiplication of the original photoelectrons. These dynode surfaces channel the original photoelectrons along with the secondary electrons liberated from the dynode surfaces in a straight path from the cathode to the phosphor.

The dynode surface, which may be a sheet of nickel treated to make it an efficient secondary electron emitter, is about 0.5 mm thick. The sheet is perforated in such a way that the openings are funnel shaped. In operation a thin perforated sheet of mica whose cylindrical perforations are in registry with the dynode surface is placed next to the photocathode surface and provides insulation between the cathode and the dynode surfaces. The dynode surface is placed so that the larger openings of the funnels are next to the mica sheet. A 200-volt dc voltage is applied between the cathode and the dynode surface so that when radiation strikes the photocathode the liberated electrons are accelerated toward the dynode surface through the mica insulating layer. These electrons strike the funnel shaped openings of the dynode and emit secondary electrons. Another similar mica layer with a second identical dynode surface, at 200 volts higher potential than the first, is placed over the first dynode surface with the openings again in registry. Additional secondary electrons are emitted and the process may be continued until the number of dynode surfaces reaches ten or more.

Since at each dynode surface one incoming electron may produce from three to ten secondary electrons, we have a method for greatly intensifying the original photoelectric emission at the cathode.

In the last dynode stage the emitted electrons find their way to the phosphor and there produce a greatly intensified image.

There are at least two [10], [11] other laboratories which have been working along similar lines and the device shows great promise. Advantages of the image intensifier are many. An area cathode can be used that is sensitive to that portion of the electromagnetic spectrum which is of interest. The area of the tube is not limited, as the mica insulators and the dynode surfaces are layered in such a manner as to produce structural reinforcement of the surface plates. This opposes the stresses on the plates caused by atmospheric pressure and a sturdy device is obtained without resorting to thick face plates. High resolution (30 lines/inch) storage, selective erasing and the production of halftones are possible future developments.

A modification of a tube of this type might be substituted in the future for the usual screen in X-ray fluoroscopy to provide brightness in excess of present screen. This would have the advantage of allowing a smaller X-ray dosage. Many other applications requiring light amplification are possible and development is taking place along these lines.

B. Displays Using Solid-State Materials

These may be further classified into: 1) the xy matrix type where an electrical signal is used to produce a light spot in a desired portion of the panel which may or may not be stored, 2) the type where one writes with a moving light source, the path of which is then stored as a visual image, 3) light intensifiers and image converters.

A brief general description of an EL cell is included at this point since the material used in categories 1, 2, and 3 for the generation of light is based on the EL cell. An excellent review paper covering the earlier experiments on EL was published in 1947 [12].

The fabrication of the EL phosphors is an art which is continually developing and many improvements of the original phosphors have already been made [13]. An EL cell operates in the following manner: A thin layer (1–4 mils) of an EL phosphor (usually zinc sulphide crystals suspended in a transparent plastic or ceramic dielectric) is placed in a strong alternating electric field of up to 10^5 volts/cm. The crystals are activated in various ways [16] and, when properly activated, they will emit light when the field is applied. The mechanism for light emission has been described adequately in the literature [12]-[15].

Initially, EL cells were not overly bright or efficient. Ten to fifty foot-lamberts is a brightness range usually mentioned. There has been much effort to achieve higher brightness [17]-[19] for EL lamps. A high figure recently quoted [20] is upward of 4000 foot-lamberts for a green cell operated at 20,000 cycles. However, it must be pointed out that these figures apply to experimental models operated at high voltages and frequencies which may result in lifetimes of a few minutes or so.

Recently, much effort has gone into improving the life of these cells. Apparently a high humidity will decrease their usable life. Consequently, even though plastic resin embedments generally give higher brightnesses than ceramics, the latter are continually being investigated [21] because they are less affected by humidity and are generally more durable. At least two industrial laboratories (Sylvania Electric and Westinghouse Electric) are engaged in a fairly large effort to improve the characteristics of the ceramic lamp.

Other advantages for the ceramic cell are its ability to withstand impact, its resistance to chemical attack, and the greater resistance of its dielectric to temperature and voltage breakdown.

In addition to the properties of the EL lamps listed above, a novel feature that they have is that their color can be changed by a shift in the applied ac frequency [13], [17], [19]. This is brought about because these phosphors produce one or more light bands in a specific wavelength region. The effect of frequency change is to alter the relative amount of light emitted in each band and in this way produce a color change.

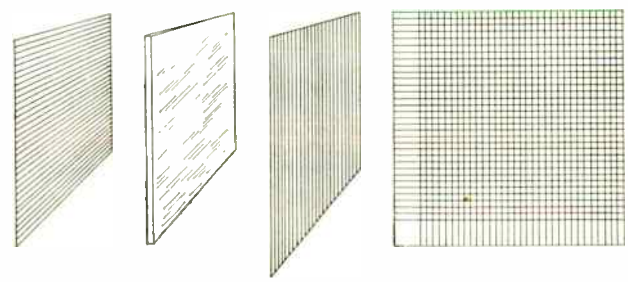
In addition to brightness and color, the efficiency of EL cells has received a good deal of attention [19]. While, as mentioned above, brightness increases with both frequency and voltage, efficiency in general does not. Optimum efficiency so far has usually been achieved at intermediate voltages. For example, plastic embedded cells, with brightnesses of 50 foot-lamberts, operating at 200 to 300 volts and between 500 to 1000 cycles and with efficiencies of 8 to 14 lumens/watt, are being produced.

Many theoretical aspects of EL cells have already been investigated [13], [22]-[28] and present activity indicates a continuing endeavor.

We will now discuss the types of panel displays which make use of the EL phosphor.

1) XY Matrix-Type With Electrical Input: The type represented here has been the subject of considerable interest on the part of many academic, governmental and industrial laboratories. These all use variations of a basic idea [29] which places an EL phosphor between an xy array of grids. If a voltage is applied to one line of x and one line of y grids, the voltage drop across the area where the grids overlap will be twice as great as across either grid. If the total voltage drop is of the proper magnitude, the EL phosphor will luminesce at that area. The variations of this idea are primarily concerned with overcoming problems of crosstalk, stability, and brightness.

In one design [30] the panel is constructed of three layers (see Fig. 2): a horizontal or x array of parallel conductors, an EL layer, and a vertical or y array of conductors. These are applied to a base which may be glass or metal.



(a) (b)

Fig. 2—Mobile-dot lamp. (a) Exploded view. (b) Front view assembled panel.

By using a suitable frequency, and establishing the proper voltage at a selected x line and y line, light will be emitted from that particular area of the EL phosphor between the crossed elements.

The outside layer which houses the arrays must be transparent and one of the conductors constituting the x and y arrays must also be transparent (tin oxide, etc.).

One serious problem encountered with an unmodified xy matrix panel is that of crosstalk. In general, when a voltage is applied to an x and y line, an attenuated light signal appears across a whole line of the x grid as well as a whole line of the y grid. Of course at the point of intersection of the two grids the brightness is much greater than along each line. The light signal that gets through adversely affects the wanted signal.

There are possible ways to overcome these difficulties and in general they employ two methods of approach. 1) A modification of the EL material in such a way that its brightness-voltage curve may have a steeper slope so that with half the applied voltage the brightness of the phosphor is so low that the contrast ratio is satisfactory. 2) Use of nonlinear elements in series with a phosphor so that the voltages across the EL phosphor can be easily modified. While approach 1) is not being neglected a great deal of interest is centered on approach 2).

Two of the laboratories [31], [32] using approach 2) are unable to supply details for publication.

The EL panel may be regarded as a lossy capacitor with a power factor of about 0.1. Since large audio frequency voltages are required for high light output, the dissipation in EL cells and driving circuits may run high. This requires a switching system which will handle large amounts of power and will switch in micro or milliseconds. Switching will be mentioned again in Section III.

One system which employs approach 2) uses ferroelectric nonlinear capacitors [33]. In these capacitors, the dielectric material is the ceramic barium strontium titanate $Ba_xSr_yTiO_3$.

One flat display screen combines the ferroelectric capacitor [34] with the EL screen (ELF) to solve some of the problems of brightness control and storage. Fig. 3 illustrates schematically one way in which the ferroelectric and EL capacitors might be arranged in a multi-element array. The EL cell emits light which varies with the alternating voltage appearing across it; this voltage is determined and changed by varying the bias or control voltage applied to the ferroelectric.

In this application of the ferroelectric, the property of interest is the change in capacitance as a function of applied voltage.

In the circuit shown in Fig. 4 an EL and ferroelectric capacitor are connected in series with an ac source called the light power supply. This might be 200 volts rms at 10,000 cycles. Provision is made for a control voltage which might be supplied by a battery in series with an isolation resistance.

Assume that the parameters are adjusted so that with zero control voltage the alternating voltage across the EL and ferroelectric cells are 150 and 50 volts, respectively. This corresponds to the ON or bright condition for the EL phosphor; if now the control voltage is increased the capacitance of the ferroelectric capacitor decreases. A larger fraction of the voltage from the light power supply appears across the ferroelectric and a smaller fraction across the EL cell. With the control voltage increased to 300 volts the alternating voltage across the EL cell might be reduced to 50 volts and the EL cell would go into the dark or OFF condition.

If a phosphor is employed which has a voltage-brightness curve as shown in Fig. 5, then for a drop in voltage from 150 to 50 volts the brightness of the cell would fall from 243 to 6 foot-lamberts, with a resulting contrast ratio of nearly 40 to 1.

Sack [34] mentions fabrication techniques and includes other circuits which improve the screen characteristics.

Included in Section III are designs which show how the EL ferroelectric may be used in switching devices.

2) Storage Display With Optical Input and Optical Output: In most of the *xy* matrix panels currently being studied, the panel is designed to be used in conjunction with a storage display. In these displays the writing is accomplished by means of a light signal. The light signal triggers a cell, which consists of an EL phosphor in conjunction with a photo element, to the ON state. The operation is carried out in such a way that the phosphor may be kept in the ON state for a predetermined length of time; of from 0.1 second to an unlimited time. This produces a stored signal which can be erased at will, as shown below.

The problems encountered in these devices are "spreading" of the stored images, dielectric breakdown of the associated EL and photo element, low brightnesses, crosstalk between cells of the matrix, and poor resolution; *i.e.*, the number of cells per inch. Where possible a description of solutions to some of problems will be given below.

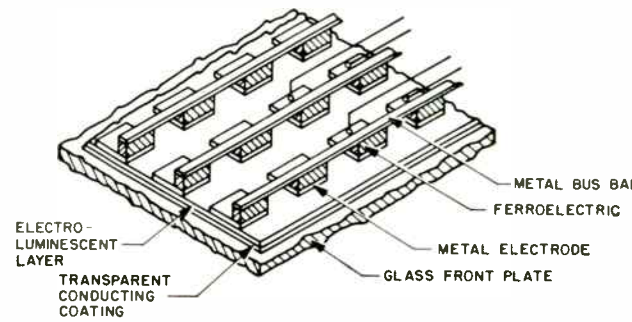


Fig. 3—Multi-element ELF array.

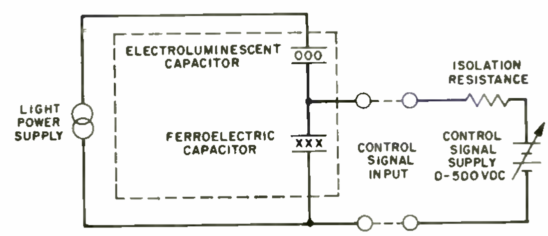


Fig. 4—Equivalent circuit of the two-component ELF elements.

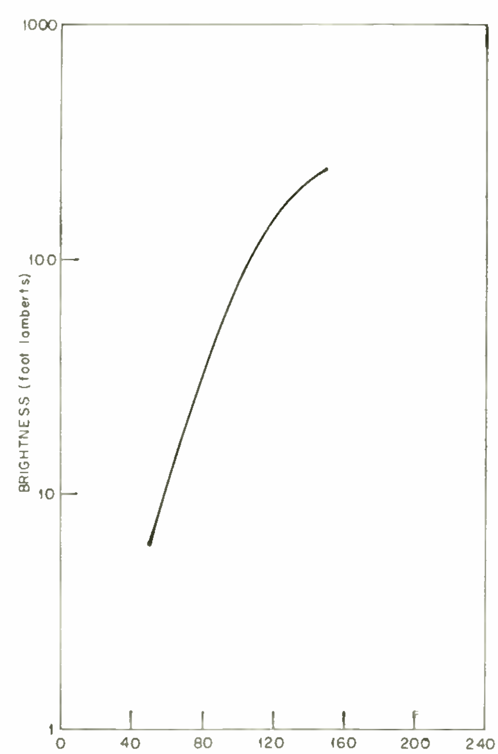


Fig. 5—Brightness of a Westinghouse electroluminescent cell as a function of voltage.

In one design [35]–[37], a transparent conductive layer is applied to a flat glass base over which is applied an EL phosphor layer. Then, as shown in Fig. 6, glass pillars, photo-formed glass or glass spheres are placed on top of the EL material. Fig. 7 shows a small section of the mosaic, modified with a cover glass.

A glass base with its transparent conductive coating, B (see Fig. 7), has the EL layer, C, applied to it. In more recent designs the embedment material of this layer is ceramic. Glass columns, K, with transparent conductive caps, D and G, on both ends are fused to the EL layer. A light attenuating layer, E, covers all the remaining surface of the EL area. Photoconductive material, F (usually made of CdS doped with $CuCl_2$ and $CdCl_2$), is placed between the columns and sintered, and then a cover plate, J, with a transparent conductive film coating, I, on its underside is placed over the whole.

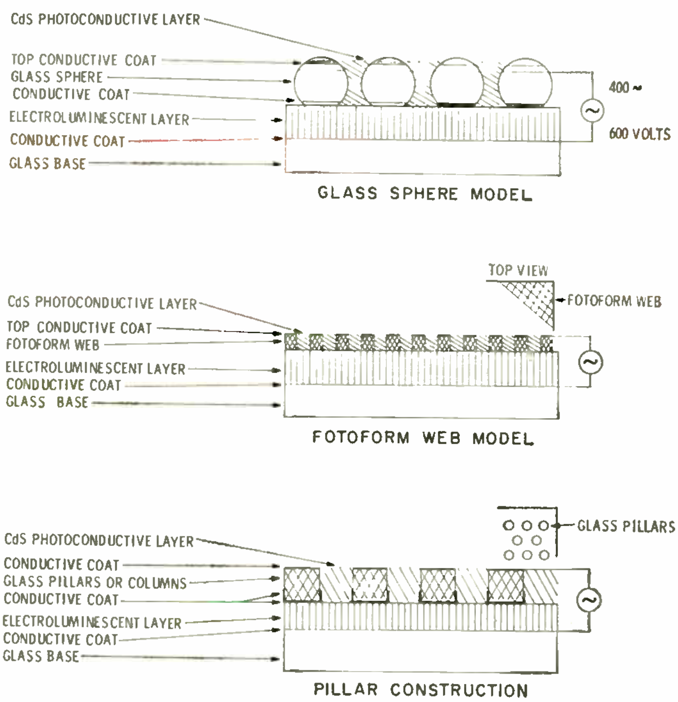


Fig. 6—Experimental models of solid-state storage devices.

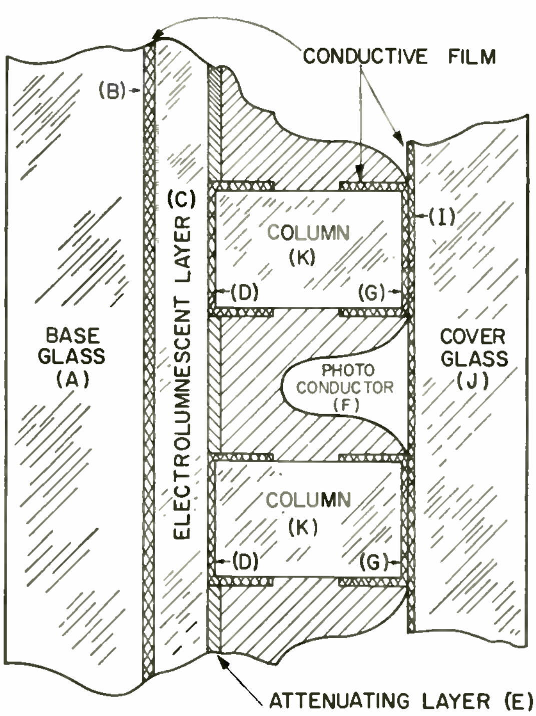


Fig. 7—Mosaic panel construction, cross section.

The operation of the mosaic panel is as follows. The conductive film, B, at one end, and the conductive film, I, at the other are connected to opposite sides of an ac supply. The conductor film caps, G, at one end of the glass column assume approximately the same potential as the conductive film, I. In total darkness or low ambient light, the photoconductive material, F, will not complete the circuit between the conductor film caps, D and G, on the columns. This is then the initial condition. Now assume that one of the elements is triggered by means of a light signal. (This may be provided by the xy matrix panel mentioned previously.) Light of sufficient brightness will pass through the cover glass, J, through the conductive caps, G, and into the column. The scattered light (the pillars are frosted) will strike the photoconductor, F, attached to the wall of the column. Note that the photoconductor is applied with grooves or channels in it so that a maximum area of the photoconductor receives light. In the presence of this light, the resistance of the photoconductor will rapidly decrease and within a few milliseconds will present a low resistance (about 10^{-4} to 10^{-6} of its dark resistance) between conductive end caps, G and D. The conductive end caps, D, will then assume the potential of the conductive film, I, and the EL layer directly adjacent to the conductor end caps, D, will then have a major part of the alternating voltage across it in this area and will emit light. This emitted light will travel in both directions; it will pass out through the glass base, A, and become one point in the display, and it will also pass up into the column. The light which passes into the column will be sufficient to keep the photoconductor in the conducting state. The original triggering light can now be removed and the element will continue to emit light until the display is erased by removal of the ac voltage.

The conductivity level of the photoconductor which is required for continuously holding an element lit is dependent upon the magnitude and frequency of the source voltage applied between the conductor films. For an excellent discussion of photoconductors, see Bube [66], [67].

The photoconductor may be of the doped CdS variety, and again as with the EL phosphors the properties of the photoconductor may be strongly influenced by the methods of production and application. A typical spectral response for the photoconductor mentioned is shown in Fig. 8. Since there is little volume conductivity of the photoconductor the material is applied thick enough so that it will absorb all the radiation.

A typical photoconductive layer response to light and

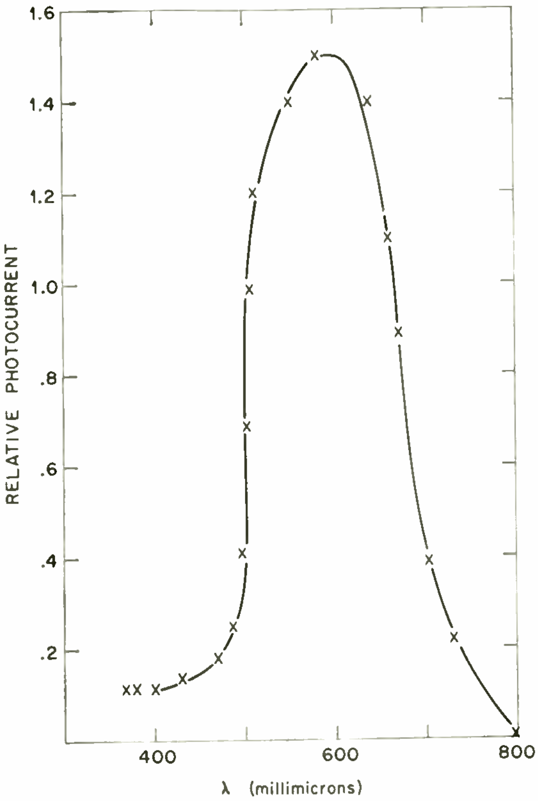


Fig. 8.

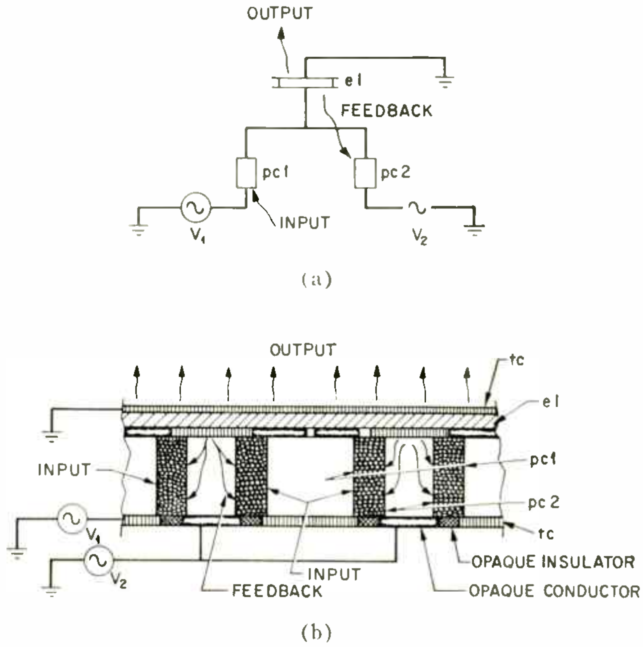


Fig. 9—2-pc EL structure (schematic). (a) Circuit schematic. (b) Structure schematic.

applied voltage is $I_{pc}=9\times10^{-5}\ EV_{pc}$; where I_{pc} is the photocurrent in ma/square, E is the illuminance in lumens/foot² and V_{pc} is the applied voltage. The dark resistance is of the order of 10^{10} ohms per square. Since a thin surface layer (less than 1 mil) absorbs all the illumination, volume conductivity is a minimum.

The transparent conductor requirements are: it must pass approximately 90 per cent of the visible light, have a resistivity of less than 10^3 ohms per square, and be easy to apply to the cover plates. The most versatile material is a tin oxide SnO_x which is formed on the glass by various techniques and then bonded at temperatures from 400° to 800°C. Construction techniques as well as operating details for the design of this panel are to be found in the literature [35], [37]. Other designs [31], [38], [39] are variations of the basic design described above and are concerned with trying to improve resolution, stability and brightness and to produce low crosstalk and long life. The use of a third element [31], [39] to improve the characteristics of the cell is being studied. A simple description of one such circuit [39] which provides negative as well as positive feedback is shown in Fig. 9.

Each cell consists of two parallel connected photoconductor elements in series with a common EL element. Photoconductor 1 (pc 1) is exposed to input radiation only, while photoconductor 2 (pc 2) receives only feedback radiation; *i.e.*, that originating in the EL layer. Each photoconductor can have an independent voltage source. With $V_2=0$ the panel operates as a simple image intensifier. If V_2 is subsequently applied and V_1 removed, the panel stores the picture present at the instant of switching. With V_1 and V_2 both applied in phase, the incoming information is stored continuously, and the panel operates in what might be called integrating storage. With V_1 and V_2 applied out of phase, an optical negative feedback is obtained making the input-output relationship more linear. An analogous situation is found with vacuum tube voltage amplifiers.

For more details see Loebner [39].

3) Image Intensifiers and Image Converters: Several image intensifiers have been described in detail in the literature [14], [36], [40], [41]. Because of space limitations, the description that follows is somewhat brief. Two general classes have been employed—one uses electroluminescence and the other photoelectroluminescence.

A description which might be considered a prototype of the first class is given below [40]. A transparent conducting (tc) coating on a glass plate (see Fig. 10) is sprayed with a thin (1-mil) EL phosphor layer. This is covered with a thin opaque layer over which is put a current-diffusing layer. A heavy layer of bonded photoconductive CdS powder is spread on and sprayed with a conducting film. Fine parallel grooves are now cut into the photoconductor, which leaves long 60° prisms of photoconductor with conducting lines at the apex. These lines are connected to a common terminal and act as one electrode for the device. The grooves of the photoconductor provide continuous conducting paths for the photocurrents from the electrode at the apex of the photoconductor to the bottom of the photoconductor grooves. By means of the current diffusing layer, photocurrents which would otherwise enter the EL layer at

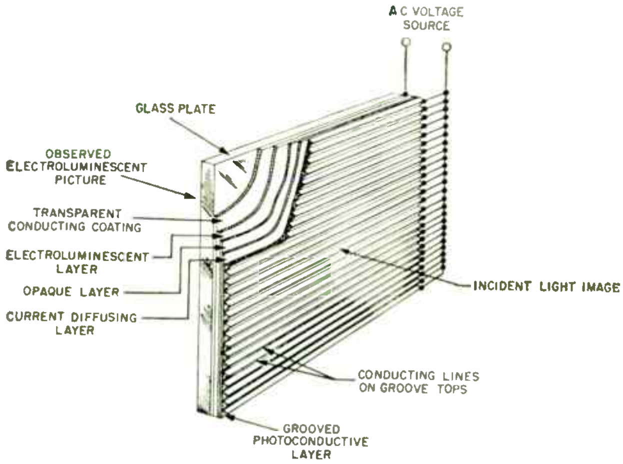


Fig. 10—Grooved photoconductor-type image intensifier.

restricted regions near the groove bottom are caused to diffuse slightly before entering the EL layer. The amount of diffusion is limited to the width of a single photoconductive groove by controlling the thickness of the current diffusing layer and so essentially all of the EL layer can be excited by the photoconductive layer and at the same time resolution of the device is not affected.

The light output of the amplifier depends on the light input, the alternating voltage and frequency of the source, as well as the characteristics of the photoconductor and EL layers.

In one operating device [40] the voltage used is 1000 volts peak-to-peak, at 400 cycles. This produced light of about 5 foot-lamberts for an input of about 0.2 foot-lambert. These values are for a tungsten filament input operating at 2870°K.

Although the usual input to the image intensifier is visible light, infrared and X-ray input images have been used with photoconductors having the required spectral sensitivities, and outputs in the visible have been obtained. Used in this way, the amplifier may act as an X-ray intensifier and image converter.

The other general class of image intensifiers utilizes the phenomenon of photoelectroluminescence. In a device using this phenomenon the incident radiation initiates and controls the light emission from the phosphor, while the power required for light emission comes from a direct voltage applied to the phosphor.

A detailed description of the preparation of a photoelectroluminescent phosphor which was initially studied because of its transparency and its lack of graininess as a phosphor screen, is in the literature [42].

These phosphor films, besides showing the phenomenon of photoelectroluminescence, produce relatively efficient luminescence under cathode ray bombardment, and may be made to exhibit electroluminescence in alternating electric fields [43] as well.

The films are produced by chemical deposition from a vapor phase reaction. The ones most easily produced are zinc and cadmium sulphide activated with arsenic, copper, manganese, phosphorus or zinc. The basic process produces a transparent film on a glass surface. It involves bringing together zinc and activator vapor at the glass surface heated to 500°–600°C in an atmosphere of a few millimeters of H_2S. Under proper conditions a clear zinc sulphide coating is deposited on the glass (see Fig. 11).

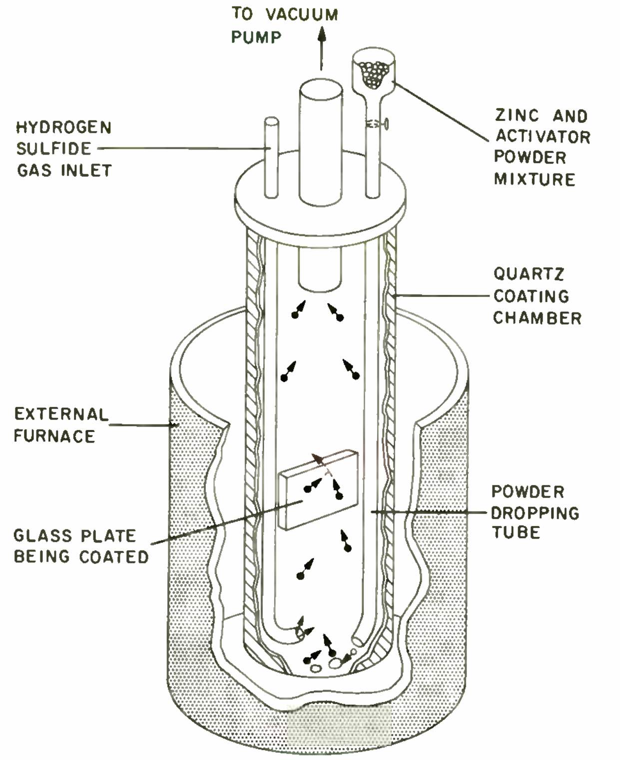


Fig. 11—Formation of zinc-sulfide coatings from vapor reaction (vapor-deposition method produces phosphor layers having desired characteristics).

Other activators may be introduced to produce different color responses. Clear coatings have been produced as thick as 5 microns in one deposit. The coatings are durable and tightly bonded to the glass and may be polished, handled and cleaned in practically the same way as the glass itself.

The advantage of this technique over one in which a powder is evaporated and then condensed onto a glass surface in a vacuum chamber is that here the activator is present at all times while in the other process the control of the amount of activator material may be difficult. Another advantage is the uniformity of the film.

Cells that exhibit electroluminescence are made in the following manner. An initial transparent coating of titanium dioxide (TiO_2) is placed on a sheet of glass, and a layer of ZnS produced by a vapor phase reaction is deposited over this coating. The TiO_2 now becomes electrically conducting as the ZnS forms on it. If now an electrode is painted or evaporated over the surface of the ZnS, an EL cell is produced.

It was while operating with a cell of this type that photoelectroluminescence was first reported [44].

The properties of the photoelectroluminescent screen can be examined by applying voltage to the electrode with ordinary batteries (see Fig. 12). In making a screen which will respond to the near ultraviolet, $ZnCl_2$ is used along with Zn and $MnCl_2$ in the vapor reaction with H_2S. When ultraviolet radiation of about 3650 Å is incident through the glass and the TiO_2 film, a faint yellow-orange photoluminescence results. If now a dc field of 50–100 volts (for a 10-micron thick film) with TiO_2 positive is applied the brightness is greatly increased.

In a typical case for low-intensity ultraviolet excitation (1–2 microwatts/cm^2), the amount of energy radiated in the visible while the field is turned on is 6 to 8 times greater than the incident ultraviolet. The amplification is nearly linear, thus good contrast is maintained during amplification.

If sudden changes are made in incident intensity the light emission requires finite times to adjust itself to new values. These times range from a few hundredths of a second to several seconds, corresponding respectively to high and low radiation intensity. One of the most probable applications of these phosphors in the near future will be intensified fluoroscopic screens.

C. Displays Using Liquids

A flat xy matrix-type display using a liquid as the display material (called an "electrofluor cell") has been developed [45] and is shown in Fig. 13. Response curves at different voltages are shown in Fig. 14.

A flat plastic sandwich with a transparent top surface is filled with a partially conducting liquid which contains a chemical indicator. Thin (0.003-inch) parallel platinum wires are attached to the underside of the top surface and run the length of the surface in the y direction. They are connected to one terminal of a dc source of the order of 1 volt. The other terminal of the voltage supply is connected to a row of dot electrodes in the x direction. These dots, which may be made of platinum, are embedded in the upper surface of the lower plastic layer and are shown to scale in Fig. 13.

By means of a suitable switching arrangement, continuous application of the direct voltage across any xy row of electrodes will cause a reaction at the intersection which will produce an electrochemical change in a fairly well-defined zone. With the proper indicator, this will be observed either as self-luminous fluorescence in the presence of ultraviolet radiation or will appear colored by selective subtractive absorption of white light. In most types, the color effects occur at the negative electrode.

Voltage pulses may be as long as a second but a prolonged dc pulse will cause permanent damage to the bead electrode. The damage results because of electrolysis with the formation of gas bubbles and loss of the luminous or color response. A reverse pulse will reverse the electrolytic reaction and cancel the response.

The brightness response, since it is due to electrolytic deposition, varies with the quantity of electric charge

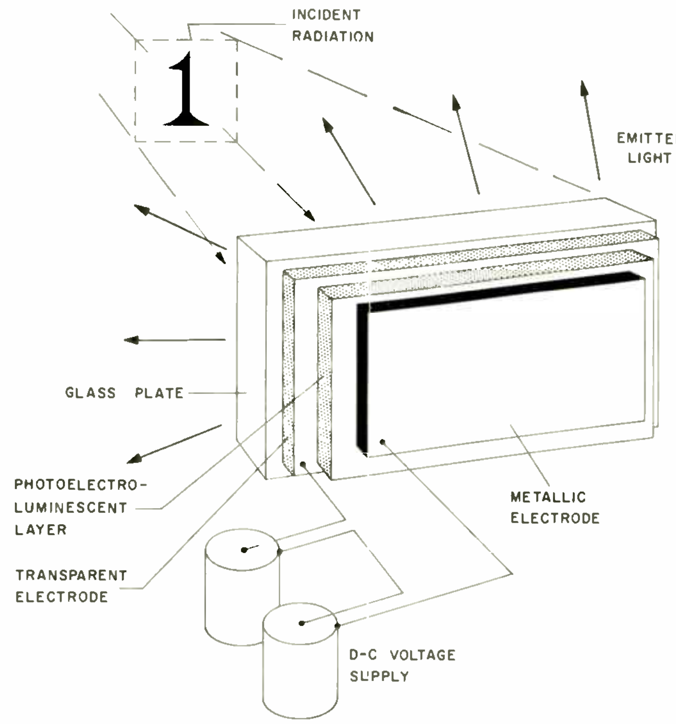


Fig. 12—Photoelectroluminescent light amplifier (incident radiation initiates and controls light emission).

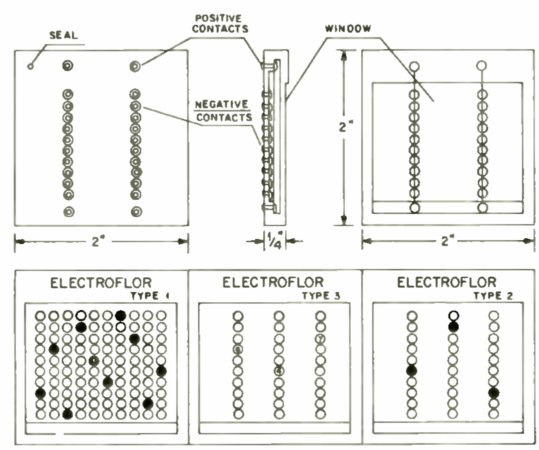


Fig. 13—Typical electrofluor cells.

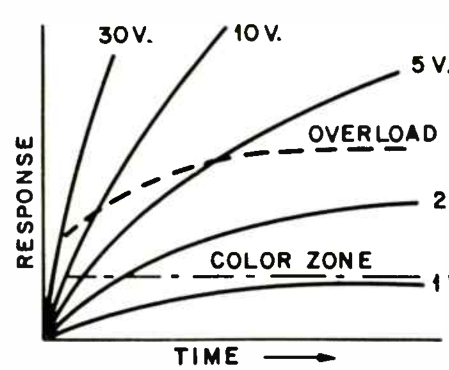


Fig. 14—Typical response curve.

passed through the cell. There is a threshold below which no response is observed. This is indicated in Fig. 14 by the dashed line called color zone.

Simple circuits are available which will enable the indication to turn on rapidly without going into overload and to erase after a time interval.

A number of flat transparent cells may be stacked together to form a 3D matrix. Each layer may be made to exhibit a different color.

D. Gaseous Displays

1) XY Matrix-Type: The flat panel using a gas as the display material may be thought of essentially as a large array of tiny closely stacked gas discharge or fluorescent bulbs connected by a crossed wire matrix.

One of the first designs developed [46] consists of three basic layers or sheets as seen in Fig. 15. These sheets are made of glass approximately one-eighth inch thick. The center sheet in the sandwich is a multicellular structure having sixteen holes per linear inch. Each cell in the center sheet has its sides coated with phosphor. The other two layers constitute the front and back plates of the sandwich and each has on its inner surface a series of parallel wires or aluminum lines so that one set of lines can form a horizontal set of stripes backing the holes and the other set can form vertical lines in front of the holes. These lines or wires are thin so that the view from front to back through the holes is virtually unobstructed. The entire device is solder glass sealed and is filled with mercury vapor and an inert gas.

The operation of the cell consists of connecting a vertical and horizontal line to a source of voltage and, as in other xy matrix devices, a voltage exceeding the firing voltage may be impressed on a single cell. This voltage ionizes the gas and initiates the glow discharge in the selected cell. Since the glow contains visible as well as ultraviolet emission, the light output may be enhanced by inclusion of a phosphor to convert the ultraviolet energy into visible light. Removal of the impressed voltage allows for deionization and extinction of the glow discharge.

In the case of a gas discharge the value of the firing voltage is fairly critical and so a cell does not fire until the voltage is exceeded. By using a proper cell geometry, gas pressure, etc., the normal constant voltage characteristics of the gas discharge cell may be altered and enough positive resistance added to each cell so that a bistable voltage ampere characteristic is obtained as in Fig. 16 [46].

This characteristic allows the device to be biased at a holding voltage which is adequate to maintain the cells in the ON or OFF condition. Then the cell might be pulsed above its firing potential and ignited or pulsed below the extinction potential thereby turning it off by applying the appropriate voltages to the designated rows and columns.

The device has several limitations in its present state of development. If large luminescent portions of the display board are required, a large amount of power will be needed to operate the device.

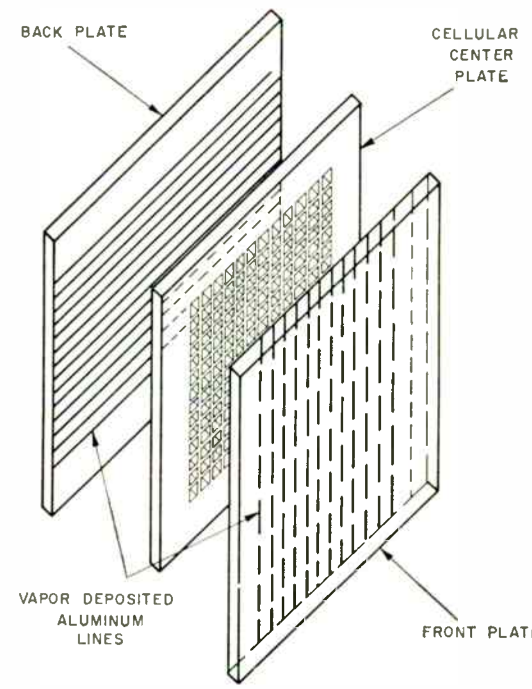


Fig. 15—Exploded view showing construction of gas discharge matrix.

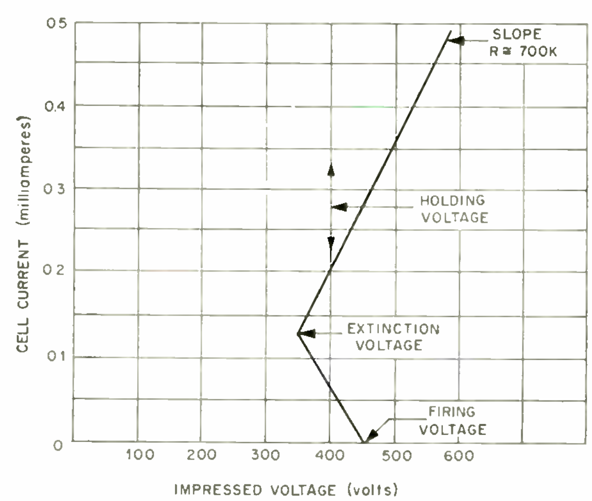


Fig. 16—Typical cell volt-ampere characteristic.

This will give rise to heat dissipation problems and cooling a large board might be difficult if large sections are luminescent. As usual with these devices, switching circuits have to be developed to handle these large amounts of power at high frequencies.

Various designs incorporate modifications of the above. For instance it is easy to modify the foregoing into a polar plot design. This has been proposed [47]. It is also possible to use a three-dimensional set of wires in a volume filled with air or with a gas at low pressures as has been done in another design [48].

E. Mechanical and Miscellaneous Displays

1) Tungsten or Neon Gas Bulb Displays: As mentioned above one of the earliest displays to be used is the one which displays news bulletins to the public at Times Square, New York, N. Y.

The switching is mechanical and consists of two conveyer belts which carry along thin bakelite slabs on each of which is placed an individual metallic letter. Under each conveyer belt is an array of contacts which are connected in such a way that the metallic letters cause individual lamps to light up corresponding to the pattern of contacts touched by the letters as is shown in Fig. 17. In this way a series of tungsten lamps are lit in succession. The effect is to produce a moving letter five feet high.

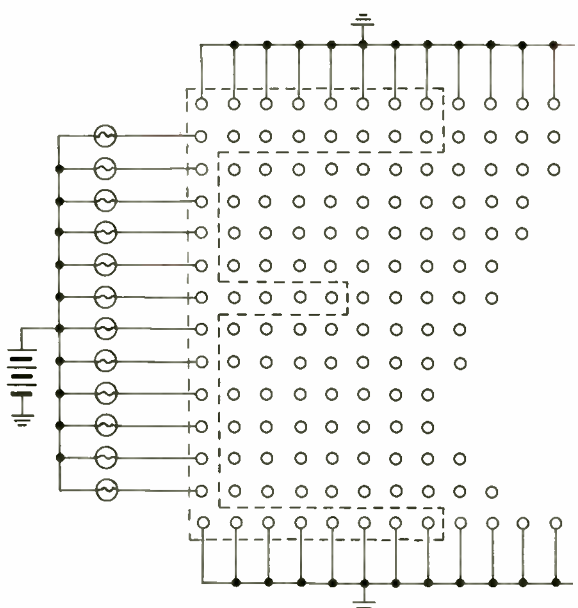


Fig. 17—The letter "F." All contacts except those grounded are attached to bulbs.

Another mechanical display used for advertising purposes is one of a series of animated displays which can display the antics of cartoon actors in black and white.

In one design [49] a bank of 4104 light bulbs are activated by 1026 photoelectric cells also arrayed as a bank in the control room. In this control room a high intensity image from a motion film projector is played on this bank of photocells. Each photocell is connected to the output of a thyratron. A light impulse from the projector hitting the photocell will trigger the thyratron and fire the four light bulbs attached to it. The time constants are such that an animated cartoon can be displayed with no difficulty.

An xy matrix of a large number of small gas discharge bulbs was proposed by the Naval Research Laboratories [50]. This could be used as a large visual display for radar information. The display would be operated by connecting all the cathodes of the bulbs in a common x row to a common bus to ground. The anodes of all the cells in a common y column could be connected to a common B^+. The bulbs operate in the same way as the individual cells in the device described in Fig. 15, except that a common envelope is not present.

Several displays of the type mentioned in the previous paragraph consisting of a small number of bulbs were built and tested. One problem encountered was the unwanted firing of extraneous cells when a given cell was energized. The problem was due mainly to variations in bulbs which allowed one bulb to fire at a lower potential than another. While this problem could be solved, the simpler method, used in the device described in Fig. 15, has many advantages, including smaller space and lower cost.

2) Units Which May be Incorporated Into Displays: These units are fairly small entities which may be grouped together to form a flat display.

One laboratory [51] makes a variety of tubes each displaying as many as ten alpha-numeric symbols. The symbols are stacked for use as a direct inline read-out device. The tube (Nixie), which comes in two sizes, $\frac{1}{2}$ inch diameter and $\frac{3}{4}$ inch high and $1\frac{1}{4}$ inch diameter and $1\frac{1}{4}$ inch high, is a gas-filled tube containing many cold cathodes and a common anode. These cathodes, up to ten in number, are stacked in such a way as to be individually visible when turned on. The same principle is featured in another tube (Pixie), but here, instead of having stacked cathodes, the glow positions are 36° apart.

Another tube, a prototype which may be called an alpha-betanumeritron, has a series of vertical, horizontal, diagonal and curved lines which can be made to glow. These used in the proper combinations can display numerals and the letters in the alphabet.

Another laboratory [52] makes a series of tubes similar to the above. In addition they make a tube which has an xy matrix in a tube filled with neon. A lattice-like structure of anodes and cathodes is provided which makes possible, by external switching, the portrayal of any of 100 luminous spots in any sequence or arrangement. The limit in this direction may be as high as 250,000 spots at approximately 400 spots per square inch.

Another unit [53] which could be incorporated into a flat display consists of a stack of small lamps in a 3×4 matrix; each lamp projects its light through a numeral stencil which has its own focusing lens system and throws a focused numeral on a ground glass plate. By switching a different lamp, a different numeral appears on the plate. While the present device can display 12 characters, a large device is being developed which can display up to 36 characters.

A similar type display [54] uses an endless belt of characters with a small driving motor instead of a projection system. The belt carries up to seventy-two display characters of 9/32 inch height and 13/16 inch width. The normal running time may be from 0.8 to 1.7 seconds depending on the number of characters.

III. Switching Devices

Many of the switching devices mentioned below are still in the development stage. Some are used to drive the flat cathode ray tubes mentioned in Section II-A, 1 and some to drive the solid and gaseous-type *xy* matrices. They are grouped into electron-beam devices and solid-state switching devices, the latter including such things as ferrite cores, transformers, ferroelectrics, transistors, etc.

A. Electron Beam Switching Devices

One design, a magnetron beam switching tube, is already commercially available [51] and has been described in the literature [55], [56]. This is a ten-position high-vacuum device with the positions mounted radially about an oxide cathode which runs along the axis of the tube. An axial magnetic field is provided by a small cylindrical magnet attached permanently to the glass envelope. Each position contains three electrodes which form, lock and switch the beam. Thus the electron beam may be formed in any one of ten positions and then switched. The tubes are available in voltage ranges from 12 to 300 volts and may be switched in from 0.1 to 1 microsecond.

Other designs which are still in the initial stages use electron guns in the conventional manner with targets which are conductors, photoconductors, and combinations of photoconductors with phosphors.

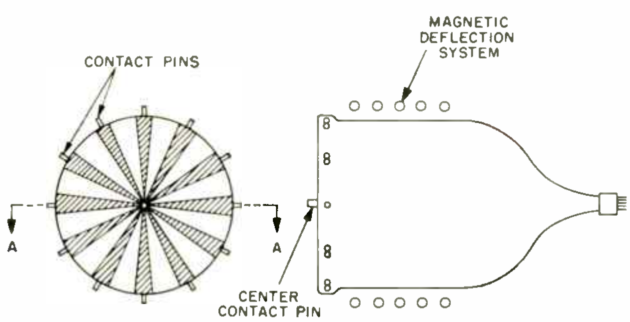


Fig. 18—Switching tube.

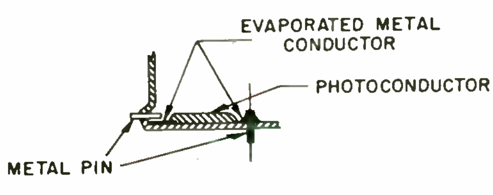


Fig. 19—Target section A-A (condition number one).

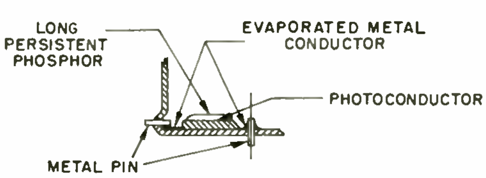


Fig. 20—Target section A-A (condition number two).

In one proposed design [57] the target of a conventional cathode ray tube is divided into a series of sectors having a common central electrode as in Fig. 18. Two possible constructions for these elements are shown in Fig. 19 and Fig. 20. In the first case they may be photoconductive layers which have the additional property of becoming conductive when bombarded with electrons as in Fig. 19, or they may consist of photoconductive layers adjacent to a cathodoluminescent phosphor, as in Fig. 20.

The photoconductive layer becomes conductive when radiation from the adjacent phosphor strikes it. The phosphor, of course, emits this radiation when bombarded by electrons from the gun.

The operation is as follows. Proper deflection is chosen so that the electron beam sweeps out a circular path around the face of the tube. The common electrode at the center is connected to a source and the lead from each sector electrode of the tube to an x array of strips of an xy matrix electroluminescent panel. A similar tube is connected to a y array as in Fig. 21. A raster may be swept out by applying a half select voltage to one x line while applying a series of half select voltages in succession to the y lines, then doing the same to an adjacent x line.

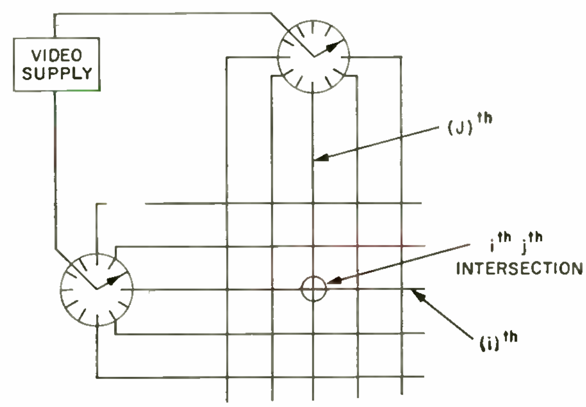


Fig. 21—Crossed-grid presentation system.

In another design [34] an EL cell in conjunction with a ferroelectric capacitor is used as the target in a cathode ray tube, as in Fig. 22. The details of the target are shown in Fig. 23. The operation of the screen has already been described in Section II-B, 1. The screen which now forms the target is modified as follows. A material having a secondary-emission ratio of greater than unity is deposited on the conducting back electrode, *a*, (see Fig. 23) and an electron collector screen is positioned behind the target as in Fig. 22. If video is applied to the collector screen, the electron beam will bring each control (back) electrode to the desired video potential as it scans the array.

The array can also be used in the manner of Fig. 24 where the control electrode is made of a material whose conductivity is changed by electron bombardment (or

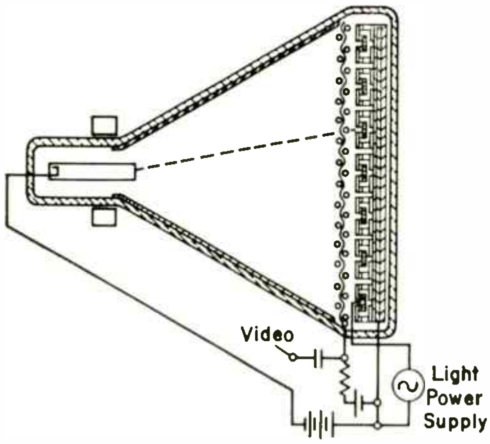


Fig. 22—Electron beam distribution to the ELF screen.

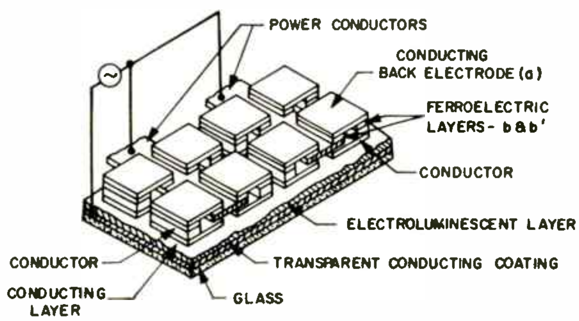


Fig. 23—ELF screen structure.

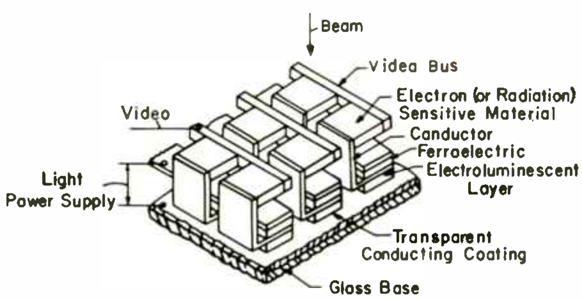


Fig. 24—Electron beam distribution using beam sensitive switches.

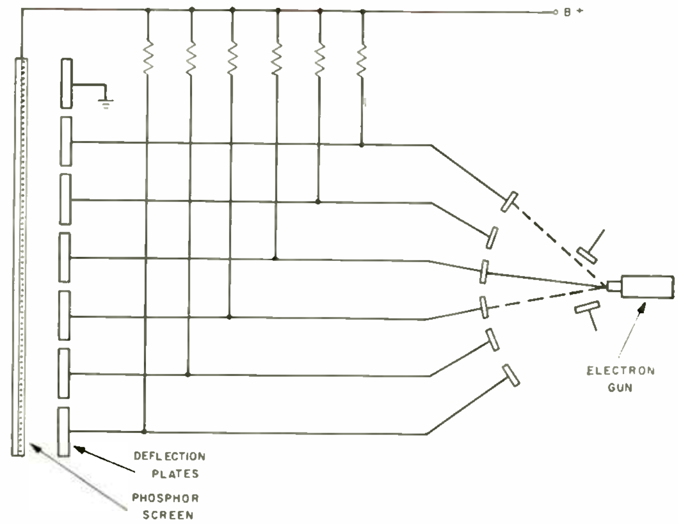


Fig. 25—Beam switching tube.

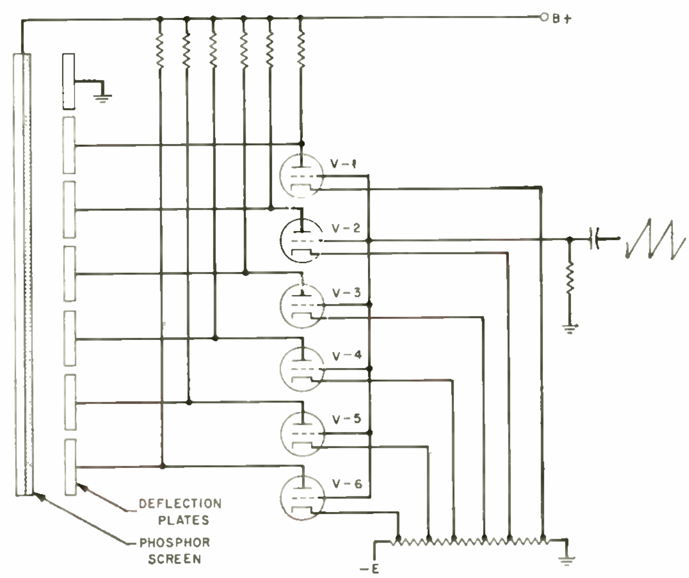


Fig. 26—Biased tube method.

radiation). The electron beam can now scan these solid-state switches which in turn commutate the video control signal among the numbers of the array. More details are to be found in the reference mentioned above.

In one scheme which is in operation [2], a beam switching tube similar to the ones described above [57] is used to achieve a sequential potential drop for switching the deflection plates of the flat tube mentioned in Section II-A, 1. This is shown in Fig. 25. Another scheme used to provide sequential voltage for the deflection plates of the same flat tube is shown in Fig. 26. There, the grids are connected so as to have an increasingly negative bias. This will allow each tube to conduct in sequence as a sawtooth voltage is applied to the grids in common. One may mention in passing that additional developments by the same laboratory to accomplish the switching function take the form of tapped delay lines and tapped pulse lines. Other laboratories [58] are experimenting with beam switching tubes and have made progress in this area.

B. Solid-State Switching Devices

Among the solid-state switching devices may be found those which make use of each of the circuit elements—capacitors, resistors, inductors and transformers. Some of these have already been used in conjunction with cathode ray tubes.

In one proposed design [59] the photoconductive layer is in the form of flat rings. The rings are spread out on a plate, a perforation is made in the center of each hole, a wire is placed in each perforation and the hole is filled with conductive material. The spaces between the rings are likewise filled with conductive material and form a common electrode. It is possible to increase the number of switch positions by using photoformed glass as the face plate. This could increase the number to as high as 1000 and possibly 10,000. This plate could now be placed so as to cover the phosphor of a cathode ray tube and switching is now accomplished by proper signals to the deflection circuits of the cathode ray tube. Details necessarily omitted because of space limitations may be obtained in the reference cited.

Another proposed design [60] uses as a switching device a small transformer with a ferrite core having a rectangular hysteresis loop as in Fig. 27. Schematically there are four windings in a circular core as in Fig. 28. A sawtooth scanning wave is applied to coil *a*, a current which provides a magnetic biasing field to coil *c*, coil *b* carries a continuous high frequency current, coil *d* is the output coil. The *b* coils on all the cores are connected in series. This is shown in Fig. 29 where only the *b* and *d* windings are shown for each coil.

If a biasing magnetic field h_1 is applied by current in coil *c*, the coil will be saturated in the negative sense. The unit will not act as a transformer, and will block the passage of high-frequency voltage to a vertical wire in the EL panel. However, as the current in coil *a* builds up, the field h_1 is neutralized and the coil changes its state from negative saturation to positive saturation. During the transition the unit acts as a transformer and passes the high-frequency voltage on the vertical wire. This gives bursts of high frequency pulses which proceed in sequence from left to right on the vertical wires of the panel, when the value of the biasing current produces magnetic biases equal to h_1, h_2, h_3, etc. in Fig. 27. The shape of this burst is shown in Fig. 30.

At the end of the scanning cycle, all the cores are switched to b^+ saturation and during the flyback they all switch back to minus saturation ready for the next scan.

Since this is to be used with an EL *xy* matrix panel the high-frequency supply should be blanked out during the flyback, otherwise a burst of light will appear over the whole panel.

For the horizontal wires the arrangement is the same with the exception that the sawtooth scanning wave is stepped while the scanning sawtooth for the vertical wires is not stepped (see Fig. 31). The steps on the scanning wave insure that a single horizontal wire is energized by the high frequency for the entire time of the horizontal scan from left to right. The steps correspond to the biasing magnetic fields in the cores. Thus the first step in the scanning wave neutralizes the biasing field in the core for the first horizontal wire and holds the magnetic bias of the core at its operating point on the hysteresis loop. The next step will neutralize the biasing field for the second core, etc. It is obvious that the output high-frequency voltage of the vertical and horizontal strip must be 180° out of phase.

The video signal is simply applied as a modulation to the high-frequency voltage. The description above applies to a particular design but similar designs have

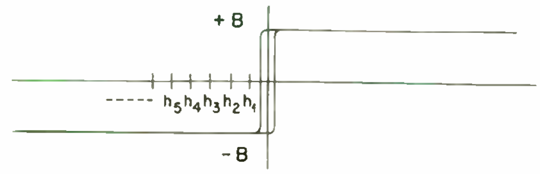


Fig. 27—Hysteresis loop of typical core.

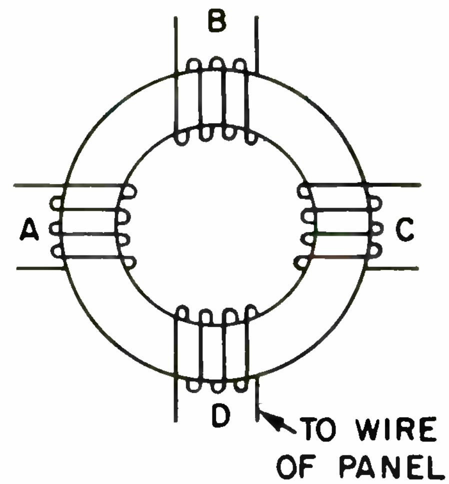


Fig. 28—Transformer detail.

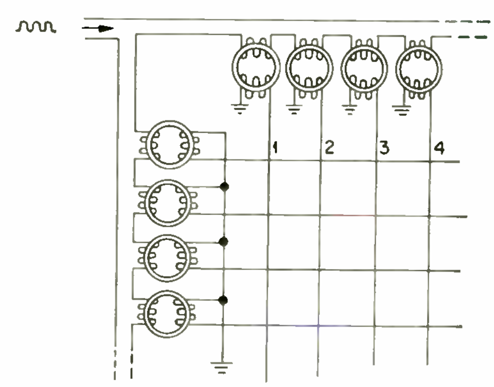


Fig. 29—Basic circuitry.

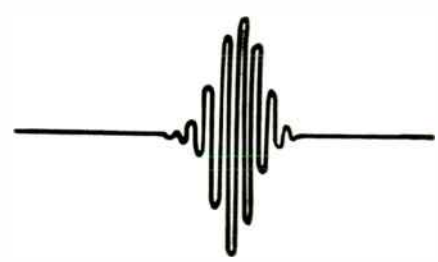


Fig. 30—Typical burst.

(a) (b)

Fig. 31—(a) Horizontal scanning wave applied to vertical wires. (b) Vertical scanning wave applied to horizontal wires.

been reported [61], [62]. The literature describes the "transfluxor" [63], a multi-aperture core, which has advantages that could make it especially suitable as a switching device for an *xy* matrix. Indeed, Rajchman and Briggs [64] indicate it has already been incorporated in such a device.

An *xy* matrix switching device using transistors in conjunction with saturable cores is in the process of development [65]. In its present state, the switching network contains 26 transistors for an *xy* array of 128×128. Each line of the *xy* array (a total of $2 \times 128 = 256$) has a half-inch saturable core which is used as a pulse transformer and as a switch. The network can deliver a pulse of up to 2000 volts, peak-to-peak, at frequencies up to 80 kc. A one-millisecond envelope of these pulses has been enough to trigger cells of an EL panel.

IV. Conclusion

From the activity noted it is apparent that the state of the art is in rapid development.

The sought-for applications mentioned in the Introduction, while not yet realized, will certainly be realities in the not-too-distant future.

While the problems of the life of these panels and their durability are still with us, the amount of time and energy being expended in this area and the advances already made indicate that progress should be forthcoming soon.

"Picture on the wall" television is already a reality in one of these designs [2] and there are already some advantages in this design, in that resolution and brightness are superior to those of existing types.

It may not be too long either before the flat display will find itself in such military and industrial applications as air traffic control boards, radar plotting boards, read-outs for computers, etc.

V. Acknowledgment

This paper is one of a series that was proposed by the Advisory Group on Electron Tubes to provide means for disseminating information thought to be of interest to the services as well as to the public.

The author is indebted to many of the people mentioned in the references for supplying him with descriptions and diagrams, many of which are included, and for the phenomenal cooperation and courtesies shown him. In particular he is indebted to Dr. W. L. Gardner, Jr. and C. L. Corderman, both of M.I.T., Lincoln Laboratory, Lexington, Mass., and to Mrs. F. R. Darne of the Bureau of Ships for the generous amount of time and assistance given during the writing of this review.

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An Improved Film Cryotron and Its Application to Digital Computers*

V. L. NEWHOUSE†, MEMBER, IRE, J. W. BREMER†, AND H. H. EDWARDS†

***Summary*—A crossed film cryotron deposited on an insulated superconductor is described and analyzed. It has a time constant of less than one μsec and is approximately 100 times faster than the original crossed film cryotron [4]. The dc dissipation is less than five μw and the active area of each element is approximately 5×10^{-4} cm^2. These cryotrons and all their interconnecting circuitry can be deposited at one and the same time in a few simple steps.**

A cryotron storage circuit and a shift register is described, based upon a principle unique to superconductors. The shift register shown is deposited in an area corresponding to 18,000 active elements per square foot. Calculations are presented to show that with this component density, a computer or memory containing more than one million elements can be operated in a one-cubic-foot container using a one-watt output liquid helium refrigerator.

Historical Introduction

THE name cryotron was applied by the late D. A. Buck to his well-known superconductive relay [1]. This device will be called the "wire-wound cryotron,"[1] and is shown in Fig. 1. It is operated at 4.2°K, the boiling point of liquid helium at atmospheric pressure. At this temperature both the niobium "control" coil and the tantalum "gate" wire are superconducting.

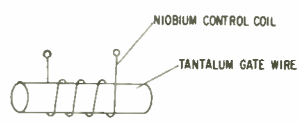

Fig. 1—Wire-wound cryotron.

The operation of the cryotron depends on the fact that the superconducting gate wire can be made resistive by means of a magnetic field generated by passing a current through the control coil. It will be recalled that any superconductor is restored to the normal resistive state while subjected to a magnetic field above the so-called

* Received by the IRE, December 1, 1959; revised manuscript received, March 7, 1960. A portion of this paper appeared in *Proceedings of the Eastern Joint Computer Conference*, December 1–3, 1959.

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[1] The IRE Committee on Superconductivity has proposed that the term "cryotron" be used for any four-terminal superconductive device in which current in the input circuit magnetically controls the superconducting-to-normal transition, and, thereby, the resistance, of the output circuit. This usage will be followed in this paper.

"critical" value H_c. The variation of H_c with temperature for different materials is shown in Fig. 2. The temperature at which H_c falls to zero for a particular material is known as the "critical temperature" T_c. Above this temperature the material is normal even in the absence of a magnetic field. A superconducting wire can also be made normal by the passage of a current larger than a critical value I_c. For thick wires of carefully annealed material, I_c is that current which gives a surface field equal to H_c. For thin films, however, the surface field corresponding to I_c can be much smaller than H_c.

It is readily seen from Fig. 2 that at 4.2°K, the operating temperature of the wire-wound cryotron, H_c for niobium is much larger than H_c for tantalum. Hence, the magnetic field produced by passing a current through the control coil, although sufficient to make the gate resistive, does not affect the control coil itself, which remains superconducting throughout. This is important whenever the coil is itself controlled by another cryotron circuit, since a cryotron gate element can best control current in a circuit in which all other elements are superconducting.

If one cryotron is to control other similar cryotrons, the critical current of the gate wire must be larger than the minimum control current required to make the gate resistive; *i.e.*, the cryotron must have gain. The gain of a wire-wound cryotron can be increased by increasing the number of turns per unit length of the control coil. This of course increases the control coil inductance and therefore reduces the operating speed of a cryotron system.

Wire-wound cryotrons are slow, due to the low resistance of the gate wire. However, when the gate is superconducting, any current it carries flows in a thin surface layer. It was thought, therefore, that the superconducting properties of the cryotron would be unaffected by replacing the gate wire by an insulated rod having a film deposited on it. The resistance of such a film in the normal state would of course be much higher than that of a solid wire, with a consequent increase in operating speed. Further fabrication advantages would result, moreover, if it should be possible to deposit the cryotron on a flat surface, because large numbers of circuits together with their interconnections could then be made at one and the same time. Such ideas appear to have occurred independently to a number of workers in the field [2], [3], including the authors of this paper. Apparently, work to develop vacuum-deposited cryotrons was started at about the same time in several laboratories.

It was shown by two of the authors [4] (and independently by workers at other laboratories), that it is indeed possible to produce a cryotron having current gain in a geometry suitable for deposition on a flat surface. This will be referred to as the crossed film cryotron (CFC). Its basic structure is shown in Fig. 3. It consists of an evaporated tin film "gate" crossed by a much narrower "control grid" of evaporated lead. The tin and the lead are insulated from each other by an evaporated film of silicon oxide. The device operates at about 3.6°K, just below the critical temperature of tin. At this temperature, the tin film has a relatively low critical field compared to that of the lead film which remains permanently superconducting. If the lead control grid is made sufficiently narrow, even a small control current produces a magnetic field which is intense enough to switch the tin gate film in the vicinity of the grid to the normal state. It is shown below that the gain of the CFC can be made arbitrarily large by increasing the width of the gate film. The CFC was first presented in public at the Electron Device Research Conference at Ithaca in June, 1959. Similar devices were described at the same meeting by M. L. Cohen, J. L. Miles, A. E. Slade, and C. R. Smallman of the A. D. Little Company, and by A. E. Brenneman and R. de Lano of the IBM Research Center.

Following a brief review of the relevant properties of superconducting films, the properties of the unshielded cryotron will be compared with the characteristics of a "shielded" cryotron which is an order of magnitude

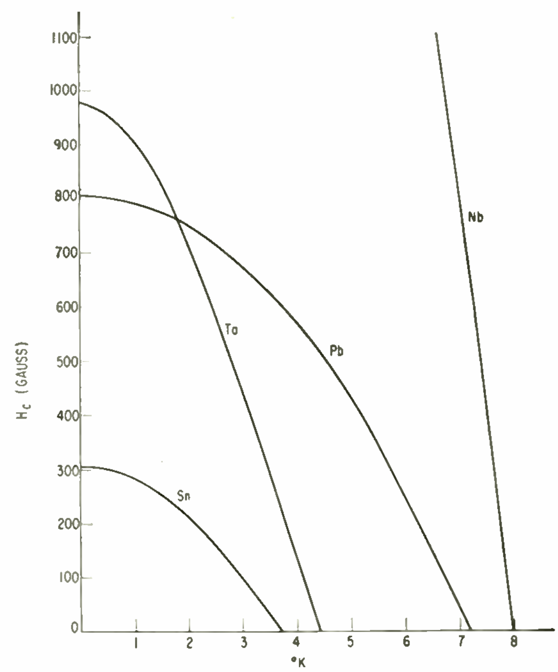


Fig. 2—Critical fields of highly purified bulk superconductors as a function of temperature.

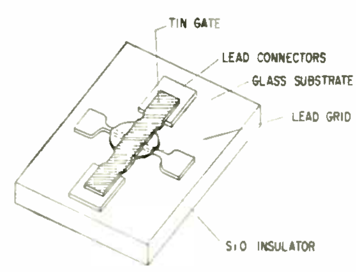


Fig. 3—Structure of the crossed film cryotron (CFC) (no shield). Typical dimensions: gate film—0.3 micron, ×2 mm.; insulator—0.4 micron; and grid film—1 micron ×25 microns.

faster. The operation of a very simple storage circuit and of a shift register will be described. In conclusion, an assessment of the role of film cryotrons in computers will be made.

Properties of Superconductive Films

Diamagnetism of Superconductors

It is well known that a superconducting body does not allow the penetration of magnetic flux, provided that the surface magnetic field nowhere exceeds H_c. Under this condition, it behaves like a perfectly diamagnetic material. If a current-carrying wire is placed in front of a superconductor, screening surface currents will be induced on the surface of the superconductor just sufficient to maintain zero internal field at the expense of increasing the external field. It is easy to show that in the case of a current-carrying wire in front of an infinite superconducting sheet, the contribution of the screening currents to the field between the external wire and the sheet is equal to that which would be produced by an "image" current. This image current is antiparallel to the external current and is the same distance behind the superconducting sheet as the real current is in front.

The case of a superconducting strip of width W carrying a current I in front of a superconducting shield is treated in Appendix I. It is shown that if the strip is close to the shield, the field between them is approximately uniform and equal to $H=0.4\pi I/W$. On the other side of the current-carrying strip the field is zero. These results are illustrated in Fig. 8. In the absence of a shield, it is easy to show that the field on both sides of the strip will be equal and opposite and of mean magnitude $H=0.2\pi I/W$. Hence, by bringing a superconducting plane close to a current-carrying strip, we double the field between the plane and the strip and reduce it everywhere else.

This effect can be used to reduce the effective inductance of superconducting elements by depositing them on top of another insulated superconductor. This is the equivalent of replacing a single wire by a "twisted pair" to reduce its inductance. It is found, moreover, and described below, that another effect of a superconducting plane is to increase the critical current of a tin film brought close to it.

Critical Temperature

The width of the temperature transition over which the resistance of our tin films falls from 90 per cent to 10 per cent of the normal value is of the order of 0.02°K. In this paper the critical temperature T_c is defined as the temperature at which the film resistance has half of its full value. A low-measuring current of the order of 0.1 ma is used to avoid heating effects.

Penetration Depth

When a thick superconducting wire or film carries a current, all of it flows in a region close to the surface. The depth of the region is of the order of magnitude of a parameter λ known as the penetration depth. λ is a function of both the material and the temperature and is given by the expression [5a]

$$\lambda = \frac{\lambda_0}{[1-(T/T_c)^4]^{1/2}}, \tag{1}$$

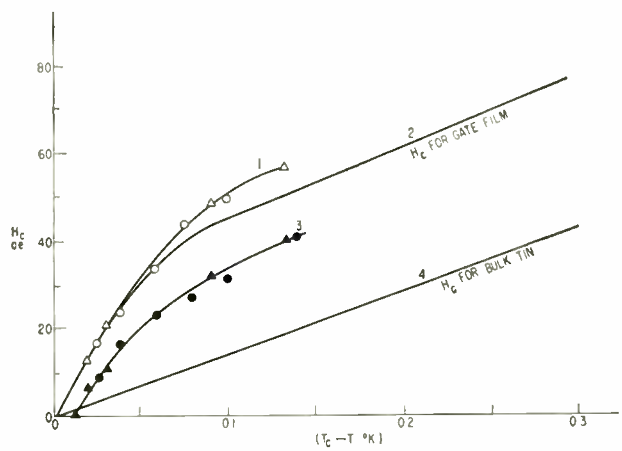
where λ_0 is the penetration depth at absolute zero. For tin, $\lambda_0=0.05$ micron, and at 3.6°K, the CFC operating temperature, $\lambda=0.1$ micron. It is shown below that the properties of superconductors differ from the bulk values when one or more of the specimen dimensions is decreased to the penetration depth.

Film Critical Field

H_c for thin films is higher than for bulk material. Published data can be fitted to the equation [5b]

$$\frac{(H_c)\text{ film}}{(H_c)\text{ bulk}} = 1 + \frac{\lambda'}{d}, \tag{2}$$

where d is the film thickness, and λ' is of the same order of magnitude as λ and has the same variation with temperature. H_c as a function of temperature for bulk tin and for a 0.3-micron film is shown in Fig. 4. In accordance with (1) and (2), (H_c) film$/(H_c)$ bulk decreases as $T\to 0$. [The other curves shown in Fig. 4 are calculated from the electrical characteristic of CFC and are discussed in connection with (7) below.] H_c for the film was measured by placing it in a solenoid producing a field parallel to the film surface, and measuring the film resistance as a function of the field. The measuring current used was kept small compared to I_c. The sample geome-



Curve 1—H_c for 0.3-micron tin gate film calculated from grid current measurements using (7). (△ 80 micron grid width, ○ 25 micron grid width.)
Curve 2—H_c for 0.3-micron tin film measured directly. [Geometry of Fig. 5(b).]
Curve 3—H_c calculated from grid current just sufficient to produce gate resistance using (7). (▲ 80 micron grid width, ● 25 micron grid width.)
Curve 4—H_c for bulk tin.
(The intercept of curve 3 with the abscissa corresponds to the width of the film resistance transition with temperature at zero field.)

Fig. 4—H_c for tin as a function of temperature.

try used is shown in Fig. 5(b). Contacts 1 and 2 are the current terminals, 3 and 4 the voltage terminals. The four leads are soldered directly to the tin film. The reason for using this geometry for the critical field measurements is that since H_c varies inversely with film thickness, the edges of the films, which are thin, remain superconducting at higher fields than the rest of the film if the measuring current used is small. The geometry of Fig. 5(b) overcomes this problem because no superconducting path between the contacts exists once the center of the film has become normal. Critical currents must be measured with the geometry of Fig. 5(a), because, in this case, the film must have a well-defined width. The thin edges of the tin film present no problem here because I_c *decreases* with film thickness.

Film Critical Current

The variation of resistance R with current for a 0.3-micron tin film is shown in Fig. 6. By applying the current in short pulses it is possible to obtain the so-called isothermal transition shown in the broken line. This curve is connected with the actual superconducting behavior of the film and is reasonably independent of other film characteristics such as resistivity and of the substrate properties. If a *slowly* rising current is passed through a film, Joule heating causes thermal "propagation" of resistive areas in the film [6]. This behavior is shown in the solid curve of Fig. 6 and is strongly dependent on substrate thermal conductivity and on film resistivity.

The critical current I_c will be defined as that current at which resistance first appears. For thin films at a temperature not far below T_c,

$$I_c = j_c W, \tag{3}$$

where W is the film width [7]. j_c increases as T approaches zero somewhat like H_c, and at absolute zero is proportional to the square root of the film thickness [8]. It appears to depend somewhat on heat treatment and film substrate.

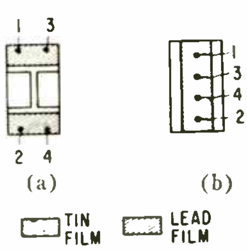


Fig. 5—(a) Specimen geometry for measuring I_c for films. (b) Specimen geometry for measuring H_c for films.

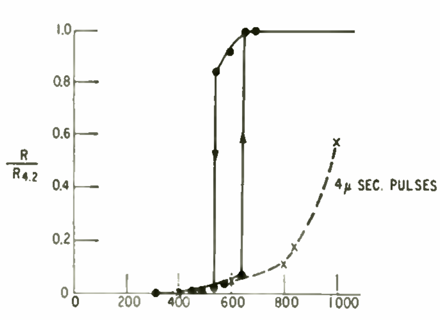


Fig. 6—Current-induced transitions for 0.3-micron thick, 4.05-mm wide tin film on sapphire substrate with $T_c - T = 0.09°K$, —●—●— slowly varying current, —×—×— 4 μsec pulses at 50 cps.

It is believed that, in a current-carrying film, the superconducting-to-normal transition starts at the film edges because the surface current density is highest there. It can be shown that the proximity of a superconducting shield plane makes the magnitude of the magnetic field near a current-carrying film more uniform than it would be in the absence of a shield plane. This reduces the current concentration near the film edges. It is to be expected, therefore, that the current at which resistance appears should be higher for a film close to a superconducting shield plane than for a film which is not. In agreement with this reasoning, it is found experimentally, and shown below, that the mean surface current density j_c, at which resistance first appears in a 0.3-micron tin film, is increased by over 300 per cent if the film is deposited on top of an insulated shield plane.

The Crossed Film Cryotron

DC Characteristics

The basic structure of a CFC is shown in Fig. 3. As stated above, a sufficiently large current passed through the superconducting grid generates a magnetic field which produces a resistive channel across the tin gate film. The characteristics of the unshielded cryotron are shown in the broken curves of Fig. 7. The curve intersecting the left-hand ordinate at 91 ma shows values of gate and grid current at which resistance just begins to appear. Its intersection with the ordinate defines I_c for the gate film used. It can be seen that the presence of gate current "helps" the grid current to make the gate resistive.

The gain of the unshielded CFC can best be defined as the ratio of the maximum current I_c which the gate can carry and remain superconducting, to the minimum grid current i_g required to make the gate resistive at low gate currents, *i.e.*,

$$g = I_c / i_g. \tag{4}$$

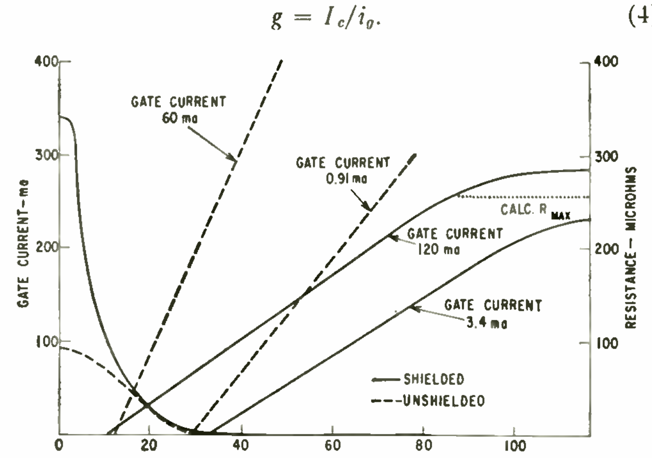


Fig. 7—Comparison of electrical characteristics of unshielded and shielded CFC. Grid width—30 microns; gate width—2 mm; $T_c - T = 0.08°K$.

For the unshielded cryotron, $I_c=91$ ma and $i_g=31$ ma at the temperature shown. The other broken line curves shown in Fig. 7 refer to the right-hand ordinate and represent the variation of resistance with grid current at constant gate current.

If the CFC shown in Fig. 3 is covered with an insulating layer followed by a film of lead, the inductances of the gate and grid will be reduced as explained above. The dc characteristics are found to be changed as shown in the solid lines of Fig. 7. It can be seen that I_c for the shielded CFC is approximately three times that for the unshielded one, with a corresponding increase in gain.

It is also noteworthy that the curves of gate resistance as a function of grid current for the shielded CFC have a smaller slope than the curves for the unshielded CFC and approach a saturation value of resistance R_{max}. This is because the field, due to a grid above a shield plane, falls off very rapidly beyond the edge of the grid since in this region the fields due to the grid current and the shield current tend to cancel one another. This is demonstrated in Fig. 8. It is to be expected, therefore, that the portion of the gate film which can be made resistive by grid current action is that portion lying under the grid only. The maximum calculated resistance of the shielded CFC on the basis of this assumption is shown dotted and is seen to be in fair agreement with experiment.

Gain of the Shielded CFC

We will now show that the gain of the shielded CFC is proportional to the ratio of the gate width to that of the grid. That this holds for the unshielded CFC was shown by Newhouse and Bremer [4]. As stated above in (3), for thin tin films I_c is proportional to the film width W, *i.e.*,

$$I_c = j_c W, \tag{5}$$

where W is now the width of the CFC gate film.

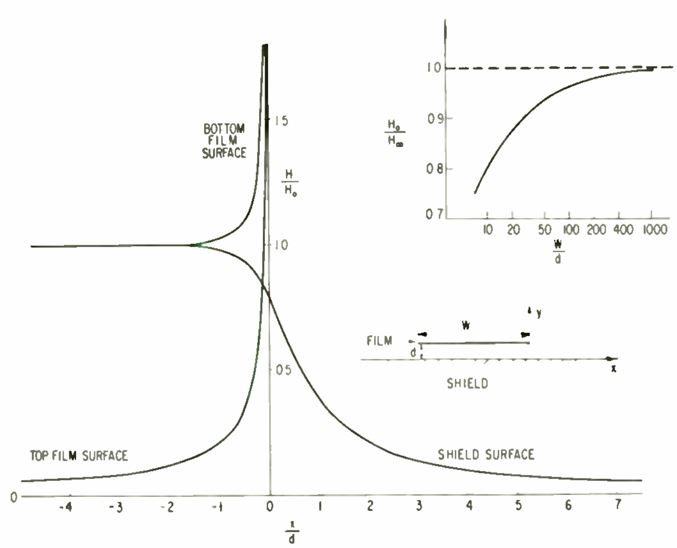


Fig. 8—Surface fields for a superconducting film of width W carrying a current I over an infinite superconducting shield plane. ($H_\infty = 0.4\pi\ I/W$.)

To calculate the field due to a current-carrying grid, we use the results of Appendix I as illustrated in Fig. 8. As mentioned above, a tin gate film will start to become resistive due to grid action when the surface field on it exceeds H_c. Before this happens, the surface field of the tin should be that of a completely superconducting plane with an adjacent current-carrying grid, *i.e.*, the middle curve of Fig. 8. The surface field on a superconducting tin gate film should, therefore, be a maximum in the middle of the grid and fall off slightly towards the edges. Under the center line of the grid, we find that the gate surface field to a very close approximation is given by

$$H = 0.4\pi I/w \tag{6}$$

where W is the grid width. If we now define H_c as the field at which the gate film reaches half its maximum resistance for low gate currents, then the grid current I_G at which the gate film resistance becomes $\frac{1}{2}R_{max}$ will be, using (6),

$$I_G = H_c w/0.4\pi. \tag{7}$$

In Fig. 4, a curve of H_c for a gate film measured directly with a solenoid is compared with values of H_c calculated from grid current measurements using (7). (Also shown are values of H calculated from the grid currents at which gate resistance begins to appear.) From the agreement between curves 1 and 2, it is clear that (7) is at least approximately correct.

If we now define the gain for the shielded CFC as $G=I_c/I_G$ and substitute for I_c and I_G from (5) and (7), we obtain

$$G = 0.4\pi \frac{j_c}{H_c} \frac{W}{w}. \tag{8}$$

We see that the gain is proportional to W/w, a function of the geometry, and to j_c/H_c, a function of the gate film properties. It is known that j_c/H_c decreases with gate film thickness and increases with $\Delta T = T_c - T$. (See, for instance, Bremer and Newhouse [7] and Crittenden [8].) The effect of temperature on the characteristics of a typical shielded CFC is shown in Fig. 9. There is, as predicted by (8), a strong increase of gain with decrease of temperature. The cryotrons described here are operated at $\Delta T = 0.08°$K, below their maximum gain, to reduce heat dissipation.

Operating Speed

The speed of a cryotron is, of course, dependent on the mode of operation. In the circuits described below [see Fig. 10(a)], the cryotron gate is in parallel with the load, which consists of the grid crossing a similar cryotron. The time constant τ at which current will be diverted from the gate XZ to the grid Y of the load cryotron is L/R where L is the load inductance and R the resistance of the driving cryotron. L is the sum of the grid inductance of the driven cryotron, the gate inductance of the driving cryotron, and the inductance of the connecting circuits.

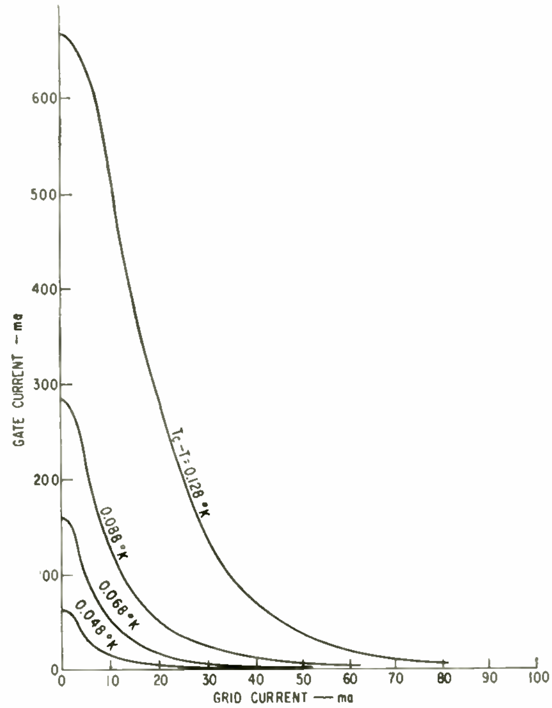


Fig. 9—Dependence of shielded CFC characteristics on temperature. Dimensions as in Fig. 3. T_c of gate film is 3.76°K.

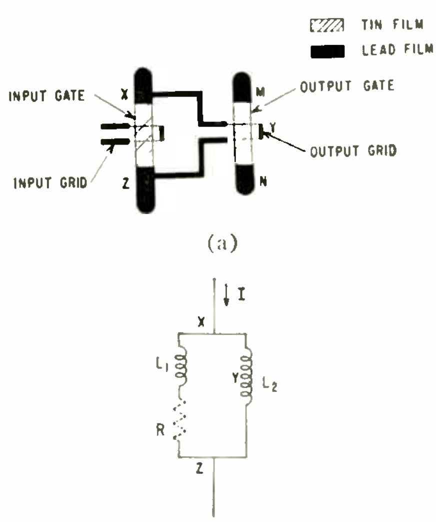


Fig. 10—(a) Cryotron storage cell. (b) Equivalent circuit.

It is shown in Appendix II that the inductance of a grid of width w spaced d cm from the superconducting shield plane is $4\pi \times 10^{-9} d/w$ henries/cm. The driven cryotron has a width W; hence, its grid has a length W and

$$L = 4\pi \frac{Wd}{w} \times 10^{-9} \text{ henries.} \tag{9}$$

(The gate of the driving cryotron and its connections to the driven cryotron are much wider than w, hence, their inductance can be neglected.) As discussed in connection with Fig. 7, the maximum portion of a shielded cryotron which becomes resistive is that part of the gate film covered by the grid. Hence, the resistance R of the driving cryotron of width W, energized by a current I_G through a grid of width w, is

$$R = \frac{1}{2} \frac{\rho w}{Wt} \text{ ohms,} \tag{10}$$

where t is the gate film thickness and ρ the effective bulk resistivity. (For very pure films, ρ is itself a function of t, but for the relatively impure films used here, this dependence can be neglected.)

From (9) and (10), we find that the effective time constant of one cryotron driving another is

$$\tau = L/R$$
$$= 8\pi \frac{td}{\rho} \left(\frac{W}{w}\right)^2 \times 10^{-9} \text{ second.} \tag{11}$$

Substituting from (8) for W/w in (11), we find the time constant in terms of the gain

$$\tau = 8\pi \frac{td}{\rho} \left(\frac{H_c}{0.4\pi j_c}\right)^2 G^2 \times 10^{-9} \text{ second.} \tag{12}$$

There are two points of interest in (12). First $\tau \propto t(H_c/0.4\pi j_c)^2$. This shows that there is an optimum value of the gate thickness t because, as we attempt to reduce τ by reducing t, $H_c/0.4\pi j_c$ increases. For solid wires, $H_c/0.4\pi j_c \rightarrow 1$, but for the 0.3-micron tin films presently used, $H_c/0.4\pi j_c$ is between 20 and 50. The second point of interest is that τ is not a function of the grid and gate widths. Hence, a reduction in cryotron area will not increase speed.

Present values for the material constants in (12) are as follows: $t=0.3$ micron, $d=1.0$ micron, $\rho=6\text{–}12\times 10^{-7}$ ohm cm, $H_c/0.4\pi j_c=20\text{–}50$. A practical value for the gain is 2. Substituting these values into (12), we obtain a theoretical range of $\tau=10\text{–}130\times 10^{-8}$ second. A typical cryotron, described below, has an experimental time constant of 38×10^{-8} second at the temperature of operation close to T_c. Fig. 9 shows that higher gain or speed is possible by operating at lower temperatures.

A Simple Storage Circuit

We will now describe a simple storage circuit used to measure τ which depends on a principle unique to superconducting networks. The principle will be illustrated with the circuit shown in Fig. 10(a).

In one mode of operation a current is applied between X and Z. Most of this flows through the path XZ rather than XYZ because the former has much lower inductance. The equivalent circuit is shown in Fig. 10(b).

If current is now passed through the input grid, XZ becomes resistive and I is diverted through the path XYZ. It is now possible to switch off the current through the input grid so that XZ becomes superconducting again. Since L_2 is still superconducting, it is to be expected that I will remain diverted through L_2, even though L_1 has become superconducting again. This does, in fact, happen experimentally. The current in XYZ can be determined conveniently with a dc measurement by measuring the resistance of MN.

If, after I has been diverted to L_2 and after L_1 has become superconducting again, I is switched off, a circulating current will remain in loop XYZ. Its magnitude can be calculated as follows:

Assume that a current $+I$ has been injected into node X and completely diverted to L_2, through L_1 having been temporarily made resistive. If now current $-I$ is injected into node X, it will divide itself between XZ and XYZ in the inverse ratio of their inductances. The current along XZ will be $I_{XZ}=-I(L_2/L_1+L_2)$ and the net current along XYZ will be

$$I_{XYZ} = I - I\frac{L_1}{L_1 + L_2}$$

$$= I\frac{L_2}{L_1 + L_2}.$$

Therefore, $-I_{XZ}=I_{XYZ}$. Hence, the circulating current is

$$I_{\text{circ}} = \frac{L_2}{L_1 + L_2} I. \tag{13}$$

Eq. (13) has been confirmed experimentally and currents have been stored for several hours.

The results which have been described in connection with the above storage circuit can be generalized. If a current is injected into a network of superconducting elements, it will distribute itself among them in inverse ratio to their inductances. By making one or more of these inductors resistive for various lengths of time, current will be diverted from the resistive elements to the superconducting ones. As soon as the resistive elements are made superconducting again, however, the current distribution stops changing, as long as the external injected current remains constant.

It is clear that in a cryotron computer, analog and digital storage will be simpler than in a transistor computer, where positive feedback circuits, or magnetic cores which cannot conveniently supply a continuous output, are required.

The circuit of Fig. 10 provides a convenient way of measuring the effective speed of the cryotrons used in it. The deflection of current from leg XZ to leg XYZ takes place with a time constant

$$\tau = (L_1 + L_2)/R. \tag{14}$$

(R is the resistance produced in XZ by the input grid current.) By applying individual pulses of known length to the input grid, we can measure the current changes in the output grid and obtain an experimental value for τ.

The results of applying 0.05-μsec current pulses to the input grid of one of the storage loops shown in Fig. 12 is shown in Fig. 11. The experimental curves have a time constant of approximately 0.38 μsec. The value given by (14), using calculated values for L_1+L_2 and experimental values for R, is 0.33 μsec. The curve of

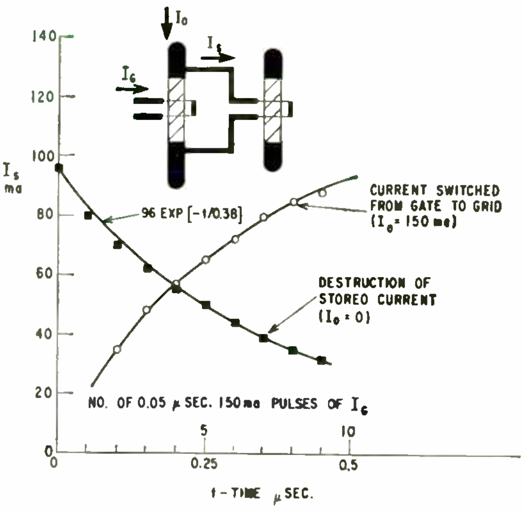


Fig. 11—Change of stored current due to 0.05 μsec pulses on input grid of a storage cell of Fig. 7. Curves are fitted to the data points corresponding to a time constant of 0.38 μsec.

rising current is obtained when 150 ma is injected into the cryotron loop and gradually diverted to the output leg. The curve of decreasing current corresponds to a stored current (with the external current switched off) being destroyed by pulses on the input grid.

A Shift Register

The memory circuit described has been applied to a shift register. A short portion of such a register is shown in Fig. 12. The register can be operated with three stages per bit using three advance and three reset current sources. Alternatively, it can be operated with four stages per bit, using two advance and two reset current sources. A diagram and a set of calculated waveforms for three-stage-per-bit operation are shown in Fig. 13.

Information travels from left to right. To inject a "1" into the register, the input winding is pulsed while advance current I_1 is on. This diverts I_1 from the first cryotron to its output grid, and, when I_1 goes off, a circulating current C_1 remains in the first storage cell. I_2 is now injected into the second storage cell. Due to the existence of C_1, I_2 will be diverted to the output grid of the second storage cell. It is necessary at this time to destroy C_1. This is done by passing current through the reset winding R_2. C_1 has to be destroyed before I_2 is switched off, since otherwise C_2, the circulating current in the second storage cell caused by the effect of C_1 on I_2, would be destroyed. After C_1 has been destroyed and I_2 switched off, the injected "1" is represented by the circulating current C_2 in cell 2. In similar fashion, C_3 is created and C_2 is destroyed. Only now can a new "1" be injected into the first storage cell. The grid of the last storage cell crosses an output cryotron whose resistance is an indication of the presence or absence of the circulating current C_4. As described earlier, the experimental and calculated time constant of the cryotron in this circuit module is 0.38 μsec. This time constant could probably be reduced to approximately 0.1 μsec by working at a lower temperature and changing the cryotron dimensions to lower the gain. The register has been used

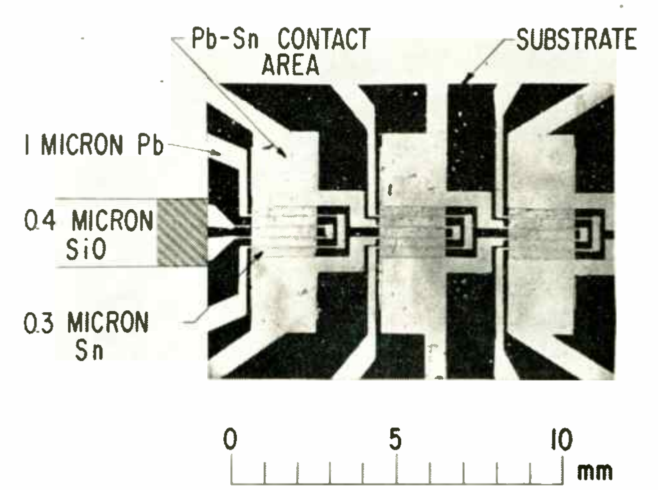


Fig. 12—Portion of experimental shift register. Note the insulator film separating the grids from the underlying gate films.

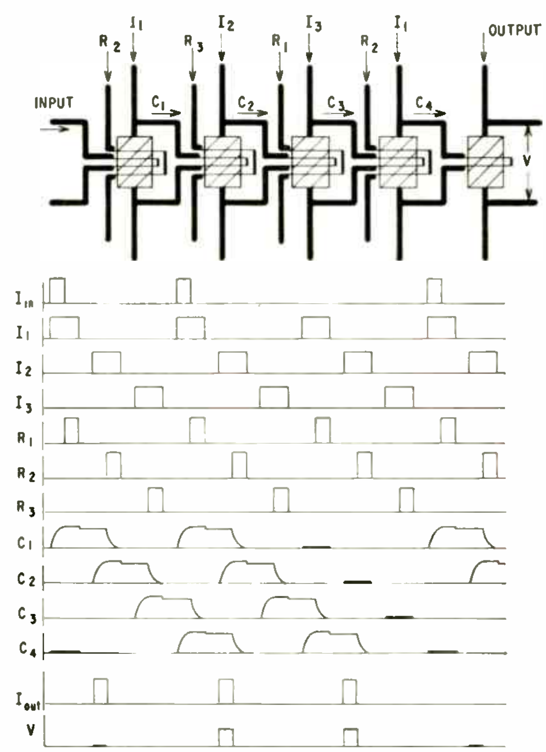


Fig. 13—Shift register schematic and calculated waveforms.

to transfer information but so far only at low repetition rates.[2]

System Considerations

A summary of present CFC characteristics is shown in Table I.

TABLE I

Time constant	0.4 μsec
Size	6 mm^2
Dissipation (dc)	$=5\times10^{-6}$ watt

[2] A 4-stage-per-bit shift register has recently been operated at a repetition rate of over 100 kc.

The lower limit in size is presently set by difficulties in getting good contact between the CFC gate and its connecting circuitry.

Using the present elements at the packing density of the shift register of Fig. 12 (this leaves enough free area for interconnections) would make it possible to accommodate 18,000 active elements on a one-foot^2 plate. This makes the assumption that every gate element is crossed by an average of four active elements. (This assumption is realistic since the average amplifier in a digital computer is fed by a logic element involving at least four diodes.) Using an evaporator which is large enough to handle a one-foot^2 plate, and using vacuum mask changing equipment, it should be possible to manufacture one 18,000 element plate per hour. Stacking two of these plates per cm gives a one million element computer in one cubic foot. Assuming each element to be on for 10 per cent of the time produces a total dissipation of $1\times10^6\times5\times10^{-7}=0.5$ watt. In addition, approximately 0.25 watt heat enters the liquid helium due to radiation and conduction through the container. It has been calculated that 100 input and output leads, 10 of which carry 200-ma dc, would contribute less than 0.2 watt in conduction and Joule heating (provided that the upper ends of these leads are cooled by liquid nitrogen). The total heat inflow is therefore approximately one watt for the system under discussion. Helium refrigerators with one-watt capacity for a one-cubic-foot volume are presently being designed in at least two laboratories.

The above considerations show that CFC are today the only active components which together with their interconnections can be manufactured and assembled in a few simple steps. This makes the cost of component manufacture and assembly negligible compared to the cost of design and testing.

Film cryotrons also constitute the first advance over transistors with respect to computer miniaturization because, although semiconductor components have been developed in the last few years which have very small size, large numbers of them cannot be used in a small volume because their dissipation is too high. CFC do not have this limitation.

As shown above, and in other ways, CFC can be used for storage. They have the advantage over magnetic cores of being compatible with the logic circuitry without the need for intermediate amplification. This makes it possible to integrate memory and logic in new ways, leading to vastly improved performance. When cryotrons are used for storage, they have of course no dissipation—a considerable advantage over all proposed semiconductor storage elements.

To summarize, CFC should make possible computers which have an order of magnitude more components than hand-assembled machines—and are correspondingly more powerful. At the same time, they may reduce the cost and size of conventional computers by an order of magnitude.

Appendix I

Surface Current Density in a Shielded Superconducting Strip of Zero Thickness

When an external magnetic field is applied to a superconductor, currents are induced in its surface sufficient to keep the internal field zero. It is easily shown that the surface current density under these conditions is given by

$$g = \frac{1}{0.4\pi} H \times n \text{ amperes/cm} \tag{15}$$

where H is the magnetic field at the surface, and, necessarily, parallel to it, and n is a unit vector normal to the surface.

In a current-free region, H can be written as the gradient of a scalar function ψ, the magnetostatic potential:

$$H = \nabla\psi \text{ oersteds.} \tag{16}$$

and, combining with (15) at the boundary of the current-free region,

$$g = \frac{1}{0.4\pi} \nabla\psi \text{ amperes/cm.} \tag{17}$$

Since the divergence of H is zero in the region of interest,

$$\nabla^2\psi = 0. \tag{18}$$

It is noticed that solving for g is the same mathematical problem as finding the surface charge on a conductor in an electrostatic field, since

$$\sigma = \frac{1}{4\pi} D = \frac{1}{4\pi\epsilon} \nabla V$$

and $\nabla^2 V = 0$, where V is the electrostatic potential, ϵ is the dielectric constant, and D is the electric flux density. Since Laplace's equation has been solved for the appropriate geometry in the electrostatic case [9], this result can be applied directly to the problem of current density.

The essential steps of the problem are summarized. If all the current is perpendicular to the cross section of the strip, then the problem becomes two dimensional and conformal mapping techniques can be used.

The situation to be analyzed is shown in Fig. 14(a). We use the approximation of an infinitely thin film. (The case of a finite thickness film has been solved, also, but the results were not sufficiently different to warrant presentation.)

To solve the case shown in the insert of Fig. 9, we use the further approximation of allowing the left-hand side of the strip to go to infinity, as shown in Fig. 14(a). This is justified if $d/w \ll 1$. To determine the magnetic field around the current-carrying strip, we define

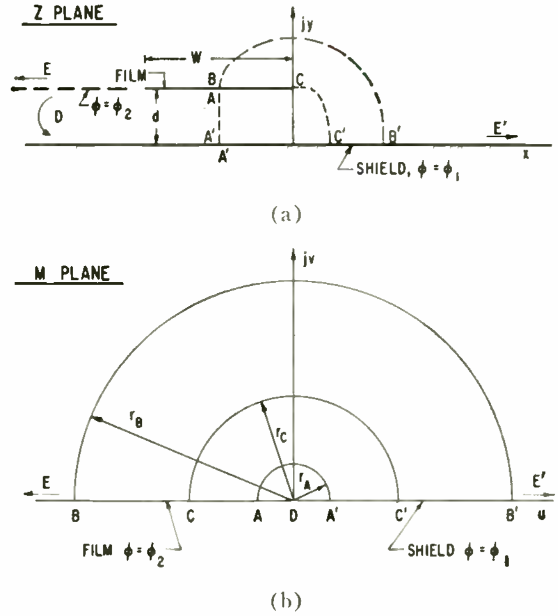


Fig. 14—Transforms used in Appendix I.

the function

$$F(z) = \phi(x, y) + j\psi(x, y) \tag{19}$$

where

$$z = x + jy. \tag{20}$$

ϕ is the axial component of vector potential analogous to the electrostatic potential, and ψ is the magnetostatic potential analogous to the electrostatic flux lines.

We use the Schwartz-Christoffel transformation shown in Fig. 14(b) to the m plane. (m is used instead of the more familiar w to avoid confusion with the grid width.) In the m plane,

$$\begin{aligned} F(m) &= \phi(u, v) + j\psi(u, v). \\ m &= u + jv, \text{ and} \\ m &= f(z). \end{aligned} \tag{21}$$

In Fig. 14(a) and 14(b), point A is the midpoint of the bottom of the film, point B is midpoint of the top of the film, and point C is the edge of the film. r_c is arbitrarily chosen to equal 1 since arbitrary selection of one scale factor is allowed in Fig. 14(b). The transformation required to go from Fig. 14(a) to 14(b) is

$$\frac{\pi z}{d} = 1 + m + \ln m. \tag{22}$$

The solution for F in the m plane is known to be

$$\begin{aligned} F(m) &= \phi(m) + j\psi(m) \\ &= -j\frac{\phi_2 - \phi_1}{\pi} \ln m - \phi_1. \end{aligned} \tag{23}$$

It is not necessary to know the value of ϕ_2 and ϕ_1 to find g. From (23),

$$\psi(m) = -\frac{\phi_2 - \phi_1}{\pi} \ln m = k_1 \ln m. \tag{24}$$

From (17), for this two-dimensional case,

$$g = K_2 \left| \nabla \psi \right|_s = K_2 \left. \frac{\partial \psi}{\partial x} \right|_s = K_2 \left(\frac{\partial \psi}{\partial r} \right)_s \left(\frac{dr}{dx} \right)_s, \tag{25}$$

where s indicates the derivative is evaluated on the film surface.

K_2 is evaluated using the fact that one-half the total current must flow between points A and B, the midpoint of the bottom and top of the film respectively. This total current is directly proportional to the change in ψ going from A to B.

Evaluating K_2 and the derivative of (25),

$$g_s = \frac{I}{W} \cdot \frac{1 - r_B + \ln r_B}{r_A - r_B} \cdot \frac{1}{1 - r}, \tag{26}$$

where r is a function of X defined by (22).

At the film surface, $(m = re^{j\pi} = -r)$ and

$$\frac{\pi x}{d} = 1 - r + \ln r,$$

$$\frac{\pi y}{d} = \pi. \tag{27}$$

At the shield surface, $\theta = 0$ and

$$g_s = \frac{I}{W} \frac{1 + r_B + \ln r_B}{r_B - r_A} \cdot \frac{1}{1 + r}. \tag{28}$$

From (26) and (27), it is seen that as W/d increases, g of the top surface $(r > 1)$ will approach 0 while $g/(I/W)$ for the bottom surface $(r < 1)$ approaches 1. W/d will vary considerably for both grids and gates. A value of 50 to 100 is not unknown for a control grid. For a gate W/d is of the order of 5000. The plot of the magnetic fields is shown in Fig. 8. H/H_0 is plotted vs x/d and is essentially correct for any value of $W/d \geq 10$. H_0 is the value of the field at the center of the bottom film surface. The variation of H_0 with W/d is shown in the insert. H_∞ is the field value at the center of the bottom film surface for $W/d = \infty$. Since g is directly proportional to H at the surface, Fig. 8 also shows the variation of g with x.

It is seen that the field is essentially uniform between the film and the shield. H_0 is evaluated from (15) and (26) (for $r \sim 0$):

$$H_0 = 0.4\pi \frac{I}{W} \frac{1 - r_B + \ln r_B}{r_A - r_B}. \tag{29}$$

As $W/d \rightarrow \infty$, $r_B \rightarrow \infty$, $r_A \rightarrow 0$, and $1 + \ln r_B \ll r_B$. Thus,

$$H_\infty = 0.4\pi \frac{I}{W}. \tag{30}$$

In practice, the film is not infinitely thin, and the current density cannot be infinite at the edge of the film as (26) states, or that portion of the film would go resistive. This fact can only serve to make the current distribution more uniform than shown.

Appendix II

Inductance of a Superconducting Strip over a Shield Plane

If I amperes are flowing in a conductor of inductance L henries, the energy in the magnetic field is

$$U = \tfrac{1}{2} LI^2 \times 10^{-7} \text{ ergs}, \tag{31}$$

or, in terms of the volume integral of the magnetic field H,

$$U = \frac{1}{8\pi} \int H^2 dV \text{ ergs}. \tag{32}$$

L is calculated by evaluating (31) and equating it to (32)

To evaluate (32), it is assumed that the field has a uniform value H_0 in the volume between the strip and the shield and zero elsewhere. This is justified in Appendix I.

From Appendix I,

$$H = 0.4\pi \frac{I}{w} \text{ oersteds}. \tag{33}$$

Combining (31), (32), (33), and solving for L,

$$L = 4\pi \frac{d}{w} (10)^{-9} \text{ henries/cm}. \tag{34}$$

Effects of the finite penetration depth of the field and current into the film have been neglected. These will tend to make L slightly larger than (34).

Acknowledgment

The authors are happy to acknowledge the contributions of D. MacKellar and H. Tanis to the device technology and to the experimental work in general.

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Gallium Arsenide Tunnel Diodes*

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Summary—**The fabrication and properties of GaAs tunnel diodes are described. The material preparation is discussed; devices are described which have been fabricated consistently with peak to valley current ratios >15:1, with voltage swings in the range from 0.9 to 1.2, and with current densities from 2000 amp/cm² to over 10,000 amp/cm² (and with correspondingly low capacitances, *e.g.*, capacitances as low as 0.2 $\mu\mu f$/ma and g/C figures of merit as high as 5×10^{10} sec^{-1}). The temperature behavior of typical units is presented. Applications particularly well suited to GaAs units are mentioned.**

Introduction

In a relatively short time the tunnel diode[1–4] has become a well-known device both as a negative resistance (active) circuit element and as a research tool in the study of degenerate semiconductors.[5–8] Initially, most tunnel diode studies were based around the two well-known materials, germanium and silicon. Almost simultaneously with this work, however, it was recognized that intermetallic compounds offered many attractive features as tunnel diode materials. This is because tunneling and tunnel diodes are not dependent upon minority carrier lifetime, which has never been very high in the intermetallics, and because such properties as low effective masses and high mobilities, which are characteristic of many of the intermetallic compounds, are of direct benefit to tunnel diode operation.

Among the various intermetallic compounds that have exhibited tunneling action or from which successful tunnel diodes have been fabricated, GaAs is one of the most attractive. First of all, the wide energy gap of GaAs (1.35–1.45 ev) makes possible tunnel diodes with a voltage swing[9] significantly larger than those made of germanium or silicon. This is desired in many switching applications. The theoretical decrease in tunneling current due to a wider band gap is more than offset by the low effective masses ($<0.1m_e$) of GaAs relative to those of germanium and silicon ($>0.1m_e$).[10] Also, the larger mobilities of carriers in GaAs, again relative to germanium and silicon, make possible a reduction in device series resistance. Finally, the lower dielectric constant (11.1) of GaAs is of some value in reducing the junction capacitance.

This paper describes the fabrication and properties of GaAs tunnel diodes. The material preparation is discussed; devices are described which have consistently been fabricated with peak to valley current ratios >15:1, voltage swings from 0.9 to 1.2, and current densities as high as 10,000 amp/cm² and higher (and correspondingly low capacitances, *e.g.*, capacitances as low as 0.2 $\mu\mu f$/ma and g/C figures of merit as high as 5×10^{10} sec^{-1}). The temperature behavior of typical units is presented. Applications particularly well suited to GaAs units are mentioned.

Device Fabrication

The devices described herein were in all cases fabricated from polycrystalline starting material of unknown properties; *i.e.*, the degree of stoichiometry and purity were either unknown or known to be not particularly good. Gallium arsenide crystals were obtained from six different (independent) suppliers and cut into wafers and etched in preparation for doping by diffusion. An arbitrary number of thin wafers were sealed into an approximately 10 ml quartz ampoule with either zinc or cadmium (acceptors) ranging in quantity in each case from a fraction of 1 mg to over 10 mg. Also, from 5 to over 30 mg of free arsenic was enclosed in the ampoule, and the system, after evacuation and sealing, was annealed at temperatures ranging from 800°C to over 1100°C. The free arsenic in each ampoule established a sufficiently high arsenic vapor pressure to prevent decomposition of the GaAs wafers. For the conditions described no deterioration of the wafer surfaces was noticed.

With cadmium as a doping agent, measured doping levels of over 2×10^{19} atoms/cm³ could be attained; with zinc, measured doping levels as high as 2×10^{20} atoms/cm³ could be obtained. The resistivities and hole mobilities corresponding to these doping levels were approximated as 0.007 ohm-cm and 40 cm²/volt-seconds, and 0.001 ohm-cm and 40 cm²/volt-seconds, respec-

* Received by the IRE, February 29, 1960; revised manuscript received, April 22, 1960.

† General Electric Co., Electronics Park, Syracuse, N. Y.

[1] L. Esaki, "New phenomenon in narrow Ge *p-n* junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January, 1958.

[2] H. S. Sommers, Jr., "Tunnel diodes as high-frequency devices," Proc. IRE, vol. 47, pp. 1201–1206; July, 1959.

[3] I. A. Lesk, N. Holonyak, Jr., U. S. Davidsohn, and M. W. Aarons, "Germanium and silicon tunnel diodes—design, operation and application," 1959 IRE WESCON Convention Record, pt. 3, pp. 9–31.

[4] R. N. Hall, "Tunnel Diodes," presented at IRE-PGED Meeting, Washington, D. C.; October 29, 1959.

[5] N. Holonyak, Jr., I. A. Lesk, R. N. Hall, J. J. Tiemann, and H. Ehrenreich, "Direct observations of phonons during tunneling in narrow junction devices," *Phys. Rev. Letters*, vol. 3, pp. 167–168; August, 1959.

[6] J. V. Morgan and E. O. Kane, "Observation of direct tunneling in germanium," *Phys. Rev. Letters*, vol. 3, pp. 466–468; November, 1959.

[7] J. I. Pankove, "Influence of degeneracy on recombination radiation in germanium," *Phys. Rev. Letters*, vol. 4, pp. 20–21; January, 1960.

[8] S. L. Miller, M. I. Nathan, and A. C. Smith, "Pressure dependence of the current-voltage characteristics of Esaki diodes," *Phys. Rev. Letters*, vol. 4, pp. 60–62; January, 1960.

[9] The voltage from tunnel region to thermal region of the V-I characteristic.

[10] N. B. Hannay, "Semiconductors," Reinhold Publishing Co., New York, N. Y., ch. 9, p. 415 and ch. 8, p. 323; 1959.

tively. These measurements were obtained on selected polycrystalline wafers in which individual crystallites were large, *i.e.*, of a size comparable to the wafer dimensions.

In fabricating the diode, a small scribed wafer is alloyed in a hydrogen ambient to a header assembly or base plate with gold-zinc alloy or with tin or lead-based alloys doped with zinc. On the opposite side of the wafer a small quantity of an alloy, largely composed of tin or lead, doped with sulphur, germanium, or tellurium is alloyed onto the parent crystal (either an individual crystallite or wafer consisting of several crystallites).[11] The parent crystal is p-type as a result of zinc or cadmium doping, and the alloyed-regrown region is n-type due to sulphur, germanium, or tellurium doping.

Fig. 1 shows a cross section of an unetched GaAs tunnel diode. The lower portion of the figure shows quite clearly the thick, uniform n-type regrown region formed on the p-type parent crystal. The upper portion of the figure shows that the regrown region is uniform under the entire alloy material. Structures such as those shown in Fig. 1 have been completely assembled, and etched to the desired peak tunnel currents, in low-inductance rectangular and circular strip line packages and in relatively low-inductance, single-ended header assemblies.

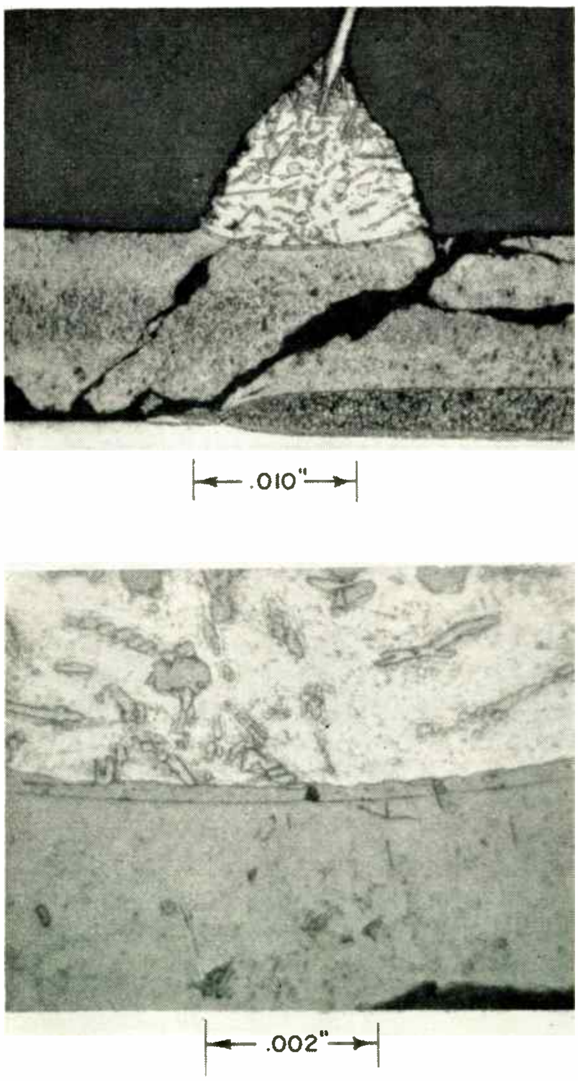

Fig. 1—Unetched gallium arsenide tunnel diode, magnified cross sections. GaAs wafer p-type (doped with Zn). Alloyed-regrown region n-type (doped with S).

Electrical Properties

The general appearance of a GaAs tunnel diode V-I characteristic as compared to that of Ge and Si is shown in Fig. 2. In terms of the magnitude of the peak to valley current ratio, the Ge characteristic is that of a "good" unit, the Si characteristic is that of a "fair" unit, and the GaAs characteristic is that of a "poor" unit. In spite of the fact that the GaAs unit has been classified as "poor," it will be noticed that it has greater than a 10:1 peak to valley current ratio and a voltage swing ~ 1 from peak tunnel current to equal thermal current. A more typical GaAs tunnel diode V-I characteristic is shown in Fig. 3. Again the voltage swing is ~ 1, and the peak to valley current ratio is greater than 20:1. Soltys and Hall[12] have managed to obtain ratios as high as 60:1 by procedures similar to those described here. In the present work typical ratios, at current densities ranging from 2000 amp/cm^2 to 10,000 amp/cm^2, are in the range from 15:1 to over 30:1. For such units the voltages corresponding to the peak tunnel currents fall in the range from 0.1 to 0.25. Finally, voltage swings greater than 1.1 are frequently observed.

Diodes fabricated on cadmium-doped material consistently have current densities in the range from 2000 to 10,000 amp/cm^2. Higher current density units can be obtained on zinc-doped material, but then mechanical problems are experienced in attempting to reduce the junction area sufficiently to give convenient magnitudes of peak tunnel current.[13] The high tunnel current densities attainable in GaAs make possible tunnel diodes having measured capacitances as low as 0.2 $\mu\mu f$ per milliampere of peak tunnel current, and perhaps lower. This is of considerable importance to the theoretically maximum frequency of oscillation

$$\omega_c = g/C\sqrt{(1/gR_s) - 1}$$

of a tunnel diode. In the expression given above, g refers to the absolute value of the negative conductance, C to the junction capacitance, and R_s to the series resistance of the tunnel diode. The ratio g/C is the important

[11] After etching, the junction area is of the order of 0.001 inch diameter or less. Hence, if one uses GaAs of reasonable crystallite sizes, no problems are encountered in using polycrystalline material.

[12] T. Soltys and R. N. Hall, private communication.

[13] Units of this type display their negative resistance regions for only an instant in switching to the thermal region (higher voltages) and then burn out because of the high currents and small areas of the junctions. These effects have thus far made it impossible to make actual current density measurements greater than 10,000 amp/cm^2.

quantity commonly referred to and has been quoted as being as high as 2×10^{10} sec^{-1} in germanium.[2] In practice it is found that it is extremely difficult to obtain g/C values of this magnitude in germanium, whereas values as high as 5×10^{10} sec^{-1} along with high (20:1+) peak to valley current ratios are readily attained in GaAs.

The series resistance R_s of GaAs tunnel diodes of 5 ma peak tunnel currents has been measured and found to fall in the range from 2 to 10 ohms. These values are considered high in terms of what can be achieved by simple mechanical improvements in the assemblies which have been fabricated to date or by developing units in which the parent wafer is n-type, rather than p-type, so that the greater mobility of n-type crystals can be used to better advantage.

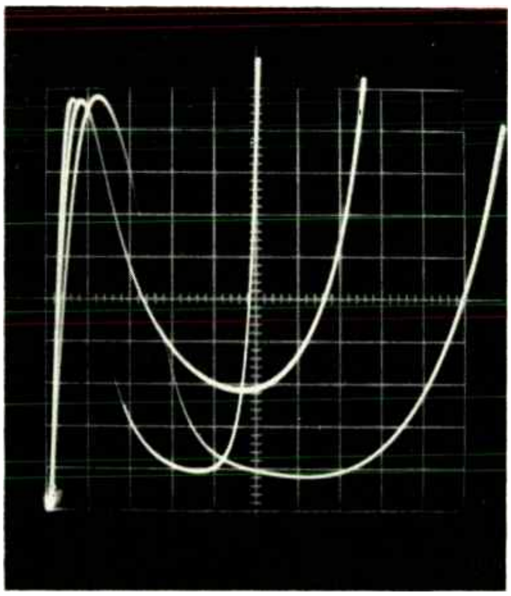

Fig. 2—V-I characteristics of germanium, silicon, and gallium arsenide tunnel diodes. The doping impurities of each diode are Ga, As; B, As; and Cd, S, respectively. Scale: 0.1 volt/div (horizontal), 0.2 ma/div (vertical).

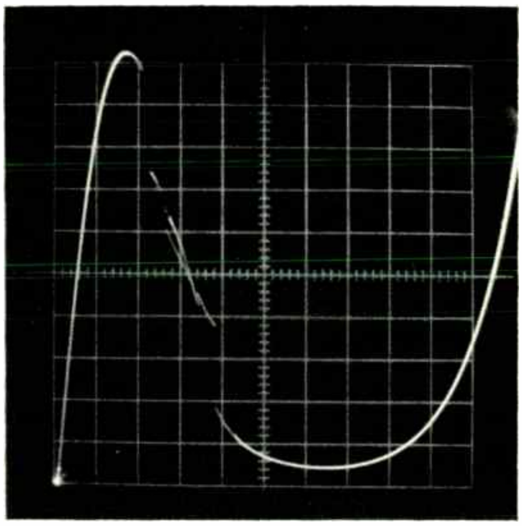

Fig. 3—Gallium arsenide tunnel diode V-I characteristic. Peak to valley current ratio 25:1. Parent crystal cadmium doped, alloyed-regrown region sulphur doped. Scale: 0.1 volt/div (horizontal), 0.5 ma/div (vertical).

The speed properties of GaAs tunnel diodes can best be attested to by the fact that units of relatively high peak tunnel currents have been operated as oscillators well into the kilomegacycle frequency range. A diode of peak current a little less than 10 ma has been inserted into one end of a strip line which nearby was capacitively shorted and presented an inductive load across the tunnel diode. With dc bias applied from a low impedance source to the strip line and tunnel diode, oscillations were obtained at 4.4 kmc. A representative unit, mounted on a single-ended header of considerable inductance, was similarly arranged in a strip line and was observed to oscillate at 2 kmc. Because of the crude nature of these circuits and experimental diodes, it is reasonable to predict that GaAs tunnel diodes will be made to oscillate well above 10 kmc. By very ingenious experimental procedures Ge tunnel diodes have been made to oscillate at 10 kmc.[4] Similar techniques should enable GaAs tunnel diodes to surpass Ge.

Temperature Properties

The temperature dependence of GaAs tunnel diode V-I characteristics has been found to resemble in many respects those previously reported in Ge and Si., Fig. 4 shows the change in peak current and valley current of a typical unit in the temperature range from that of liquid nitrogen to 200°C. The valley current, as shown, increases monotonically with temperature. Beyond 200°C the rate of increase in valley current with temperature becomes steeper. Although the data do not go beyond 200°C, some units have been operated up to 300°C with a peak to valley current ratio slightly greater than 3:1. As shown, the peak current decreases by roughly 20 per cent when the temperature is lowered from room temperature to that of liquid nitrogen. Slightly beyond room temperature the peak current decreases monotonically with increasing temperature. This behavior is expected because of the increase in energy gap (and decrease in tunneling) as the temperature is lowered and because of the smearing out of the Fermi function at higher temperatures. In all units examined, the maximum in peak current occurred at or near room temperature.

At liquid nitrogen temperatures (77°K) and below, many GaAs tunnel diodes exhibit a V-I characteristic much as shown in Fig. 5. In the region where the current increases from its valley magnitude (4 ma, 0.7 v) to some higher magnitude (34 ma, 1.2 v) in the thermal region, a "hump" is noticed in the V-I characteristic. A similar effect has been observed in Si tunnel diodes fabricated by alloying boron-doped aluminum onto a substrate heavily doped ($\sim10^{20}$ atoms/cm^3) with phosphorus but has not been observed in silicon doped with arsenic. This effect was first observed in germanium by

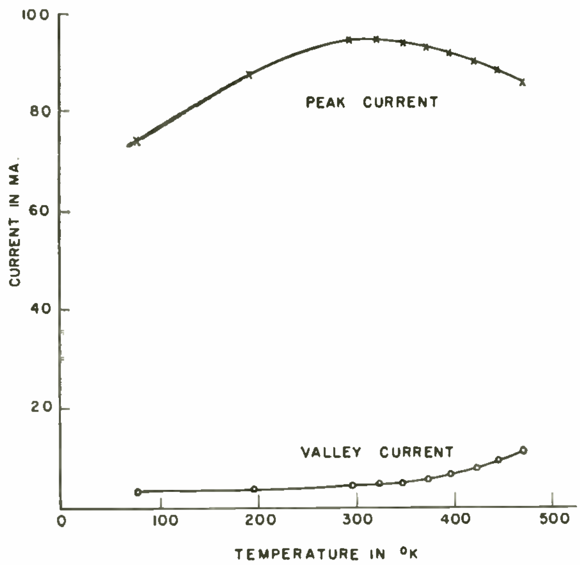

Fig. 4—Gallium arsenide tunnel diode peak current and valley current temperature dependence.

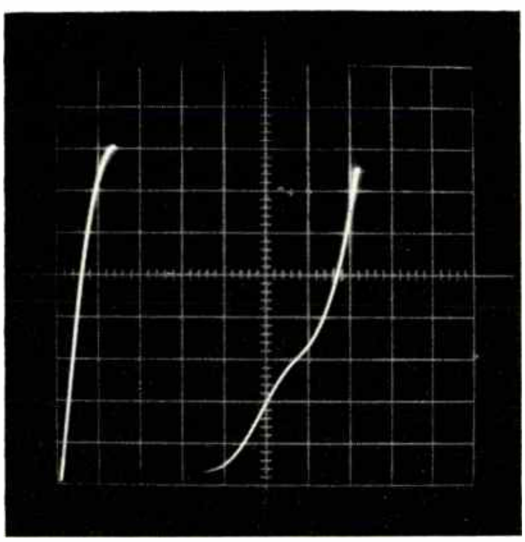

Fig. 5—Gallium arsenide tunnel diode V-I characteristic at 77°K. Parent crystal cadmium doped, alloyed-regrown region sulphur doped. Scale: 0.2 volt/div (horizontal), 10.0 ma/div (vertical).

Esaki[14] and later was described in greater detail by Longo,[15] who attributes it to tunneling to states formed in the forbidden region on one side of the junction. This has been predicted by Ehrenreich[16] and possibly others. In the case of GaAs it is not certain at this time whether the type of chemical doping element is responsible for this effect and/or whether it is due to some type of structural or strain property (*e.g.*, interstitial impurities) associated with high doping levels and particular impurities.

Applications

As is almost directly obvious from the work previously performed with Ge and Si, GaAs tunnel diodes can be used, with minor changes in circuitry, as direct replacements for Ge and Si units. In addition, many applications are possible which formerly could not be considered possible or practical for Ge or Si. For example, the large voltage swings and high peak to valley current ratios of GaAs tunnel diodes make them eminently well suited as threshold elements to switch germanium transistors, and silicon transistors and controlled rectifiers.[17] In such applications the GaAs tunnel diode acts as a shunt to the input signal until it attains a magnitude equal to the peak current. At that point the tunnel diode switches (rapidly) to its valley region and presents the transistor or controlled rectifier with the requisite current and voltage drive.

The wider voltage swing of GaAs tunnel diodes suggests that it may be possible to build logic circuits in which germanium rectifier diodes may be used to steer the logic functions in a prescribed direction. Preliminary work by Chow[18] indicates that this is indeed true.

One of the newest and perhaps most novel applications of GaAs tunnel diodes could be to build a coupled series pair of GaAs and Ge tunnel diodes where the GaAs unit has a peak current slightly higher than that of the Ge unit and where the valley current of the GaAs unit is smaller than that of the Ge unit. The over-all V-I characteristic of such a series pair turns out to be a multivalued function of both current and voltage.[19] It has been found possible, under suitable conditions, to increase the current through the series pair and cause first the Ge unit to turn on, then to turn off just as the GaAs unit turns on, and finally cause both units to turn on, in the sequence outlined. This type of behavior suggests further possibilities for performing computer and logic operations. Gallium arsenide tunnel diodes are perfectly suited for this application since their peak to valley current ratios are generally much greater than those of Ge.

Another application which is highly suited to GaAs is that of building a higher power oscillator making use of an active compound element consisting of a series connection of an odd number of matched tunnel diodes. This idea is due to Kim[20] and has been initially explored with Ge tunnel diodes.

Another area in which GaAs tunnel diodes (Ge and Si units also) may find considerable application is in the

[14] L. Esaki, "Properties of heavily-doped germanium and narrow *p-n* junctions," in "Solid State Physics," *Proc. Internatl. Conf., Brussels; June, 1958*, Academic Press, London, Eng., and New York, N. Y., vol. 1, pt. 1, p. 519; 1960.

[15] T. A. Longo, "On the nature of the maximum and minimum currents in germanium tunnel diodes," *Bull. Am. Phys. Soc.*, ser. 2, vol. 5, p. 160; March, 1960.

[16] H. Ehrenreich, private communication.

[17] These applications were first explored by T. P. Sylvan.

[18] W. F. Chow, private communication.

[19] This effect was first examined by C. O. Harbourt and R. P. Nanavati, Syracuse University, to be published.

[20] C. S. Kim, private communication, and J. J. Stoker, "Nonlinear Vibrations," Interscience Publishers, Inc., New York, N. Y., ch. 5, p. 133; 1950.

performance of electronic functions in ambients of high nuclear radiation intensity. Experiments are in progress in many laboratories to determine tunnel diode radiation tolerances. (These experiments are also yielding information concerning effects of band structure on the V-I characteristic and vice versa.) Tunnel diodes should be quite insensitive to gamma radiation (and light) since their characteristics display such high dc conductances everywhere, and because minority carrier lifetimes can be (and generally are) so small. Electron bombardment in the mev energy range of sufficient intensity causes an increase in the valley current[15] in much the same way as does heating. Fast neutrons also cause a permanent effect like that of heating, *i.e.*, an increase in valley current and a decrease in voltage swing. At sufficiently high integrated neutron flux levels, the negative resistance disappears. Germanium, silicon, and gallium arsenide tunnel diodes show noticeable changes in their V-I characteristics at fast neutron doses of 10^{15}–10^{16} nvt. By 10^{17} nvt, germanium and silicon units are inoperable while gallium arsenide units still show appreciable negative resistances.[21]

[21] F. J. Reid, L. W. Aukerman, and R. D. Baxter, Battelle Memorial Institute, private communication, to be published.

Conclusion

Because of the remarkably good electrical properties of GaAs tunnel diodes, it is obvious that they hold many advantages over their Ge and Si counterparts, particularly in the greater number of applications for which GaAs tunnel diodes are better suited. Their speed properties thus far are as good and better than those of Ge units. Their ability to operate over a wide temperature range and to resist radiation damage make them competitive in these respects also. The fact that relatively poor quality GaAs can be fabricated into high quality tunnel diodes by simple processes makes GaAs more than competitive with most known semiconductor materials which have been used or considered for use in making tunnel diodes.

Acknowledgment

The authors are indebted to U. S. Davidsohn, Y. C. Hwang, R. E. Hysell, and J. Sparks for performing some of the measurements reported. They would like to acknowledge especially the assistance of S. F. Bevacqua throughout this work, and B. G. Hess, F. A. Carranti, and Mrs. M. Roehrig for their assistance in parts of this work. Finally, they are particularly indebted to R. N. Hall and T. Soltys for helpful discussions.

The Reflected-Beam Kinescope*

H. B. LAW†, FELLOW, IRE, AND E. G. RAMBERG†, FELLOW, IRE

Summary—**The reflected-beam kinescope is a short tube particularly suited for picture sizes of 21 inches or more. It retains the conventional gun and external deflection components and is axially symmetric. Although the effective deflection angle is nearly 180°, the deflection power required is equivalent to that used in conventional 90° kinescopes.**

The electron beam scans through an apertured phosphor screen and is reflected back to the screen by the faceplate. The screen attenuates the beam by a factor of 4, except in the case of a radial scan where a cutout in the center of the screen may be made to permit free passage of the beam.

The results of a theoretical and experimental study of the tube are presented. They show the raster to be inherently barrel-distorted. However, the distortion may be corrected in several ways. The resolution is somewhat lower than in standard tubes, although it is considered to be adequate. High detail contrast is obtained.

* Received by the IRE, December 4, 1959; revised manuscript received, February 12, 1960.

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Introduction

In the last few years, the trend in kinescope development has been to shorter and shorter tubes, while a degree of stability has been reached in regard to screen size. The reduction in tube length has been obtained by increasing the deflection angle and, more recently, by the development of shorter electron guns. Further gains to be achieved by these approaches appear to be limited.

In addition to short kinescopes, there have appeared electron-beam display devices of an unconventional nature designed to be very short.[1,2] Sometimes these de-

[1] W. R. Aiken, "A thin cathode ray tube," Proc. IRE, vol. 45, pp. 1599–1604; December, 1957.

[2] D. Gabor, "A new flat picture tube," *J. Television Soc.*, vol. 8, pp. 142–145; October, 1956.

vices are referred to as thin cathode ray tubes. They achieve their compactness by the use of traveling electrostatic deflection fields in at least one of the two scanning directions, utilizing the strong focusing action of the deflection field to obtain a sharp spot. It is the purpose of this paper to describe still another approach to a thin display device that involves a radical change in tube design, but has essentially the same operative characteristics as a conventional kinescope.

General Description

A schematic drawing of the tube for television display purposes is shown in Fig. 1. The spherically curved tube face has a transparent conductive coating on the vacuum side that is operated at cathode potential and serves to reflect the electron beam. Mounted a few inches from the reflector and curved in a manner similar to the reflector, is an apertured, shell-like structure operating at high potential. On the convex side of this structure is deposited the phosphor screen. The electron beam scans through the holes in the screen, is reflected by the faceplate, falls back to the apertured screen, and strikes the phosphor that lies on the surface between the holes. The picture is, therefore, viewed directly on a front-surface phosphor screen that lies a few inches back of the faceplate.

A conventional deflection yoke is mounted very close to the screen on the axis of the tube and is activated by standard deflection circuits. The gun is in the usual place with respect to the deflection yoke.

Although a perforated screen is required for the central area only, the entire screen is perforated to achieve screen uniformity. If, however, a radial-type scan is used, such as for radar displays, a hole of about 3-inch diameter may be cut out of the center to permit free passage of the electron beam, and the screen may be placed on a continuous surface instead of an apertured structure.

Tube Characteristics

Deflection

Up to the point where the electron beam leaving the gun penetrates the apertured phosphor screen, the deflection of the beam is the same as in a standard kinescope. The beam, upon emerging from the phosphor screen electrode, is decelerated in a retarding field and finally falls back to the screen. A study of the raster shape to be expected under these circumstances has been made, assuming the arrangement shown in Fig. 2, where for mathematical simplicity the phosphor screen and reflector are considered to be spherical and concentric. Moderate deviations from the condition of concentricity will lead to only moderate deviations of the observed results from those given by the formulas developed below.

The distance of the point of return of the scanning spot from the tube axis is seen from Fig. 2 to be

$$r_g = R \sin \phi_g. \tag{1}$$

It is shown in the Appendix that

$$\phi_g = \arcsin \left\{ \sin\theta \left(\sqrt{1 - \left(1 - \frac{p}{R}\right)^2 \sin^2\theta} - \left(1 - \frac{p}{R}\right) \cos\theta \right) \right\} + 2 \arcsin \frac{2q\left(1-\frac{p}{R}\right)\sin\theta\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}}{\sqrt{A'^2R^2 - 4q(A'R-q)\left(1-\frac{p}{R}\right)^2\sin^2\theta}} \tag{2}$$

where

$$A' = (1 - V_s/V)(1 + q/R) \tag{3}$$

and the remaining symbols are indicated in Fig. 2.

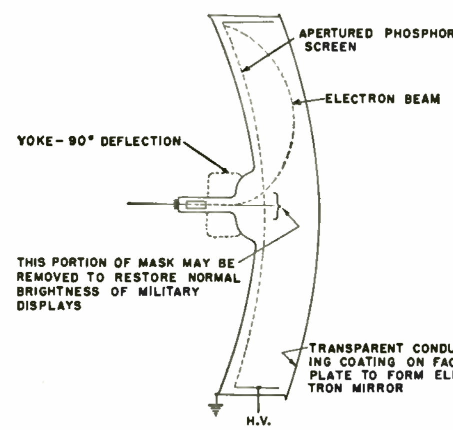

Fig. 1—Schematic drawing of reflected-beam kinescope.

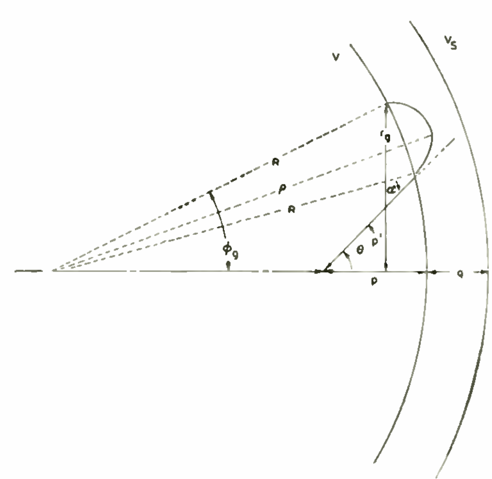

Fig. 2—Definition of symbols.

If the faceplate voltage V_s is to be held constant, the highest value of V_s for which the beam electrons do not strike the faceplate is $V_s=0$ corresponding to $A'=1+q/R$.

In Fig. 3, the landing distance of the electron beam from the tube axis vs the half scan angle has been plotted with q as a parameter. A representative tube geometry of $R=40$ inches and $p=2.4$ inches has been assumed.

The bending over of the curves in Fig. 3 shows that the raster is, in general, barrel-distorted, the distortion becoming more severe for larger q values and larger scan angles.

The curves in Fig. 3 apply for the case where the screen and reflector extend well beyond the extreme landing position of the beam. In practice, the beam path near the edge of the raster is influenced by the electrode arrangement at the edge. For example, if this electrode is at phosphor screen potential, as shown in Fig. 1, the effect is to lengthen the electron path near the end, reducing the barrel distortion.

Another factor in raster distortion is the deflection yoke. Any slight pincushion distortion of the yoke will tend to offset the barrel distortion introduced by the reflection-type scanning system. In fact, the ease with which pincushion distortion can be built into a yoke makes this an attractive approach for minimizing distortion in the picture.

In Fig. 3 it will be noticed that the spot displacement r_g does not reach a maximum as a function of angle of deflection or angle of elevation. This behavior is the result of assuming a rather strong curvature for the screen and reflector and a center of deflection some distance below the screen, *i.e.*, between the screen and the gun rather than at the screen surface.

If a flat phosphor screen and reflector is assumed, (1) becomes:

$$r_g = p \tan \theta + 2q \sin 2\theta. \tag{4}$$

In Fig. 4, r_g vs θ for (4) is plotted where $p=2.4$ inches with q as the parameter.

Comparing the curves in Figs. 3 and 4, where the screen is curved and flat respectively, it is seen that the raster is smaller for given values of p and q when the screen is curved than when it is flat.

Reducing p also results in reducing the raster size as shown in Fig. 5. Here r_g vs θ is plotted for the same conditions as in Fig. 4 except that $p=1$ inch instead of 2.4 inches.

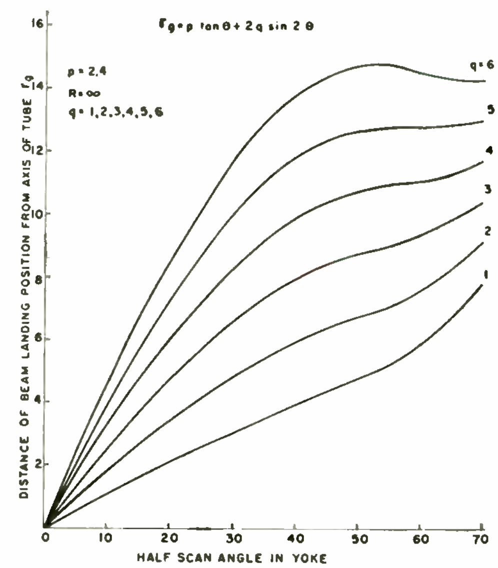

Fig. 4—Reflected-beam kinescope raster characteristics.

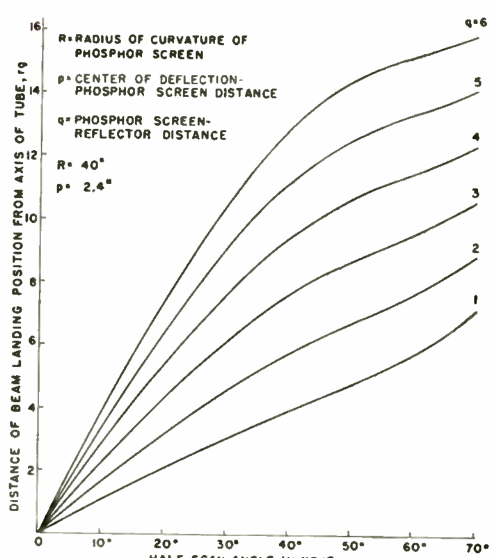

Fig. 3—Reflected-beam kinescope raster characteristics.

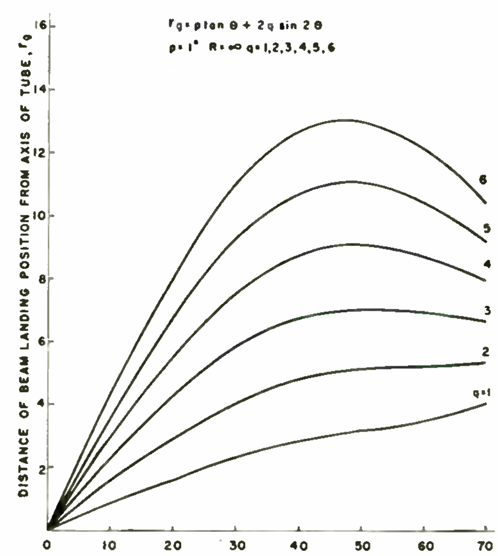

Fig. 5—Reflected-beam kinescope raster characteristics.

When $p=0$, (4) reduces to the well-known range equation for a projectile.

Up to this point, the reflector was assumed to be at zero or cathode potential. On the tube axis, for $\theta=0$, the beam grazes the reflector. As θ increases, the distance of closest approach decreases.

In principle, the beam may be made to graze the reflector for all trajectories by proper modulation of the potential of the reflector in synchronism with the scan. This modulation is given by

$$A' = (1 + q/R)\left(1 - \left[\frac{R-p}{R+q}\right]^2 \sin^2\theta\right). \tag{5}$$

Not only is the picture size increased but also the raster distortion is changed from barrel to pincushion. To show the effect, the above value of A' has been substituted in (2) and plotted in Fig. 6, where $R=40$ inches, $p=2.4$ inches, and $q=4$ inches. A plot for an unmodulated reflector ($A'=1+q/R$) is included for comparison. With proper modulation it should be possible to correct distortion and at the same time obtain a larger picture for a given tube length than without modulation of the reflector.

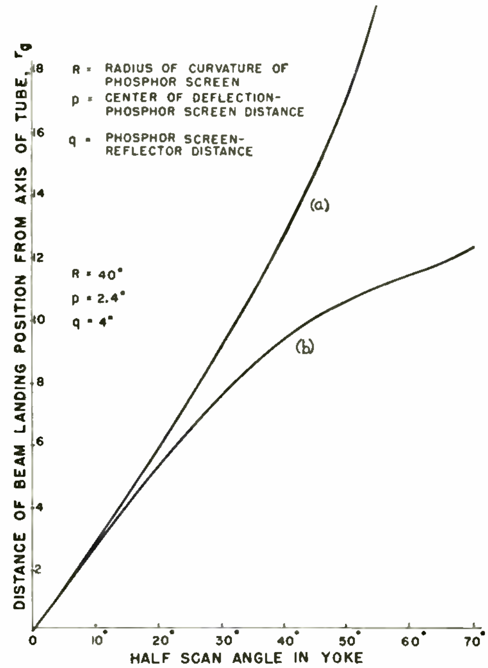

Fig. 6—Reflected-beam kinescope raster characteristic for (a) modulated and (b) unmodulated reflector.

Beam Focus

For a television-type display the electron beam scans through an apertured phosphor screen as indicated in Fig. 1. In this arrangement, there are two effects not normally encountered in a conventional cathode ray tube that tend to reduce resolution. First, the electrostatic fields about the apertures in the phosphor screen impart lateral components of velocity to the electrons passing through the mask apertures; and second, the retarding field between the phosphor screen and reflector has unilateral focusing action in the radial direction and thus distorts the spot in this direction.

In the Appendix, it is shown that the total spot-size increase in the radial direction is

$$\Delta_r D_s = \frac{2B\left(1 - 2\left(1-\frac{p}{R}\right)^2 \sin^2\theta\right)}{1+\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}} + \frac{8q(D_c - D_s)\left(1-\frac{p}{R}\right)^2 \sin^2\theta\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}}{(4q+AR)\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta} - AR\left(1-\frac{p}{R}\right)\cos\theta} \tag{6}$$

where

B = diameter of apertures in screen
D_s = spot diameter for conventional kinescope with same electron gun and equal picture size and deflection angle
D_c = beam diameter in deflection plane
$A = (1 - V_s/V)$

and the remaining symbols are indicated in Fig. 2.

The first term gives the electron spot-size increase resulting from the transverse electric field components at the apertures in the phosphor screen; and the second term gives the increase caused by the retarding field between the phosphor screen and reflector.

In the tangential direction, the spot-size enlargement is shown in the Appendix to be

$$\Delta_t D_s = 2B\left\{1 - \frac{\left(1-\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}\right)}{\left(1-\frac{p}{R}\right)^2\sin^2\theta}\right\}. \tag{7}$$

On the axis, it is seen that the spot size is increased by the aperture diameter. A plot of the increase in spot diameter as a function of the half scan angle θ is shown in Fig. 7 for a given set of conditions.

If the center of the phosphor screen is open so that the electron beam does not have to penetrate the apertures, the first term in the expression for $\Delta_r D_s$ drops out. The residual radial spot enlargement is then due to the decelerating field between the phosphor screen and reflector. In Fig. 7, it may be found by subtracting the dotted curve from the curve for $\Delta_r D_s$.

Fig. 8 shows a monoscope pattern taken from the 21-inch tube shown in Fig. 9 with the electrode struc-

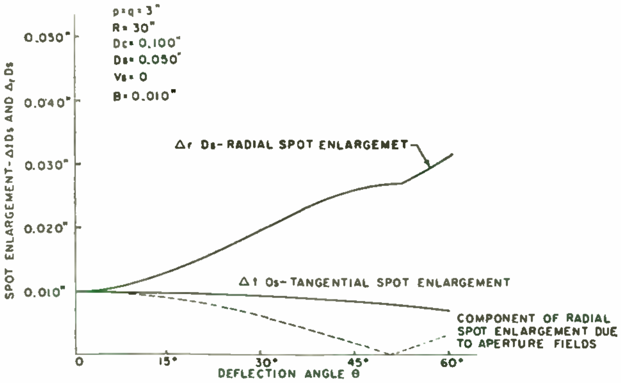

Fig. 7—Spot-size increase in the radial and tangential direction for a given set of conditions.

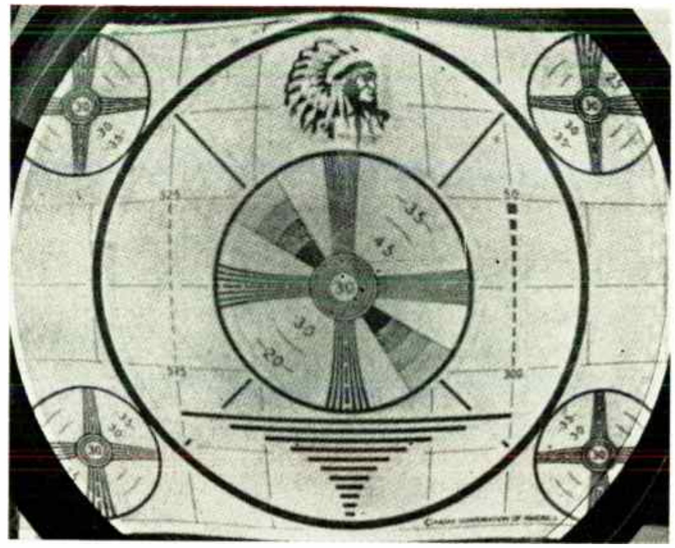

Fig. 8.

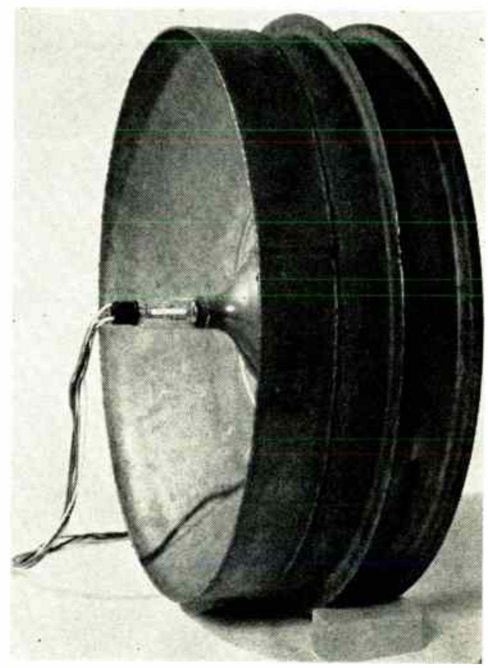

Fig. 9.

ture in Fig. 1. In this tube, $q=3.5$ inches and the radius of both the screen and reflector is 26 inches. The tube was deflected by a standard 110° deflection yoke driven by conventional circuits.

Brightness

In a television-type display where the beam must penetrate the screen apertures, the fraction of the beam that is reflected back to the phosphor screen is just the transmission factor, T, of the screen. Since the screen surface is apertured, the fraction of the total area covered with phosphor is $(1-T)$. The brightness obtainable is then proportional to $T(1-T)$. The brightness becomes a maximum for $T=50$ per cent corresponding to an attenuation by a factor of 4. The maximum is broad so that the light output is affected only slightly by the exact value of the screen transmission.

Another factor influencing brightness is the nature of the front-surface screen. The light output of a conventional aluminized screen has been compared with that of a front-surface screen in which the screen was placed on a highly reflecting surface. Within the experimental error, the light outputs were the same for equal areas of phosphor.

If the light attenuation of the conducting coating on the reflector is considered to be equivalent to that of the grey glass normally used in standard tube faces, the reflected beam kinescope for television display purposes has a brightness potential, when all factors are considered, of one-fourth that of a conventional tube. If the center of the screen is cut out for radar-type displays, the potential brightness is the same as for conventional tubes.

Contrast

A particular advantage of the reflected beam kinescope is the absence of halation.[3] This results from the fact that the screen is not in optical contact with the faceplate. As a consequence the picture exhibits high detail contrast.

Large-area contrast might be expected to suffer from high-velocity scatter of electrons from the phosphor screen. The scattered electrons encounter a retarding electric field between the screen and reflector which causes them to fall back to the phosphor screen. An intense area of illumination, therefore, has the effect of reducing contrast in surrounding areas. However, it does not produce a halo effect comparable to that observed in post-acceleration kinescopes. The reason is that in the reflected-beam kinescope the screen-reflector distance is such that the maximum range of the scattered electrons is approximately half the diameter of

[3] R. R. Law, "Contrast in kinescopes," Proc. IRE, vol. 27, pp. 511–524; August, 1939.

the picture area. The large area over which the electrons scatter dilutes their effect. Hence, high-velocity scatter is less objectionable than in tubes where concentrated halos are formed.

Picture Size vs Tube Length

The tube shown in Fig. 9 is not a production design but was fabricated from parts of existing tubes. It has an over-all length of 10 inches. A gun in a 110° funnel was used. Assuming that the same gun and funnel were used for other tube sizes, an estimate has been made of the over-all tube length required. These estimates are plotted in Fig. 10, together with lengths of existing conventional tubes. It is apparent from the curves that the length advantage of the reflected-beam kinescope increases with tube size.

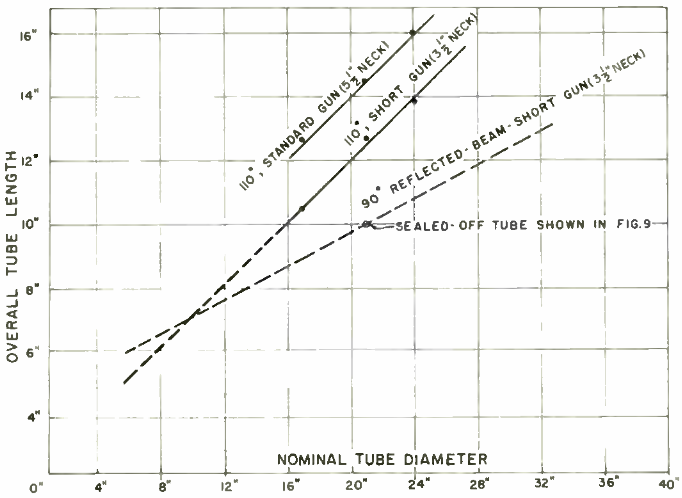

Fig. 10—Reflected-beam tube length compared to standard kinescopes.

Fabrication of Tube

Screen

A color-tube aperture mask with enlarged holes was used as the foundation structure for the phosphor screen. Phosphor was settled on the convex side of the mask.

A problem peculiar to the reflected-beam kinescope is the fabrication of the screen in such a way that the phosphor at the edge of the holes cannot be excited by the beam in its first passage through the screen. If such excitation were allowed, a small raster would appear in the center of the tube and mar the appearance of the raster formed by reflection of the beam. Fig. 11 shows the small raster produced in the center. In this figure, the picture consists of a white rectangle bordering the raster. The small raster in the center marks the first passage of the beam through the mask. Note that the small raster is slightly pincushioned while the picture is slightly barreled.

In the original picture, a third faint rectangle appears because of reflections from the tube face. Furthermore,

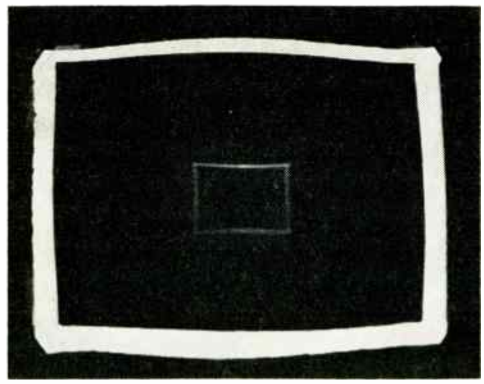

Fig. 11.

direct light from the gun produced a pattern at the center of the field. The picture has been retouched to remove these patterns so as not to confuse the illustration of the effect under discussion.

A formal solution to the problem is either to remove the phosphor around the holes or to cover it up so that the beam cannot strike it. The latter method has been found to work satisfactorily by application of a carbon black spray in the central area on the concave side of the screen. In effect, the phosphor in the holes that the beam can see is covered up with carbon. Satisfactory suppression of the central raster is achieved with less than a one mil coating of carbon.

Reflector

The reflector consists of a semitransparent conducting layer applied to the inside of the faceplate of the tube. The surface resistivity can be quite high since no beam current strikes the surface. In the tube in Fig. 9, an evaporated coating of rhodium is used with an optical transmission of 85 per cent and a surface resistivity of 2×10^5 ohms. Since low reflectivity as well as high transmission is desirable, other materials such as the alloy Inconel[4] may be better suited for the coating material.

Envelope

The nature of the electron path, as shown in Fig. 1, indicates that an unconventional bulb design may be used to reduce the tube length or effectively increase the deflection angle. Primarily, the shortening comes about by being able to recess or telescope the neck of the tube into the bulb. The relation of tube length and picture size has already been discussed and is represented in Fig. 10.

The tube shown in Fig. 9 was made by piecing together a 21-inch metal color-tube topcap and a metal cone. For simplicity in construction, the phosphor screen was electrically connected to the metal cone so that the cone operates at high voltage. If the screen were insulated from the cone, the cone could run at ground potential.

[4] L. Holland, "Vacuum Deposition of Thin Films," John Wiley and Sons, Inc., New York, N. Y., p. 195; 1956.

In the test pattern in Fig. 8, there is a distortion at the top caused by the feedthrough contact to the semi-transparent reflector coating. This effect would also disappear if the cone were designed to operate at ground potential.

The concave section in the cone represents available space for placement of circuit components if extreme compactness is desired.

The reflection of the beam by the retarding field between screen and faceplate necessitates recessing the picture behind the faceplate of the tube. This limits the viewing angle of the picture. On the other hand, the recess makes the picture more tolerant of ambient lighting conditions.

Conclusions

A kinescope having axial symmetry but an unconventional design has been described. The characteristics of the tube and the principal problems may be summarized as follows.

1) The tube length is less than that for conventional designs, especially for larger screen sizes.

2) Ninety-degree deflection power is used to obtain effective deflection angles of approximately 180°.

3) Standard-type deflection yokes and deflection circuits may be used with the tube.

4) Lack of halation makes for good detail contrast since the front surface of the phosphor is viewed.

5) Large area contrast may be somewhat lower than for a standard type tube.

6) The envelope may be run at ground potential.

7) There is space in a recessed part of the tube for circuit components.

8) The brightness for television display is down 4 to 1 by loss of beam current at the phosphor screen. This does not apply for radial scan where no information is to be displayed in the center. In this case, the brightness is that of a conventional tube.

9) The picture viewing surface is recessed within the tube.

10) Suppression of a vestigial image at the center of the tube is necessary. This image is produced by the first passage of the beam through the screen electrode. Suppression is not required in a radial type scan where the center may be removed.

11) Although the resolution is somewhat lower than for a standard tube, observation indicated that acceptable resolution may be obtained.

Appendix

For reasons of mathematical simplicity, all the formulas given below relate to tubes with concentric spherical phosphor screens and faceplate reflectors. For convenient reference, the symbols employed in the formulas are listed alphabetically below; the geometric meaning of some of them is indicated in Fig. 2.

Definition of Symbols

$A = 1 - V_s/V$

$A' = (1 - V_s/V)\ (1 + q/R)$

B = diameter of screen apertures

b = distance between point of incidence of beam on screen and point of sharp focus of the beam

D_c = diameter of beam in deflection plane

D_s = diameter of focused spot (at distance b from mask) in absence of retarding field

$-e$ = charge of electron

m = mass of electron

p = distance between plane of deflection and center of screen

q = separation of screen and inner (conducting) surface of faceplate

r_g = distance of point of return to screen (scanning spot) from tube axis

r_0 = distance of point of incidence on screen from tube axis

R = radius of curvature of screen

t_r = transit time of electron between incidence on screen and return to screen

V = voltage of screen (with respect to gun cathode)

V_s = voltage of faceplate electrode (with respect to gun cathode)

α = angle of incidence on screen ($\sin\alpha = (1 - p/R)\sin\theta$)

$\Delta_t D_s$ = spot enlargement in tangential direction

$\Delta_r D_s$ = spot enlargement in radial direction

$\Delta\sqrt{2eV_r/m}$ = radial velocity component imparted by aperture fields to electron grazing edge of aperture

$\Delta\sqrt{2eV_t/m}$ = tangential velocity component imparted by aperture fields to electron grazing edge of aperture

θ = angle of deflection

ρ = distance of point between screen and faceplate from their common center of curvature

ϕ = angular coordinate of point with respect to tube axis, measured with center of curvature as origin

ϕ_0 = central angle of point of incidence on screen

ϕ_g = central angle of point of return to screen

$\xi = 1/\rho - A(1 + q/R)/2q \sin^2\alpha$

(dot) = derivative with respect to time

(prime) = derivative with respect to ϕ

Derivation of General Formulas

The path equations in the central field between screen and faceplate are

$$\ddot{\rho} - \rho\dot{\phi}^2 = -\frac{eV}{m}\frac{AR^2}{q\rho^2}\left(1 + \frac{q}{R}\right) \qquad (8)$$

$$\frac{d}{dt}(\rho^2\dot{\phi}) = 0; \qquad \rho^2\dot{\phi} = R\sqrt{\frac{2eV}{m}}\sin\alpha. \tag{9}$$

The second equation makes it possible to translate the variation with time in the first equation into a variation with central angle ϕ:

$$\rho'' - \frac{2\rho'}{\rho} - \rho + \frac{A(1+q/R)\rho^2}{2q\sin^2\alpha} = 0. \tag{10}$$

The substitution

$$\xi = \frac{1}{\rho} - \frac{A(1+q/R)}{2q\sin^2\alpha} \tag{11}$$

leads to

$$\xi'' + \xi = 0. \tag{12}$$

This differential equation is solved by

$$\phi - \phi_0 = \arcsin(\xi/\xi_v) - \arcsin(\xi_0/\xi_v) \tag{13}$$

with

$$\xi_0 = \frac{1}{R} - \frac{A(1+q/R)}{2q\sin^2\alpha} \tag{13a}$$

$$\xi_v{}^2 = \left(\frac{1}{R} - \frac{A(1+q/R)}{2q\sin^2\alpha}\right)^2 + \frac{\cot^2\alpha}{R^2}. \tag{13b}$$

Eq. (13) can also be written in the form

$$\frac{1}{\rho} = \frac{A(1+q/R)}{2q\sin^2\alpha} + \xi_v\sin(\phi - \phi_0 + \arcsin(\xi_0/\xi_v)). \tag{14}$$

This is the equation of a hyperbola with the focal point at the origin.

Formulas for the Spot Displacement as Function of Deflection Angle

We find directly from (13):

$$\phi_g = \theta - \alpha + 2\arcsin\frac{q\sin 2\alpha}{\sqrt{A'^2R^2 - 4q(A'R-q)\sin^2\alpha}}$$

$$= \arcsin\left\{\sin\theta\left(\sqrt{1-\left(1-\frac{p}{R}\right)\sin^2\theta} - \left(1-\frac{p}{R}\right)\cos\theta\right)\right\}$$

$$+ 2\arcsin\frac{2q\left(1-\frac{p}{R}\right)\sin\theta\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}}{\sqrt{A'^2R^2-4q(A'R-q)\left(1-\frac{p}{R}\right)^2\sin^2\theta}} \tag{15}$$

$$r_g = R\sin\phi_g. \tag{16}$$

If the faceplate voltage V_s is to be held constant, the highest value of V_s for which the beam electrons do not strike the faceplate is $V_s = 0$, corresponding to

$$A = 1, \qquad A' = 1 + q/R. \tag{17}$$

For this condition, which yields maximum deflections for fixed voltages, we obtain

$$\phi_g - \phi_0 = 2\arcsin$$

$$\cdot\frac{2q\left(1-\frac{p}{R}\right)\sin\theta\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}}{\sqrt{(R+q)^2 - 4qR\left(1-\frac{p}{R}\right)^2\sin^2\theta}}. \tag{18}$$

For a flat mask, we obtain simply

$$r_g = p\tan\theta + 2q\sin 2\theta. \tag{19}$$

It is theoretically possible to increase the scanning range by modulating the faceplate voltage, making it increasingly positive as the deflection angle increases. The maximum modulation which can be used corresponds to the grazing of the faceplate by the beam for all angles of deflection. This modulation is given by

$$A' = (1+q/R)\left(1 - \left(\frac{R-p}{R+q}\right)^2\sin^2\theta\right). \tag{20}$$

With condition (20) we obtain

$$\phi_g - \phi_0 = 2\arcsin$$

$$\cdot\frac{\frac{2q}{R}\,\frac{R-p}{R+q}\sin\theta\sqrt{1-\left(\frac{R-p}{R}\right)^2\sin^2\theta}}{1-\left(\frac{R-p}{R+q}\right)^2\left(1+\frac{2q}{R}\right)\sin^2\theta}. \tag{21}$$

Formulas for the Increase in Spot Diameter Above That for a Normal Kinescope

The formulas for the spot diameter given below have been computed for plane parallel fields between mask and faceplate. Since the radii of curvature of these electrodes are large compared to their separation, this simplification leads to small error. If the beam is focused for maximum spot sharpness on the screen side of the mask, the length of the beam becomes effectively

$$L = p' + b = p' + \sqrt{\frac{2eV}{m}}\,t_r \tag{22}$$

where, for a spherically symmetric system, the transit time t_r is given by

$$t_r = \sqrt{\frac{m}{2eV}}\left(\frac{\pi}{2}\,\frac{A'R^2\sqrt{q}}{(A'R-q)^{3/2}} - \frac{A'R^2\sqrt{q}}{(A'R-q)^{3/2}}\right.$$

$$\cdot\arcsin\frac{A'R-2q}{\sqrt{A'^2R^2-4q(A'R-q)\sin^2\alpha}}$$

$$\left. + \frac{2qR\cos\alpha}{A'R-q}\right). \tag{23}$$

For a flat field ($R\rightarrow\infty$), we obtain from (23) simply

$$t_r = \sqrt{\frac{m}{2eV}}\,\frac{4q}{A}\cos\alpha. \qquad b = \frac{4q}{A}\cos\alpha. \tag{24}$$

In terms of the deflection angle θ, we can write

$$p' = R\left\{\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta} - \left(1-\frac{p}{R}\right)\cos\theta\right\} \tag{25}$$

$$b = \frac{4q}{A}\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta} \tag{26}$$

We shall denote by D_s the diameter of the spot which would be formed at a distance $p'+b$ from the deflection plane.

The actual spot dimensions exceed D_s for two reasons: 1) The electrostatic fields about the apertures in the mask impart lateral components of velocity to the electrons passing through the mask apertures; and 2) the retarding field between mask and faceplate has unilateral focusing action in the radial direction, which distorts the spot in this direction.

In an azimuthal direction, the spot-size increase is given simply by

$$\Delta_t D_s = 2\Delta\sqrt{\frac{2eV_t}{m}}\,t_r = \frac{2B\tan(\alpha/2)}{\tan\alpha}. \tag{27}$$

In the radial direction, the aperture fields effect a maximum change in direction $\Delta\alpha$, given by

$$\sin(\alpha+\Delta\alpha) - \sin\alpha = \Delta\sqrt{\frac{V_r}{V}};$$

$$\Delta\alpha = \sec\alpha\,\Delta\sqrt{\frac{V_r}{V}} = \frac{BA}{8q}\sec^2(\alpha/2). \tag{28}$$

The corresponding spot enlargement is

$$2\cdot\frac{4q}{A}\cos(2\alpha)\Delta\alpha = \frac{2B\cos(2\alpha)}{1+\cos\alpha}. \tag{29}$$

The effect of the mask apertures is thus to increase the spot size both in the tangential and in the radial direction by an amount, equal to the spot diameter on the tube axis, which decreases with increasing deflection angle.

In order to determine the focusing of the rays of the beam which are not deflected by the aperture fields, we consider the extreme rays of the beam in the meridional plane, *i.e.*, those that pass from the top and the bottom of the disk of diameter D_c in the deflection plane to the top of the spot disk of diameter D_s.

In the plane of the mask, we have for the distance of the ray in question from the central ray

$$\Delta r_0 = \frac{1}{2}\,\frac{p'D_s \pm bD_c}{p'+b} \tag{30}$$

and for the angular deviation from the central ray

$$\Delta\alpha = \frac{D_s \mp D_c}{2}\;\frac{\cos\alpha}{p'+b}. \tag{31}$$

We thus obtain for the distance of these rays from the center of the beam upon their return to the mask:

$$\Delta r_s = \Delta r_0 + 4q\cos(2\alpha)\Delta\alpha/A \tag{32}$$

$$= \frac{D_s}{2} + (\pm D_c - D_s)\sin^2\alpha\left(\frac{b}{p'+b}\right). \tag{33}$$

In all practical cases (and in particular for $\alpha<45°$), the upper sign indicates the ray establishing the boundary of the spot. Hence, the total increase in the spot diameter becomes

$$\Delta_r D_s = \frac{2B\cos(2\alpha)}{1+\cos\alpha} + 2(D_c - D_s)\sin^2\alpha\,\frac{b}{p'+b}$$

$$= \frac{2B\left(1-2\left(1-\frac{p}{R}\right)^2\sin^2\theta\right)}{1+\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}} + \frac{8q(D_c-D_s)\left(1-\frac{p}{R}\right)^2\sin^2\theta\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta}}{(4q+AR)\sqrt{1-\left(1-\frac{p}{R}\right)^2\sin^2\theta} - AR\left(1-\frac{p}{R}\right)\cos\theta}. \tag{34}$$

Thus, on the axis, the spot size is increased by the aperture diameter. For large deflection angles, this spot-size increase is augmented if the beam diameter in the deflection plane is relatively large and reduced if it is very small. Since, in general, $D_c \gg D_s$, the added term generally signifies an increase in the spot size.

Acknowledgment

The authors are indebted to Dr. D. W. Epstein and Dr. H. Johnson, under whose direction the work was done, for their interest and many helpful discussions. The work of L. Korsakoff in the Princeton Laboratory was of great value in solving some screening problems peculiar to the reflected-beam kinescope, while the sealed-off tube shown in Fig. 9 was designed and fabricated by F. van Hekken and his co-workers at the RCA Tube Division in Lancaster, Pa.

Shot Noise in Tunnel Diode Amplifiers*

J. J. TIEMANN†, MEMBER, IRE

Summary—**The contributions of Johnson noise and shot noise to the noise in a tunnel diode are analyzed according to a simple theoretical model. The current stream flowing across the junction in one direction is assumed to be uncorrelated with that flowing across the junction in the opposite direction, and both current streams are assumed to produce full shot noise. A simple parabolic band structure is also assumed. There is qualitative, but not quantitative, agreement between the predictions of this model and the experiment. On the basis of some simplifying assumptions (which are consistent with the experiment), the noise figure of a tunnel diode amplifier is calculated. A simple graphical method of determining the approximate noise figure of a diode from its current voltage characteristic is presented. It is found that the noise figure has a broad minimum centered about a point slightly higher in bias than the bias value for maximum negative conductance.**

I. Introduction

IT HAS been reported in the literature[1,2] that tunnel diodes display little or no excess noise in the bias region where the I-V characteristic displays a negative dynamic conductance. This paper consists of a more quantitative measurement of the noise in this bias region and an analysis of the noise figure of a tunnel diode amplifier based on these measurements.[3] This analysis corrects some errors which have appeared in the literature.[4] The next section describes a simplified model for the tunnel diode, and uses this model to predict its noise properties. Section III relates the noise properties of the tunnel diode to the noise figure of a small signal amplifier, and the last two sections describe the experimental procedure and results.

II. Theoretical Model

Following Esaki,[5] we assume that there are two current streams flowing across the junction in opposite directions. These currents are due to the fact that an electron occupying a state on one side of the junction can cross the barrier by means of the quantum mechanical tunnel effect if there is an appropriate unoccupied state available on the other side (see Fig. 1). It will be seen that the current flowing in the reverse direction is the familiar Zener current. For the sake of definiteness, we will call the current stream flowing in the forward direction the Esaki current.

The Zener current I_Z from the valence band to the conduction band, and the Esaki current I_E flowing from the conduction band to the valence band are taken as follows:

$$I_E = A \int_{E_c}^{E_v} f_c(E)\rho_c(E)Z_{c\to v}[1 - f_v(E)]\rho_v(E)dE, \quad (1a)$$

$$I_Z = A \int_{E_c}^{E_v} f_v(E)\rho_v(E)Z_{v\to c}[1 - f_c(E)]\rho_c(E)dE, \quad (1b)$$

where A is an appropriate arbitrary constant.

In the expression for I_E, $f_c(E)$ is a Fermi function denoting the probability that a state at energy E in the conduction band (on the n-type side) is occupied, and $\rho_c(E)$ is the density of such states at energy E. The quantity $Z_{c\to v}$ is the probability per unit time that an electron in a particular state in the conduction band will appear on the other side of the barrier, and that the remaining factors are the probability that a particular state in the valence band is empty, and the density of such states at the energy E, respectively. The integration ranges over the overlap region where the two bands are "crossed." The factors in the expression for I_Z have a similar meaning.

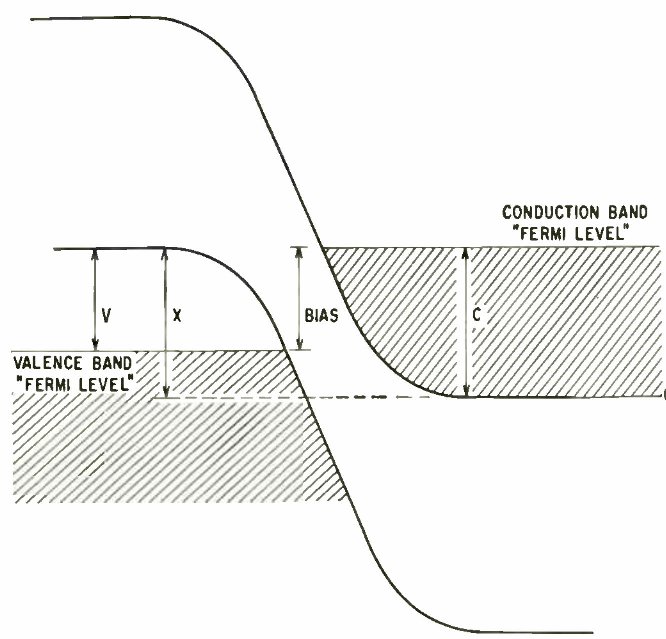

Fig. 1—The electron potential is plotted vs the dimension perpendicular to a doubly degenerate p-n junction for the condition of zero bias. The Fermi level is shown at an energy C within the conduction band on the n-type side, and at an energy V within the valence band on the p-type side. The integration shown in the expression ranges over the region where the bands overlap.

* Received by the IRE, February 8, 1960; revised manuscript received, March 25, 1960.

† General Electric Res. Lab., Schenectady, N. Y.

[1] T. Yagima and L. Esaki, "Excess noise in narrow germanium p-n junctions," *J. Phys. Soc. Japan*, vol. 13, pp. 1281–1287; November, 1957.

[2] H. S. Sommers, Jr., "Tunnel diodes as high-frequency devices," Proc. IRE, vol. 47, pp. 1201–1206; July, 1959.

[3] Most of the material in this paper was presented orally by the author at the IRE Solid-State Device Research Conference, Cornell University, Ithaca, N. Y., June, 1959.

[4] K. K. N. Chang, "The optimum noise performance of tunnel-diode amplifiers," Proc. IRE, vol. 48, pp. 107–108; January, 1960. Because of some errors in analysis and interpretation, some of the conclusions drawn in this letter are incorrect.

[5] L. Esaki, "New phenomenon in narrow germanium p-n junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January, 1958.

In actual fact, the Z's are functions of both the integration variable and the applied bias. The calculation of the value of Z has been the subject of several publications.[6] For the present analysis, however, we shall continue to follow Esaki and set

$$Z_{c\to v} = Z_{v\to c} = \text{const.}$$

The net electron current flowing across the junction is the difference between I_E and I_Z and is given by

$$i = I_E - I_Z = A' \int_{E_c}^{E_r} [f_c(E) - f_v(E)]\rho_c\rho_v dE. \quad (2)$$

For calculation of the mean square noise current, we shall assume that both current streams are completely uncorrelated.[7] Both of these current streams will then contribute full shot noise, and the equivalent saturated diode current for the tunnel diode is the sum of I_Z and I_E (see Fig. 2).

$$i_{eq} = I_Z + I_E = A' \int_{E_c}^{E_r} [f_c + f_v - 2f_cf_v]\rho_c\rho_v dE. \quad (3)$$

It is shown in the Appendix that this expression reduces to the correct value for Johnson noise in the region of zero bias. This model is therefore not in conflict with the requirements of thermodynamics.

Upon assuming a density of state function, and fixing the location of the band edges with respect to the Fermi level, the expression for i and i_{eq} are easily evaluated as functions of applied bias and temperature. Two such curves are shown in Fig. 3. One feature worth noting is that i and i_{eq} approach each other when more than a few times kT/e of bias is applied. This fact will be used later to obtain an approximate expression for the noise figure of a tunnel diode amplifier in terms of easily measured quantities.

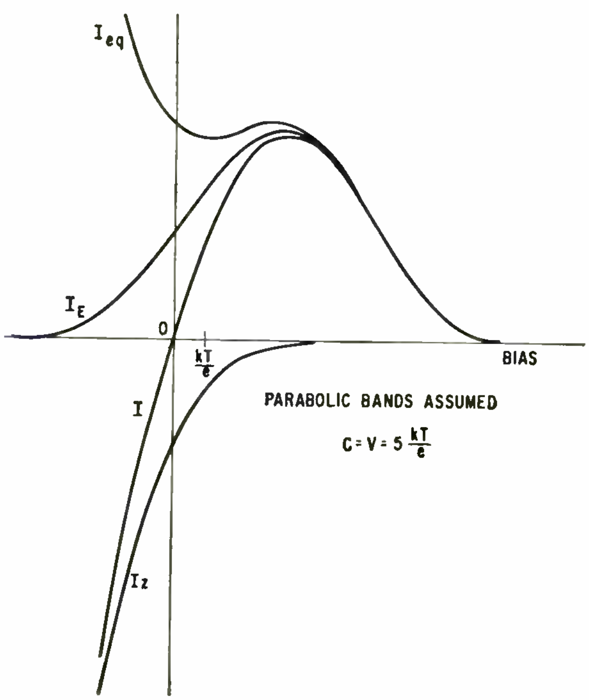

Fig. 2—The net diode current I is composed of two current streams flowing in opposition across the junction. I_Z, flowing in the reverse, is the familiar Zener current. The "Esaki" current is responsible for the negative resistance. Since these two streams are uncorrelated, their effects add (rather than subtract) in determining the noise properties of the device.

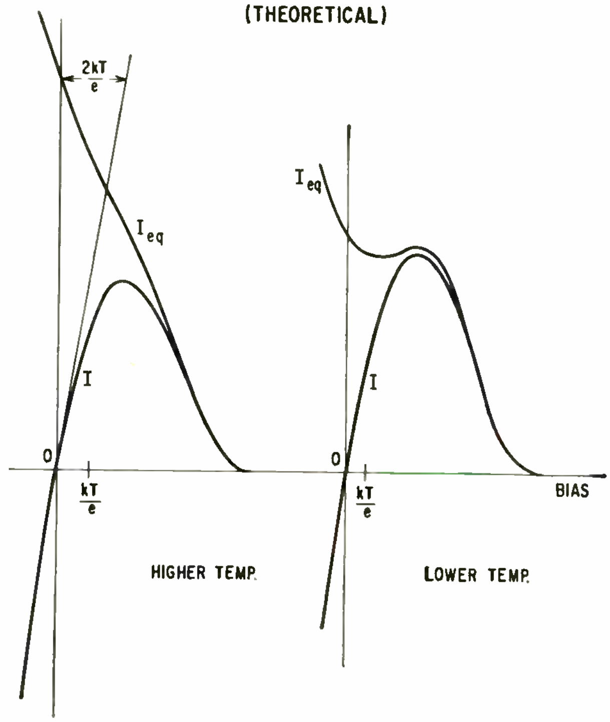

Fig. 3—Thermodynamics requires that for zero bias,

$$I_{eq} = 2kT \Big/ e \left.\frac{dI}{dv}\right|_{v=0}$$

In our model, this comes about because both I_Z and I_E fall to zero more rapidly at lower temperatures. This fact is also reflected by the higher value of the peak current at lower temperatures.

III. The Noise Figure of a Small Signal Amplifier

We now turn our attention to the relationship between the noise properties of a tunnel diode and the noise figure of an amplifier using it. Fig. 4 shows two amplifier circuits which are essentially identical.[8–10]

[6] E. O. Kane, "Observation of direct tunneling in germanium," *Phys. Rev. Letters*, vol. 3, pp. 466–468; November, 1959. See also Kane, footnote 11. This paper contains several references to earlier work.

[7] This assumption will be discussed in detail in a future publication.

[8] For additional discussion on small signal tunnel diode amplifier see, for example, U. Davidsohn, *et al.*, "Designing with tunnel diodes," *Electronic Design*, vol. 8, pp. 50–55, February 4, 1960; vol. 8, pp. 66–71, February 17, 1960.

[9] I. A. Lesk, N. Holonyak, and U. Davidsohn, "The tunnel diode—circuits and application," *Electronics*, vol. 32, pp. 60–64; November 27, 1960.

[10] E. Gottlieb, "Using the tunnel diode," *Electronic Industries*, vol. 19, pp. 110–113; March, 1960.

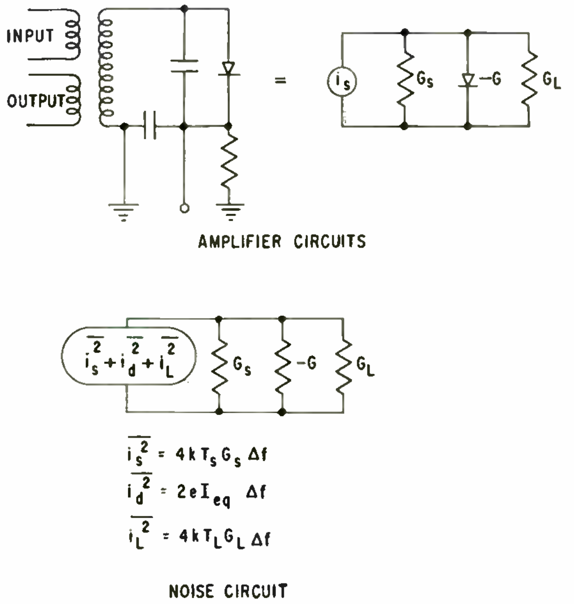

Fig. 4—Small signal tunnel diode amplifiers. The circuit on the left is a typical narrow-band amplifier circuit. At the center frequency, it is equivalent to the circuit on the right, provided G_s and G_L are the transformed source and load conductances, respectively. The generators in the lower circuit represent the fluctuating noise currents.

The only difference is that in the video amplifier the admittance seen by the diode is independent of frequency. The current generators in the equivalent circuit represent the Johnson noise in the conductances and the shot noise in the diode, and are given by

$$\overline{i_s^2} = 4kT_sG_s\Delta f,$$

$$\overline{i_d^2} = 2ei_{eq}\Delta f \approx 2ei\Delta f,$$

$$\overline{i_L^2} = 4kT_LG_L\Delta f, \tag{4}$$

where T_s and T_L are the absolute temperatures of the source conductance and the load conductance, respectively. In replacing i_{eq} by i, it is assumed that the temperature of the diode is low enough that $I_Z/I_E \ll 1$ at the operating point and that correlation effects are negligible.

Since all of these sources are in parallel, we arrive at the noise figure simply by taking the ratio of the total mean square current to that contributed by the source alone as follows:

$$F = (\overline{i_s^2} + \overline{i_d^2} + \overline{i_L^2})/\overline{i_s^2}$$

$$\approx 1 + \frac{T_LG_L}{T_sG_s} + \frac{ei}{2kT_sG_s} \cdot \tag{5}$$

In order to appreciate what (5) implies regarding an actual amplifier, let us consider the operation of the amplifier of Fig. 3 in greater detail.

The impedance into which the signal current source feeds its current is

$$Z = (G_s + G_L - G)^{-1}. \tag{6}$$

The output power is therefore

$$W_0 = e^2G_L = (iZ)^2G_L.$$

The available power from the signal source is in the same units

$$W_i = \frac{i^2}{4G_s},$$

and the gain of the amplifier is

$$G = W_0/W_i = 4G_sG_LZ^2. \tag{7}$$

According to (7) and (6), the gain can become arbitrarily large as G approaches G_s+G_L. On the other hand, we see from (5) that the minimum noise figure for the stage occurs when G_s is as large as possible. This means that for the lowest noise figure and large gain we should choose

$$G_s \gtrsim G$$

$$G_L \sim 0.$$

For this case, (5) becomes

$$F \approx 1 + \frac{ei}{2kT_sG} = 1 + \frac{T_{eff}}{T_s} \tag{8}$$

where T_{eff} is defined by (8) and is the inherent noise "temperature" of the diode.

$$T_{eff} = \frac{e}{2k}\left(\frac{i}{G}\right). \tag{9}$$

Referring to Fig. 5, the quantity i/G appearing in (9) is the horizontal distance along the voltage axis between the points at which the tangent to the operating point crosses it to the point directly below the operating point. The most favorable operating point for a given diode as well as an absolute figure of merit for the shot noise of the diode can therefore be easily obtained from the I-V characteristic. It should be pointed out that although there is a minimum value of i/G for any particular diode, there is no known general argument which states what this minimum value can be. Therefore, until more is learned theoretically about the properties of heavily doped semiconductors, nothing can be said about an ultimate limit on the noise figure of a tunnel diode amplifier. Generally speaking, we have found that diodes made from larger band gap semiconductors have larger values of (i/G) at the optimum operating point than diodes made from semiconductors with small band gaps. From the point of view of noise performance, therefore, small band gap materials are preferable. Two prominent disadvantages of small band gap tunnel diodes are the fact that they are unable to operate over a wide temperature range, and the fact that their output voltage is correspondingly lowered.

If the tunnel diode stage is the first stage of an amplifier system and if the subsequent stages of amplification are extremely noisy, then one requires a very large gain

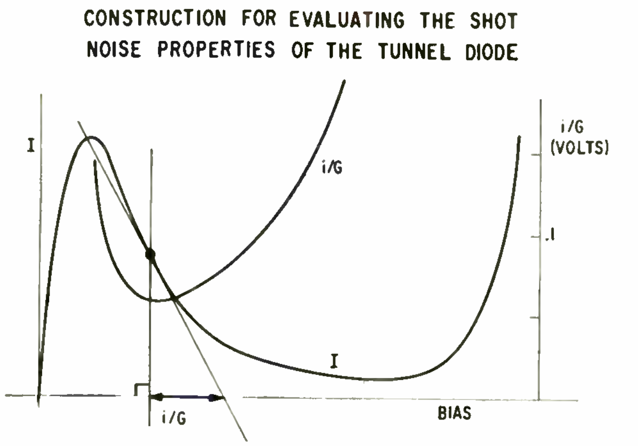

Fig. 5—In the above construction, a tangent to the characteristic curve at the operating point is drawn to the voltage axis and a perpendicular is erected through it. The horizontal distance between the intersection of these lines with the voltage axis is proportional to the effective noise temperature of the device.

from the tunnel diode to get the best noise figure, and (8) represents the best noise figure possible. On the other hand, if the second stage is not too noisy compared to the tunnel diode, not as much gain is needed from the tunnel diode, and G_s can be increased above G by a greater amount. In this case, the noise figure of the tunnel diode stage is less than that given by (8), and the over-all noise figure can also be less.

IV. Experimental Results

Fig. 6 is a plot of the I-V characteristic of a typical germanium tunnel diode. The donor concentration on the n-type side is about 4×10^{19}/cc and the acceptor concentration on the p-type side is about 10^{20}/cc. Also plotted is the equivalent saturated diode current. This quantity is defined as

$$i_{eq} = \frac{\overline{i^2}}{2e\Delta f},$$

and for an applied bias large compared to kT/e would be identical with the dc current flowing if full shot noise was present.

The circuit used at 0.5 mc is shown in Fig. 7. The circuit used at 100 mc was similar. The experimental procedure is as follows: With the tunnel diode removed and the resistance R set at some convenient value, one measures the noise power from the narrow-band amplifier. Then one increases the current in the noise diode until this power is doubled. The diode current required to do this is the equivalent saturated diode current for the preamplifier stage and the Johnson noise of the resistance. One then increases the output of the signal generator until the power meter reads a convenient value large compared to the noise. Next, one inserts the tunnel diode and replaces the resistance R by a different convenient value and adjusts the bias on the diode until the signal produces the same reading as before. At this point, the combined conductance of the

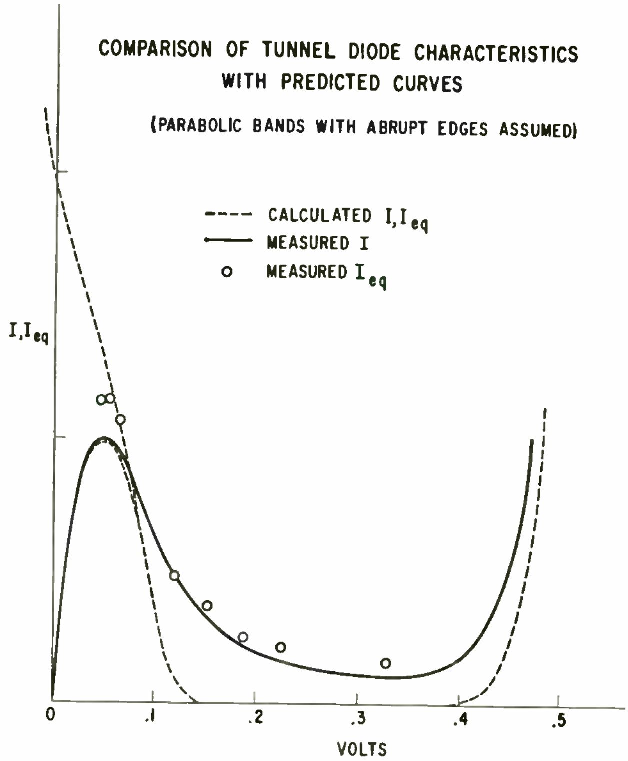

Fig. 6—Although the relationship between the dc current and the noise are similar to the calculated relationship, it is clear that our simple model (particularly as regards the abrupt band edges) is not in agreement with the measurements.

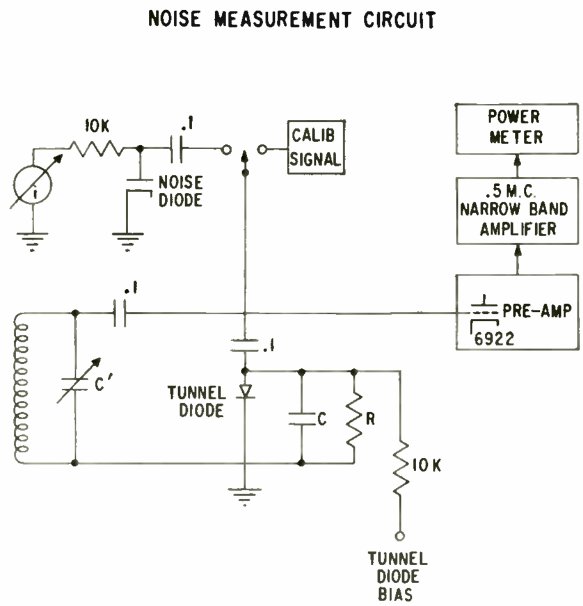

Fig. 7—The circuit for measuring the equivalent saturated diode current of a tunnel diode.

new resistor plus the tunnel diode is equal to the conductance of the original resistance alone. Under this condition, the equivalent saturated diode current for the preamplifier will be as before, and the additional noise diode current required to double the noise power is due to the diode and the added conductance. By subtracting the contributions from the conductance and the preamplifier, one obtains the value for the tunnel diode alone.

Referring to Fig. 6, we notice that the behavior of the noise is in qualitative accord with our model. That is, for applied bias in excess of a few times kT/e, the equivalent saturated diode current drops from the Johnson noise value to essentially full shot noise. We notice, however, that the negative conductance is not as great as is predicted from the calculation based on parabolic density of state functions for the conduction and valence bands. Since this fact implies a poorer noise figure for the diode than would be expected, it is of interest to examine the possible reasons why this is so. There are two possibilities: either the conduction band electrons lose energy through interactions with other entities (such as phonons, photons, and free carriers) in the crystal during the tunnelling process, or there are states for the electrons within the forbidden gap. These alternatives are represented in Fig. 8. The first of these possibilities has been examined by Kane,[11] and has been found to be unlikely. The second possibility seems at present to be the more likely. There are two types of states which could be present in the forbidden gap. There could be deep levels caused by some unknown impurity or by a complex of donor or acceptor impurity atoms, or there could be a general fuzzing of the band edge caused by the large impurity concentrations present in the crystal. The latter case is in agreement with the observed current voltage characteristic and the fact that the small band gap materials, which require fewer impurities to become degenerate, show a lower value of i/G. If deep level traps are involved, the bound states must have a lifetime less than about 2×10^{-9} second; otherwise the noise would have exceeded the shot noise value by a measurable amount.

MODELS FOR EXCESS CURRENT

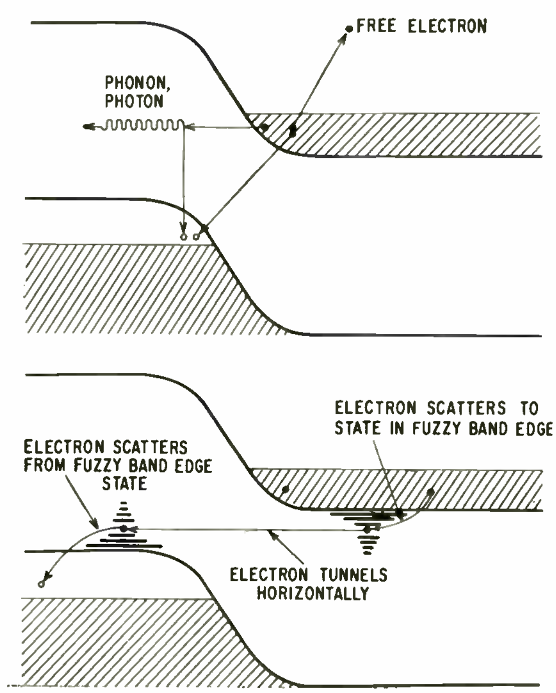

Fig. 8—In the first process, the electron loses energy through the involvement of some other entity in the tunnelling process. In the second process, the electron tunnels without loss of energy by means of states within the forbidden gap. The energy is lost both before and after tunnelling.

[11] The author is indebted to E. O. Kane for a pre-publication copy of his manuscript, "Electron-electron scattering in n-p tunneling," *Bull. Am. Phys. Soc.*, vol. 5, p. 160; March 21, 1960.

Appendix

The Relation Between Johnson Noise and Shot Noise in a Tunnel Diode

According to very general thermodynamic arguments, the noise current in any two terminal devices at thermal equilibrium (zero applied bias) must be given by

$$i_J^2 = 4kTG\Delta f, \tag{10}$$

where

$$G = \left.\frac{\partial i}{\partial v}\right|_{v=0}$$

is the conductance at thermal equilibrium. Now, according to our model, there are two opposing but uncorrelated current streams crossing the barrier. Their difference is the current i flowing, but their sum determines the shot noise.

For the mathematical steps which follow, it is convenient to express these current streams in terms of dimensionless variables. We take our unit of voltage as kT/e. By comparing Fig. 1 with Fig. 9, it is easily

ELECTRON POTENTIAL FOR A TUNNEL JUNCTION

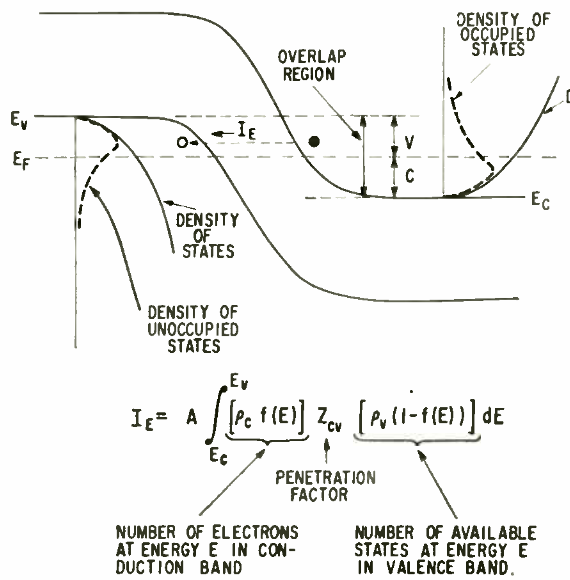

Fig. 9—The variables used in the Appendix.

seen that (2) and (3) become

$$i = A\int_0^X \sqrt{E}\sqrt{X-E} \cdot [f(E-C) - f(E+V-X)]dE, \quad (11)$$

$$i_s^2 = 2eA\Delta f\int_0^X \sqrt{E}\sqrt{X-E}[f(E-C) + f(E+V-X) - 2f(E-C)f(E+V-X)]dE, \quad (12)$$

where

$f(z) = 1/1+e^z$ is the Fermi function,
$(kT/e)\ C$ is the position of the Fermi level in the conduction band,
$(kT/e)\ V$ is the position of the Fermi level in the valence band.

The radicals appearing in the integrands represent the density of states functions and represent a simple parabolic band shape.

In terms of the dimensionless variables C, V, and X, the applied voltage is

$$v = \frac{kT}{e}(C + V - X);$$

hence

$$\frac{di}{dv} = -\frac{e}{kT}\frac{di}{dx}.$$

Now

$$\frac{di}{dx} = \Big[A'\sqrt{E}\sqrt{X-E}[f(E-C) - f(E+V-X)]\Big]^X + A'\int_0^X \sqrt{E}\frac{1}{2\sqrt{X-E}}[f(E-C) - f(E+V-X)]dE + A'\int_0^X \sqrt{E}\sqrt{X-E}\frac{\exp(E+V-X)}{[1+\exp(E-V-X)]^2}dE.$$

At zero bias, $X = C+V$. Hence, the first two terms vanish. Thus

$$\left.\frac{di}{dv}\right|_{v=0} = \frac{eA'}{kT}\int_0^{C+V}\sqrt{E}\sqrt{C+V-E}\frac{\exp(E-C)}{[1+\exp(E-C)]^2}dE,$$

and the Johnson noise is

$$i_J^2 = 4eA'\Delta f\int_0^{C+V}\sqrt{E}\sqrt{C+V-E} \cdot \frac{\exp(E-C)}{[1+\exp(E-C)]^2}dE. \quad (13)$$

Similarly at $X = C+V$, (11) gives for the shot noise

$$i_s^2 = 4eA'\Delta f\int_0^{C+V}\sqrt{E}\sqrt{C+V-E}\left[\frac{1}{1+\exp(E-C)} - \frac{1}{[1+\exp(E-C)]^2}\right]dE$$

$$= 4eA'\Delta f\int_0^{C+V}\sqrt{E}\sqrt{C+V-E}\frac{\exp(E-C)}{[1+\exp(E-C)]^2}dE,$$

in agreement with (13).

Acknowledgment

The author would like to thank R. L. Watters of this laboratory for making the noise measurements and Dr. H. Hurwitz for several fruitful discussions relating to the noise figure of tunnel diode amplifiers.

A Parametric Device as a Nonreciprocal Element*

A. K. KAMAL†, MEMBER, IRE

Summary—**A parametric device has been proposed which is equivalent to a passive nonreciprocal phase shifter. A first-order analysis is made assuming the series resistance effect in the variable capacitance diode to be negligible and a dispersion theory for matching and bandwidth of the nonreciprocal element is developed. The nonreciprocal phase behavior of the device was exhibited experimentally over a narrow frequency band.**

I. Introduction

THE importance of parametric amplifiers using semiconductor diodes has stimulated the development of various other devices. At the higher frequencies (of the order of one kmc and higher), where ferrite isolators and circulators are easily available, some problems of the negative resistance type of parametric amplifier are easily solved. At lower frequencies, however, of the order of few hundreds of megacycles, for example, it becomes difficult to separate the input and output circuitry. In other words, one should provide a nonreciprocal element between the load and the parametric device, first, in order to eliminate the effect of load impedance variations, and second, to keep the noise produced by the hot load from entering the device, being amplified and returning to the load.

In this paper certain potentialities for nonreciprocal behavior in parametric devices are exploited. A nonreciprocal circuit element has been proposed which can be used as an isolator or circulator at lower frequencies where it is not possible to obtain the ferrite equivalents.

The idea is based on the fact that if we employ an up converter parametric amplifier, as shown on the left of AA' in Fig. 1, we note that the upper sideband $\omega_u = \omega_p + \omega_s$ has a phase angle of $\phi_p + \phi_s$, where ϕ_s is the phase angle associated with the signal and ϕ_p is the angle associated with the pump. The right-hand side of this figure is a down converter parametric amplifier where the pump frequency remains the same but the pump phase angle is changed to ϕ_p'. Hence, the output is at the signal frequency with a phase angle of $\phi_p - \phi_p' + \phi_s$.

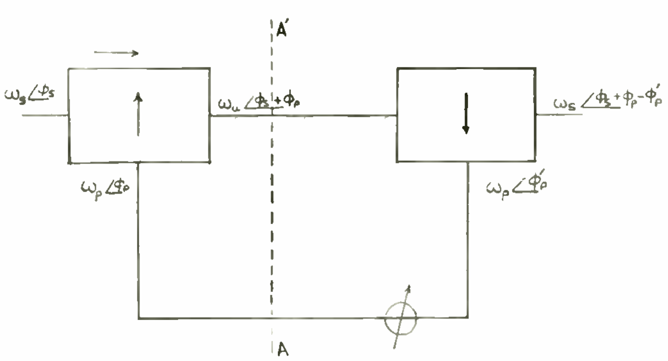

Fig. 1—Block diagram of the nonreciprocal device.

Considering the transmission from right to left, we will get the output signal frequency at a phase angle of $\phi_p' - \phi_p + \phi_s'$, which demonstrates the phase nonreciprocity. As an example, suppose that $\phi_p - \phi_p' = 90°$. In this case, the transmission from left to right will produce a phase angle of $90° + \phi_s$, while from right to left it will give $-90° + \phi_s'$. Hence, in the two directions, the device will produce a 180° phase shift, which shows a complete nonreciprocity. The amplitude of the signal is not affected because the diodes are assumed to be lossless.

II. Analysis

The analysis is based on the transfer or $ABCD$ matrix methods as described in the appropriate forms by Guillemin.[1]

The equation for a reactance parametric amplifier can be written in the usual admittance matrix form[2]

$$\begin{bmatrix} I_s \\ -I_m^* \\ I_u \end{bmatrix} = i \begin{bmatrix} C_0\omega_s & C_1\omega_s & C_1^*\omega_s \\ C_1^*\omega_m & C_0\omega_m & 0 \\ C_1\omega_u & 0 & C_0\omega_u \end{bmatrix} \begin{bmatrix} V_s \\ V_m^* \\ V_u \end{bmatrix}, \quad (1)$$

where subscripts s, m and u represent the quantities at signal, image (or lower sideband), and upper sideband, respectively; the star indicates the conjugate value.

From (1), reversing the sign of I_u and suppressing the lower sideband, the equation for an upper sideband up converter in $ABCD$ matrix form can then be expressed as

$$\begin{bmatrix} V_s \\ I_s \end{bmatrix} = \begin{bmatrix} -\dfrac{C_0}{C_1} & \dfrac{i}{\omega_u C_1} \\ i\omega_s C_1^* - i\omega_s \dfrac{C_0^2}{C_1} & -\dfrac{\omega_s}{\omega_u} \cdot \dfrac{C_0}{C_1} \end{bmatrix} \begin{bmatrix} V_u \\ I_u \end{bmatrix}. \quad (2)$$

* Received by the IRE, September 28, 1959; revised manuscript received, March 17, 1960.

† School of Elec. Engrg., Purdue University, Lafayette, Ind. This work was done by the author during the summer of 1959 at Bell Telephone Labs., Murray Hill, N. J.

[1] E. A. Guillemin, "Communication Networks," John Wiley and Sons, Inc., New York, N. Y., vol. 2, ch. 4; 1935.

[2] M. E. Hines, "Amplification in Nonlinear Reactance Modulator," presented at the 15th Conference on Electron Tube Research, Berkeley, Calif.; June, 1957.

which can be written[3] as

$$\begin{bmatrix} V_s \\ I_s \end{bmatrix} = \sqrt{\frac{\omega_s}{\omega_u} \cdot \frac{C_1^*}{C_1}} \begin{bmatrix} a & b \\ c & d \end{bmatrix} \begin{bmatrix} V_u \\ I_u \end{bmatrix} \tag{3}$$

$$= K \begin{bmatrix} a & b \\ c & d \end{bmatrix} \begin{bmatrix} V_u \\ I_u \end{bmatrix} \tag{4}$$

$$= K[\beta] \begin{bmatrix} V_u \\ I_u \end{bmatrix} \tag{5}$$

where

$$K = \sqrt{\frac{\omega_s}{\omega_u} \cdot \frac{C_1^*}{C_1}}, \qquad a = -\frac{\omega_u C_0}{\Omega |C_1|},$$

$$b = \frac{i}{\Omega |C_1|}, \qquad c = i\Omega |C_1| \left(1 - \frac{C_0^2}{|C_1|^2}\right),$$

$$d - \frac{\omega_s C_0}{\Omega |C_1|} \quad \text{and} \quad \Omega^2 = \omega_s \omega_u.$$

The operator $[\beta]$ in (5) represents a passive reactive network which does not include any pump phase relationships. The determinant of this operator is unity.

On the other hand, K is the active element and involves the pump phase relationship, *i.e.*,

$$K = |K| e^{i\phi} \quad \text{where} \quad |K| = \sqrt{\frac{\omega_s}{\omega_u}}$$

and

$$i\phi = + \frac{1}{2} \ln \frac{C_1^*}{C_1}$$

so that

$$\phi = - \text{ phase of } C_1. \tag{6}$$

But C_1 depends on the pump voltage V_p, therefore,

$$\phi = - \text{ phase of } V_p. \tag{7}$$

Consequently, it is possible to display the network represented by (3) schematically, as in Fig. 2. Eq. (3) and Fig. 2 characterize a simple nonlinear reactance diode which has as a usual representation, that shown in Fig. 3.

If such an up converter is terminated by a down converter having a similar diode as a parametric element, it is easy to write down a circuit configuration of the whole system, as shown in Fig. 4. $[\beta']$ represents $[\beta]$ with ω_s and ω_u interchanged. The gain in a lossless system is unity, since

$$V_u I_u^* = \frac{1}{|K|^2} V_s I_s^* \qquad \text{for the up converter,}$$

and

$$V_s I_s^* = |K|^2 V_u I_u^* \qquad \text{for the down converter.}$$

The phase, however, will be changed by an amount

$$\pm(\text{phase of } V_{p_1} - \text{phase of } V_{p_2}), \tag{8}$$

which is the phase difference between the two pumps.

The above relationship can also be shown simply by writing down the transmission, for example, from left to right, as follows:

$$\text{Transfer function} = \frac{1}{\underset{\text{(up converter)}}{\sqrt{\frac{\omega_s}{\omega_u} \cdot \frac{C_1^*}{C_1}}} \times \underset{\text{(down converter)}}{\sqrt{\frac{\omega_u}{\omega_s} \cdot \frac{C_1'}{C_1^{*\prime}}}}}$$

$$= e^{-i(\phi_1 - \phi_2)}, \tag{9}$$

where ϕ_1 and ϕ_2 are the phase angles of the pumps.

III. Matching and Bandwidth

It will be of great importance to consider the matching problem between the two networks $[\beta]$ and $[\beta']$. Matching is accomplished by inserting a coaxial line between the up converter and down converter parametric devices, as shown in Fig. 5. The figure represents the case of two lossless diodes, which can be justified by neglecting the series resistance at lower frequencies provided that the Q is very high.

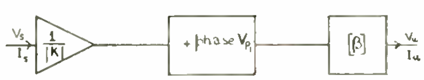

Fig. 2—Up converter parametric amplifier representation.

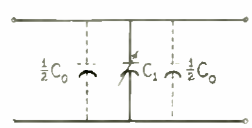

Fig. 3—Nonlinear reactance diode.

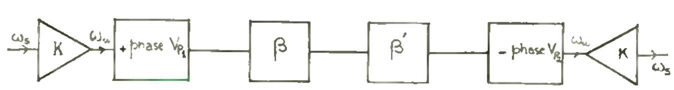

Fig. 4—Circuit configuration of an up converter-down converter parametric amplifier.

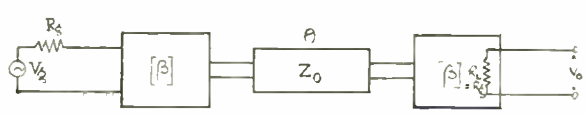

Fig. 5—Matching by a coaxial line of length θ and characteristic impedance Z_0.

[3] H. A. Haus, "Equivalent circuit for a passive nonreciprocal network," *J. Appl. Phys.*, vol. 25, pp. 1500–1502; December, 1954.

In the normalized matrix form, one can easily write that

$$\begin{bmatrix} V_g \\ I_g \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} A & iB \\ iC & D \end{bmatrix}\begin{bmatrix} \cos\theta & i\sin\theta \\ i\sin\theta & \cos\theta \end{bmatrix} \cdot \begin{bmatrix} D & iB \\ iC & A \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 1 & 1 \end{bmatrix}\begin{bmatrix} V_0 \\ 0 \end{bmatrix} \qquad (10)$$

where

$$A = -\frac{\omega_u C_0}{\Omega|C_1|}, \qquad B = \frac{1}{R_g\Omega|C_1|},$$

$$C = R_g\Omega|C_1|\left(1 - \frac{C_0^2}{|C_1|^2}\right),$$

$$D = -\frac{\omega_s C_0}{\Omega|C_1|} \text{ and } R_{\text{generator}} = R_{\text{load}} = Z_0 = R_g. \qquad (11)$$

Eq. (10) is represented by Fig. 6.

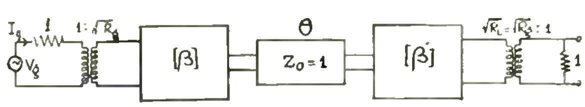

Fig. 6—Normalized representation of Fig. 5.

Performing the matrix multiplication, one gets

$$\begin{bmatrix} V_g \\ I_g \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} D(A\cos\theta - B\sin\theta) - C(A\sin\theta + B\cos\theta) & i\begin{Bmatrix} B(A\cos\theta - B\sin\theta) \\ +A(A\sin\theta + B\cos\theta) \end{Bmatrix} \\ i\begin{Bmatrix} D(C\cos\theta + D\sin\theta) \\ +C(D\cos\theta - C\sin\theta) \end{Bmatrix} & A(D\cos\theta - C\sin\theta) - B(C\cos\theta + D\sin\theta) \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 1 & 1 \end{bmatrix}\begin{bmatrix} V_0 \\ 0 \end{bmatrix}, \qquad (12)$$

which may be written more compactly in the form

$$\begin{bmatrix} V_g \\ I_g \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}\begin{bmatrix} T_{11} & iT_{12} \\ iT_{21} & T_{22} \end{bmatrix}\begin{bmatrix} 1 & 0 \\ 1 & 1 \end{bmatrix}\begin{bmatrix} V_0 \\ 0 \end{bmatrix} \qquad (13)$$

$$= \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix}[T]\begin{bmatrix} 1 & 0 \\ 1 & 1 \end{bmatrix}\begin{bmatrix} V_0 \\ 0 \end{bmatrix} \qquad (14)$$

where

$$T_{11} = T_{22} = (AD - BC)\cos\theta - \sin\theta(BD + CA),$$

$$T_{12} = 2AB\cos\theta - B^2\sin\theta + A^2\sin\theta, \text{ and}$$

$$T_{21} = 2CD\cos\theta - C^2\sin\theta + D^2\sin\theta. \qquad (15)$$

Since the determinants of

$$[\beta],\ [\beta'] \text{ and } \begin{pmatrix} \cos\theta & i\sin\theta \\ i\sin\theta & \cos\theta \end{pmatrix}$$

are unity separately, the determinant of $[T]$ must also be 1. Hence,

$$T_{11}T_{22} + T_{12}T_{21} = 1. \qquad (16)$$

At any frequency for which match is obtained, it can be shown that

$$T_{12} = T_{21}. \qquad (17)$$

The insertion gain of such a network is, in general, given by[4]

$$G_0 = \frac{4}{[\sum a_{ij}]^2} \qquad (18)$$

since

$$G_0 = \frac{\text{Available power in the unit load} = |V_o|^2}{\text{matched power} = \dfrac{|V_g|^2}{4} = \dfrac{|V_o|^2}{4}[a_{ij}]^2}, \qquad (19)$$

$$G_0 = \frac{4}{|T_{11} + T_{22} + i(T_{12} + T_{21})|^2}. \qquad (20)$$

At the center frequency, that is, when perfect matching conditions occur,

$$G_0 = \frac{4}{4(T_{11}^2 + T_{12}^2)} = 1 \qquad (21)$$

since

$$T_{11} = T_{22} \quad \text{and} \quad T_{12} = T_{21}.$$

The bandwidth of the previous system can be evaluated by varying one of the variable parameters, for example, the signal frequency ω_s. In general, one can write the matrix transform $[T]$ of (14) as

$$[T] = [T_0] + [\delta T] \qquad (22)$$

where $[T_0]$ is the value of $[T]$ at match frequency and $[\delta T]$ is the change from the center value.

In terms of a small variation $\delta\omega_s$ in the signal frequency, the gain is shown in Appendix I to take the form of

$$G = \frac{1}{1 + \dfrac{1}{4}(\delta\omega_s)^2\left[\dfrac{\partial(T_{12} - T_{21})}{\partial\omega_s}\right]^2}. \qquad (23)$$

[4] H. Seidel and G. F. Hermann, "Circuit aspect of parametric amplifier," 1959 IRE WESCON CONVENTION RECORD, pt. 2, pp. 83–90.

The three-db bandwidth can be found by equating

$$\frac{1}{4}(\delta\omega_s)^2\left[\frac{\partial(T_{12}-T_{21})^2}{\partial\omega_s}\right]^2 \text{ to } 1.$$

We will take a practical example by assuming some specific values and solving for the gain from (23) and (39) (see Appendix II). We will assume the following values:

$$\begin{aligned}
\omega_s &= 2\pi \times 500 \times 10^6 \text{ cycles/second,}\\
\omega_u &= 2\pi \times 2000 \times 10^6 \text{ cycles/second,}\\
C_0 &= 10\ \mu\mu\text{fds,}\\
|C_1| &= \sqrt{10}\ \mu\mu\text{fds, and}\\
R_g &= 50 \text{ ohms.}
\end{aligned}$$

The three-db bandwidth, as shown in Appendix III, is given by

$$\delta\omega^2 = \left(\frac{1}{122.6}\right)\omega_s^2,$$

$$\delta\omega = \frac{\omega_s}{10} = 2\pi \times 45.2 \text{ mc.}$$

Consequently, the bandwidth

$$2\Delta f = 90.4 \text{ mc.} \tag{24}$$

For a coaxial line, (39) becomes

$$\left[\frac{\partial}{\partial\omega_s}(T_{12}-T_{21})\right]^2(\delta\omega_s)^2 = 368.6\left(\frac{\delta\omega}{\omega_s}\right)^2,$$

hence, the gain by (23) becomes

$$G = \frac{1}{1 + 92\left(\frac{\partial\omega}{\omega_s}\right)^2}, \tag{25}$$

which gives the three-db bandwidth condition that

$$\delta\omega^2 = \left(\frac{1}{92}\right)\omega_s^2$$

or

$$\delta\omega = \left(\frac{1}{9.6}\right)\omega_s$$

$$= \frac{1}{9.6} \times 2\pi \times 500 \text{ mc.}$$

Hence, the bandwidth

$$2\Delta f = 104.2 \text{ mc.} \tag{26}$$

The Q of the system can also be calculated, since

$$Q_{\text{coax}} = \frac{f}{2\Delta f} = \frac{500}{104.2} \simeq 4.8 \tag{27}$$

for the coaxial case, and

$$Q_{\text{guide}} = \frac{f}{2\Delta f} = \frac{500}{90.4} \simeq 5.5 \tag{28}$$

for the guide.

IV. Experimental Set-Up

The experimental arrangement is somewhat more complicated than one might infer from the block diagram of Fig. 1. This is due to the necessity of matching each element in and out, which in turn requires a liberal use of tuning stubs. This obviously tends to make the whole arrangement of narrower bandwidth than that predicted by the theory outlined in Section III.

A simplified block diagram for a three-port circulator is given in Fig. 7, since, in this form, the gyrator action of the parametric network was most amenable to observation.

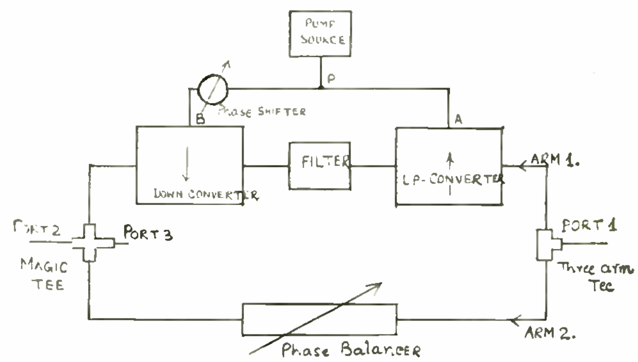

Fig. 7—Block diagram of a three-port circulator.

A magic tee is used to provide the third port. At the input port, a simple three arm tee is employed which divides the signal equally into two arms, one of which consists of the nonreciprocal parametric device and the other which has just a phase balancer in it.

In the parametric device arm, a filter for eliminating the pump is provided between the up and the down converter parametric amplifiers. This filter plays a very important part in determining the bandwidth characteristics of the device. The same pump source feeds both the up converter and down converter, the pump phase difference being introduced by a phase shifter in one arm. We have used a variable coaxial line for this purpose. It is easy to see that when the path PA is equal to PB, the two pumps are fed in phase at the points A and B. But changing the length PB gives a phase difference at the points A and B, and, consequently, one gets a nonreciprocal device. It is possible to produce any amount of phase nonreciprocity since the signal phase change is directly proportional to the pump phase change. In other words, when the two pumps have a zero phase difference, the device is completely reciprocal. As the pump phase goes from 0° to 180°, the parametric device becomes a perfect gyrator at 90° phase difference, and again becomes reciprocal at 180°.

We have also used a variable coaxial line for the phase balancer in the second arm.

If the signal is applied, for example, to port 1, it divides into two equal parts, one passing through the parametric device and the other through the phase balancer, combining at port 2, where a magic tee is provided. The terminal 3 provides the third arm of the circulator and is terminated. When the phase shift of the pump was 90°, the transmission fron port 1 to port 2 was adjusted to add in phase at port 2. On the other hand, transmission from ports 2 to 1 under those conditions resulted in cancellation of the two signals, giving a minimum at terminal 1.

V. Experimental Results

A signal frequency of 500 mc and a pump frequency of 1500 mc were chosen. The up converter thus gave a 2000 mc upper sideband which in turn became the signal for the down converter parametric amplifier which again gave out 500 mc. Under perfect conditions, there should not be any power loss since the gain in the up converter is exactly the same as the loss in down converter parametric amplifier. We did get loss, however, which, we were able to reduce to less than one db. With the elaborate tuning arrangement, a bandwidth of 5 mc was observed.

When the whole unit was operated as an isolator, by loading the third port of the circulator, we were able to obtain an isolation of more than 46 db between the two directions.

VI. Discussion

The loss of less than one db was exceptional when one considers the number of elements, such as filters, etc., included in the experimental circuit. It is quite possible that under such conditions the converters might have acted as negative resistance amplifiers and provided gain.

This negative resistance behavior limits the bandwidth of the system and makes the phase relationships more complicated than those discussed above. To eliminate the lower sideband, the circuit has to be designed properly.

The other limiting factor for the bandwidth is the filter which is necessary in this device. Consequently, it is important to design a broad-band and low insertion loss filter. Matching elements should also be designed carefully so as not to limit the obtainable bandwidth.

One of the future efforts will be to broaden the band to 100 or 150 mc.

Appendix I

Calculation of Gain Function Denominator

$$\begin{bmatrix} V_g \\ I_g \end{bmatrix} = \begin{bmatrix} 1 & 1 \\ 0 & 1 \end{bmatrix} \begin{bmatrix} T_{11} + \delta T_{11} & i[T_{12} + \delta T_{12}] \\ i[T_{21} + \delta T_{21}] & T_{22} + \delta T_{22} \end{bmatrix} \cdot \begin{bmatrix} 1 & 0 \\ 1 & 1 \end{bmatrix} \begin{bmatrix} V_o \\ 0 \end{bmatrix} \tag{29}$$

where

$$T_{11} + \delta T_{11} = T_{22} + \delta T_{22}. \tag{30}$$

Since the determinant still remains unity,

$$(T_{11} + \delta T_{11})(T_{22} + \delta T_{22}) + (T_{12} + \delta_{12})(T_{21} + \delta T_{21}) = 1$$

which gives to first order in the variation of the matrix elements:

$$T_{11}\delta T_{22} + T_{22}\delta T_{11} + T_{12}\delta T_{21} + T_{21}\delta T_{12} = 0. \tag{31}$$

The insertion gain will become

$$\begin{aligned} G &= \frac{4}{|(T_{11}+T_{22}+\delta T_{11}+\delta T_{22})+i(T_{12}+T_{21}+\delta T_{12}+\delta T_{21})|^2} \\ &= \frac{4}{(T_{11}+T_{22}+\delta T_{11}+\delta T_{22})^2+(T_{12}+T_{21}+\delta T_{12}+\delta T_{21})^2} \\ &= \frac{4}{(T_{11}+T_{22})^2+(T_{12}+T_{21})^2+\delta(T_{11}^2+T_{22}^2+T_{12}^2+T_{21}^2)}. \end{aligned} \tag{32}$$

From (21),

$$(T_{11} + T_{22})^2 + (T_{12} + T_{21})^2 = 4.$$

Hence, (32) may be written in the form

$$G = \frac{4}{4 + \delta[(T_{11} - T_{22})^2 + (T_{12} - T_{21})^2 + 2(T_{11}T_{22} + T_{12}T_{21})]}. \tag{33}$$

$T_{11}T_{22}+T_{12}T_{21}$ is the determinant of the operator $[T]$ which is equal to unity; hence, the variation of the determinant is zero, which gives

$$G = \frac{4}{4 + \delta[(T_{11} - T_{22})^2 + (T_{12} - T_{21})^2]},$$

but $T_{11} = T_{22}$ under the assumptions we have made. Consequently,

$$\begin{aligned} G &= \frac{4}{4 + \delta(T_{12} - T_{21})^2} \\ &= \frac{1}{1 + \frac{1}{4}\delta(T_{12} - T_{21})^2} \end{aligned} \tag{34}$$

which finally gives

$$G = \frac{1}{1 + \frac{1}{4}(\delta\omega_s)^2\left[\frac{\partial(T_{12} - T_{21})}{\partial\omega_s}\right]^2}. \tag{35}$$

Appendix II

Derivation of $(T_{12}-T_{21})$

From earlier definitions in (11) and (15), one obtains

$$
\begin{aligned}
T_{12} - T_{21} &= 2\cos\theta\left[R_g\omega_s C_0\left(1-\frac{C_0^2}{|C_1|^2}\right)-\frac{C_0}{R_g\omega_s|C_1|^2}\right] \\
&\quad + \sin\theta\left[\frac{\omega_u}{\omega_s}\cdot\frac{C_0^2}{|C_1|^2}-\frac{\omega_s}{\omega_u}\cdot\frac{C_0^2}{|C_1|^2}-\frac{1}{R_g^2\Omega^2|C_1|^2}\right. \\
&\quad \left. + R_g^2\Omega^2|C_1|^2\left(1-\frac{C_0^2}{|C_1^2|}\right)^2\right].
\end{aligned} \tag{36}
$$

Taking the first derivative,

$$
\begin{aligned}
\frac{\partial}{\partial\omega_s}(T_{12}-T_{21}) &= -2\sin\theta\cdot\frac{d\theta}{d\omega_s}\cdot R_g\omega_s C_0(1-\gamma^2)+2\cos\theta R_g C_0(1-\gamma^2) \\
&\quad + 2\sin\theta\cdot\frac{d\theta}{d\omega_s}\,\frac{C_0}{R_g\omega_s|C_1|^2}+2\cos\theta\cdot\frac{C_0}{R_g\omega_s^2|C_1|^2} \\
&\quad + \sin\theta\cdot\gamma^2\left(\frac{\omega_s-\omega_u}{\omega_s^2}\right)+\cos\theta\cdot\frac{d\theta}{d\omega}\cdot\frac{\omega_u}{\omega_s}\gamma^2-\cos\theta\cdot\frac{d\theta}{d\omega}\cdot\frac{\omega_s}{\omega_u}\gamma^2 \\
&\quad - \sin\theta\cdot\gamma^2\left(\frac{\omega_u-\omega_s}{\omega_u^2}\right)-\cos\theta\cdot\frac{d\theta}{d\omega_s}\cdot\frac{1}{R_g^2\Omega^2|C_1|^2} \\
&\quad + \sin\theta\cdot\frac{2}{R_g^2\Omega^3|C_1|^2}\cdot\frac{d\Omega}{d\omega_s} \\
&\quad + \cos\theta\cdot\frac{d\theta}{d\omega}\cdot R_g^2\Omega^2|C_1|^2(1-\gamma^2)^2 \\
&\quad + 2\sin\theta\cdot R_g^2\Omega|C_1|^2(1-\gamma^2)^2\frac{d\Omega}{d\omega}
\end{aligned} \tag{37}
$$

where

$$\gamma^2=\frac{C_0^2}{|C_1|^2}\quad\text{and}\quad\frac{d\omega_u}{d\omega_s}=1.$$

Since

$$\Omega=(\omega_s\omega_u)^{1/2},$$

$$\frac{d\Omega}{d\omega_s}=\frac{1}{2\Omega}(\omega_s+\omega_u),$$

and

$$\theta=\frac{2\pi l}{\lambda_g}=\frac{2\pi}{\lambda_u}\left[1-\left(\frac{\lambda_u}{\lambda_c}\right)^2\right]^{1/2}l,$$

where l represents the length of coaxial line or waveguide, λ_g the guide wavelength, and λ_c the cut-off wavelength. Then

$$\frac{d\theta}{d\omega_s}=\left(\frac{d\theta}{d\lambda_u}\right)\left(\frac{d\lambda_u}{d\omega_s}\right)\Bigg|_{\text{at center frequency}}$$

$$=\theta\,\frac{\dfrac{f_u}{c}}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}\cdot\left(\frac{1}{2\pi}\cdot\frac{c}{f_u^2}\right),$$

as can easily be shown. Hence,

$$\frac{d\theta}{d\omega}=\frac{\alpha}{\omega_s}\cdot\frac{\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}\quad\text{where}\quad\alpha=\frac{\omega_s}{\omega_u}. \tag{38}$$

From the theory of the parametric amplifier,[4] one knows that $R_g^2\Omega^2|C_1|^2=1$, and, if we call $R_g\omega_s C_0=\beta$, then (37) can be written

$$
\begin{aligned}
&\left[\frac{\partial}{\partial\omega_s}(T_{12}-T_{21})\right]^2(\delta\omega_s)^2 \\
&= \frac{\cos^2\theta}{\omega_s^2}\left\{\left[-2\beta(1-\gamma^2)\frac{\alpha\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}+\frac{2\beta\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}\right.\right. \\
&\quad + \gamma^2\left(1-\frac{1}{\alpha}\right)-\gamma^2\alpha(1-\alpha) \\
&\quad \left. + (1+\alpha)+(1-\gamma^2)^2(1+\alpha)\right]\tan\theta \\
&\quad + 2\beta(1-\gamma^2)+\frac{2\beta}{\alpha}+\frac{\gamma^2}{\alpha}\left[\frac{\alpha\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}\right] \\
&\quad - \gamma^2\alpha\frac{\alpha\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}-\frac{\alpha\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}} \\
&\quad \left. + (1-\gamma^2)^2\left[\frac{\alpha\theta}{\left(1-\dfrac{f_c^2}{f_u^2}\right)^{1/2}}\right]\right\}^2(\delta\omega_s)^2.
\end{aligned} \tag{39}
$$

Appendix III

Calculation of Gain Function

Taking as the assumed values

$$\gamma^2=\frac{C_0^2}{|C_1|^2}=10,$$

$$\alpha=\frac{\omega_s}{\omega_u}=\frac{1}{4},$$

$$\beta=R_g\omega_s C_0=\sqrt{\beta^2}=\sqrt{\gamma^2\alpha}$$

since

$$R_g^2\omega_s\omega_u|C_1|^2=1=\sqrt{\frac{10}{2}}. \tag{40}$$

Before substituting the above values in (39), it will be necessary to find the value of θ, which is obtained at the matched condition, namely, from (17). That is,

$$2AB\cos\theta - B^2\sin\theta + A^2\sin\theta = 2CD\cos\theta + D^2\sin\theta - C^2\sin\theta,$$

which gives

$$\tan\theta = \frac{2[DC - AB]}{(A^2+C^2)-(B^2+D^2)}. \quad (41)$$

Substituting the values of A, B, C and D, we get

$$\tan\theta = \frac{2\left[-R_g\omega_s C_0\left(1-\frac{C_0^2}{|C_1|^2}\right)+\frac{C_0}{R_g\omega_s|C_1|^2}\right]}{\frac{\omega_u}{\omega_s}\cdot\frac{C_0^2}{|C_1|^2}+R_g^2\Omega^2|C_1|^2\left(1-\frac{C_0^2}{|C_1|^2}\right)^2-\frac{1}{R_g^2\Omega^2|C_1|^2}-\frac{\omega_s}{\omega_u}\cdot\frac{C_0^2}{|C_1|^2}}.$$

From (40) we obtain

$$\tan\theta = \frac{2\left[-\beta(1-\gamma^2)+\frac{\gamma^2}{\beta}\right]}{\frac{\gamma^2}{\alpha}+(1-\gamma^2)^2-1-\alpha\gamma^2} = \frac{1}{2.86}.$$

Hence,

$$\cos\theta = \frac{2.86}{3.04}.$$

Substituting these values, (39) becomes

$$\left[\frac{\partial}{\partial\omega_s}(T_{12}-T_{21})\right]^2(\delta\omega_s)^2 = \frac{\cos^2\theta}{\omega_s^2}\cdot(\delta\omega)^2\left\{\frac{\theta}{\left(1-\frac{f_c^2}{f_u^2}\right)^{1/2}}\right.$$

$$\cdot\left[\gamma^2-\gamma^2\alpha^2-\alpha+(1-\gamma^2)^2\alpha\right.$$

$$\left.-\frac{2\beta\alpha}{2.86}(1-\gamma^2)+\frac{2\beta}{2.86}\right]$$

$$+\frac{1}{2.86}\left[\gamma^2\left(1-\frac{1}{\alpha}\right)-\gamma^2\alpha(1-\alpha)\right.$$

$$\left.+1+\alpha+(1+\alpha)(1-\gamma^2)^2\right]$$

$$\left.+2\beta(1-\gamma^2)+\frac{2\beta}{\alpha}\right\}^2$$

$$=\frac{(\delta\omega)^2}{(\omega_s)^2}\cos^2\theta\left\{\frac{\theta}{\left(1-\frac{f_c^2}{f_u^2}\right)^{1/2}}(33.1)+\frac{70.72}{2.86}-15.61\right\}^2$$

$$=\left(\frac{\delta\omega}{\omega_s}\right)^2\cdot\left(\frac{2.86}{3.04}\right)^2\left\{\frac{33.1}{\left(1-\frac{f_c^2}{f_u^2}\right)^{1/2}}+9\right\}^2. \quad (42)$$

For a coaxial line, (42) becomes

$$\left[\frac{\partial}{\partial\omega_s}(T_{12}-T_{21})^2\right](\delta\omega_s)^2 = \left(\frac{\delta\omega}{\omega_s}\right)^2\left(\frac{2.86}{3.04}\right)^2\{10.2+9\}^2 \quad (43)$$

since

$$\frac{f_c^2}{f_u^2}=0.$$

For a waveguide, putting $f_c^2/f_u^2=\frac{1}{2}$, we obtain

$$=\left(\frac{\delta\omega}{\omega_s}\right)^2\left(\frac{2.86}{3.04}\right)^2\left\{\frac{10.2}{(\frac{1}{2})^{1/2}}+9\right\}^2. \quad (44)$$

Solving (44), we get

$$\left[\frac{\partial}{\partial\omega_s}(T_{12}-T_{21})\right]^2(\delta\omega_s)^2 = (490)\left(\frac{\partial\omega}{\omega_s}\right)^2.$$

Hence, the gain by (23) becomes

$$G = \frac{1}{1+122.6\left(\frac{\delta\omega}{\omega_s}\right)^2}. \quad (45)$$

Acknowledgment

The author is indebted to H. Seidel for suggesting this problem as well as certain analytical techniques, to L. Wetzel for many helpful discussions, and to F. J. D'Allessio for carrying out the experiments.

The Compatibility Problem in Single-Sideband Transmission*

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Summary—Under the assumption of simultaneous amplitude and phase modulation of a carrier, a study is made of the relations that must hold between the envelope and the phase of a single-sideband wave. In particular, it is shown that absolute compatibility with standard AM receivers employing a linear envelope detector cannot possibly be achieved with the spectral economy of conventional single sideband. On the other hand, if one conveys the message function in the square of the envelope rather than in the envelope itself, it is shown that a phase function can be found for which the hybrid wave occupies a spectral width equal to that of a conventional single-sideband (SSB) system. Distortionless detection is achieved with a square-law envelope detector. The operations required to generate this square-law SSB signal are described in detail.

Introduction

It is well known that a single-sideband signal, generated by any conventional single-sideband technique, can be regarded as the resultant of quadrature modulation of a carrier by a pair of signals in phase quadrature. If $s(t)$ is an arbitrary message function, and $\sigma(t)$ is its *harmonic conjugate* [obtained from $s(t)$, ideally, by a network whose amplitude response is unity, and whose phase response is a constant 90° lag at all frequencies], then it follows that the modulated wave

$$f(t) = s(t) \cos \omega_0 t - \sigma(t) \sin \omega_0 t \tag{1}$$

is an upper-sideband signal with no spectral components below the carrier angular frequency ω_0. In fact, any band-limited wave (*e.g.*, band-pass noise) can be expressed in the form of (1), if ω_0 is chosen to be below the band edge.

We can write (1) in the form

$$\begin{aligned} f(t) &= \sqrt{s^2(t) + \sigma^2(t)} \cos\left[\omega_0 t + \tan^{-1} \frac{\sigma(t)}{s(t)}\right] \\ &= \alpha(t) \cos [\omega_0 t + \phi(t)], \end{aligned} \tag{2}$$

regarding the single-sideband signal as a hybrid amplitude- and phase-modulated wave. In a conventional SSB system, the message is normally conveyed by the in-phase component $s(t)$, detection being accomplished by synchronous demodulation with a carrier generated locally at the receiver. It is conceivable, however, that the message could be conveyed by the envelope $\alpha(t)$, if a corresponding phase $\phi(t)$ can be found that yields a single-sided spectrum for the hybrid wave. Reception would then be accomplished by a conventional envelope detector, and a single-sideband transmission could be made compatible with standard AM receivers.[1,2]

Since the signals $s(t)$ and $\sigma(t)$ are uniquely related to one another, it is apparent that the envelope $\alpha(t)$ and the phase $\phi(t)$ must be related also. The compatibility problem consists in the study of conditions on the envelope $\alpha(t)$ that yield to a realizable phase function $\phi(t)$ and in determining the relation that must hold between them. The solution is most easily obtained through the concept of the *analytic signal*.[3]

Mathematically, the function $\sigma(t)$ is related to $s(t)$ by the Hilbert transform[4] relation,

$$\begin{aligned} \sigma(t) &= \frac{1}{\pi} \int_{-\infty}^{\infty} \frac{s(\tau)}{t - \tau} d\tau \\ &= \frac{1}{\pi} \int_{0+}^{\infty} \frac{s(t - \tau) - s(t + \tau)}{\tau} d\tau. \end{aligned} \tag{3}$$

If the spectrum $S(\omega)$ is derived from $s(t)$ through the Fourier integral

$$S(\omega) = \int_{-\infty}^{\infty} s(t) e^{-i\omega t} dt, \tag{4}$$

then it follows directly from (3), (4), and the inverse Fourier integral that the spectrum $\sum(\omega)$ of the harmonic conjugate function $\sigma(t)$ is given by

$$\sum(\omega) = \begin{cases} -iS(\omega) & \omega > 0 \\ 0 & \omega = 0 \\ iS(\omega) & \omega < 0. \end{cases} \tag{5}$$

Thus for positive frequencies, the spectrum of $\sigma(t)$ is identical to that of $s(t)$ except for the multiplying factor $(-i)$ corresponding to a 90° phase lag.

It can further be shown that the spectrum $F(\omega)$ of the single-sideband wave of (1) is given by

$$F(\omega) = \begin{cases} 0 & |\omega| < \omega_0 \\ S(|\omega| - \omega_0) & |\omega| > \omega_0. \end{cases} \tag{6}$$

* Received by the IRE, August 2, 1959; revised manuscript received, March 28, 1960. The research for this paper has been sponsored in part by the Electronic Res. Directorate, AF Cambridge Res. Center, Air Res. and Dev. Command, under Contract No. AF 19(604)-4976.

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If the spectrum of $s(t)$ is limited to a base-bandwidth of W cycles per second, then the signal $f(t)$ occupies the same bandwidth above the carrier frequency $\omega_0/2\pi$ cps.

Now if we define a complex signal $\psi(t)$ by

$$\begin{aligned}\psi(t) &= s(t) + i\sigma(t)\\ &= \alpha(t)e^{i\phi(t)},\end{aligned} \tag{7}$$

it follows that the spectrum $\Psi(\omega)$ of $\psi(t)$ becomes

$$\Psi(\omega) = \begin{cases} 2S(\omega) & \omega > 0 \\ 0 & \omega < 0, \end{cases} \tag{8}$$

and is characterized by the property that it *vanishes* for negative frequencies. The nonphysical signal $\psi(t)$ has been called the *analytic* signal by some authors[3] because as a function of a complex variable z, $\psi(z)$ is analytic in the upper-half z plane. By comparing (2) and (7), it can be seen that the magnitude and phase of the complex signal $\psi(t)$ are identical to the envelope and phase of the single-sideband wave. Thus a study of single-sideband can be made through the analytic signal without reference to the arbitrary carrier frequency ω_0.

In order to study the compatibility problem in SSB communication, one needs only to study the relationships among the real and imaginary parts, magnitude and phase of the complex function $\psi(t)$, to obtain the same relationships between the in-phase and quadrature components, envelope and phase of the single-sideband wave.

The Compatibility Problem

Let us turn now to the first important result in this study, the spectrum of the squared envelope. From (7) we have

$$\alpha^2(t) = |\psi(t)|^2. \tag{9}$$

By taking the Fourier transform of both sides,

$$\begin{aligned}\int_{-\infty}^{\infty} \alpha^2(t)e^{-i\omega t}dt &= \int_{-\infty}^{\infty} |\psi(t)|^2 e^{-i\omega t}dt\\ &= \frac{1}{2\pi}\int_{-\infty}^{\infty}\psi(t)e^{-i\omega t}\int_{-\infty}^{\infty}\Psi^*(\lambda)e^{-i\lambda t}d\lambda dt\\ &= \frac{1}{2\pi}\int_{-\infty}^{\infty}\Psi^*(\lambda)\int_{-\infty}^{\infty}\psi(t)e^{-i(\lambda+\omega)t}dtd\lambda\\ &= \frac{1}{2\pi}\int_{-\infty}^{\infty}\Psi^*(\lambda)\Psi(\lambda+\omega)d\lambda\\ &= \frac{2}{\pi}\int_{0}^{\infty} S^*(\lambda)S(\lambda+\omega)d\lambda,\ \omega > 0.\end{aligned} \tag{10}$$

Now if we assume the signal $\psi(t)$ [hence also $s(t)$] to be limited to a bandwidth of W cps, then

$$\int_{-\infty}^{\infty}\alpha^2(t)e^{-i\omega t}dt = \frac{2}{\pi}\int_{0}^{(2\pi W-\omega)} S^*(\lambda)S(\lambda+\omega)d\lambda,\ \omega > 0. \tag{11}$$

The right-hand side vanishes for $\omega > 2\pi W$. We have thus established:

Theorem 1: The bandwidth of the square of the envelope of a single-sideband wave (or any band-pass limited wave) is equal to the spectral width of that wave.

A similar result was obtained by Rice[5] in his study of noise through a square-law device. The theorem in the above form was proved by Dugundji.[6] It should be remarked that no statement can be made concerning the bandwidth of the envelope itself since, in general, the envelope is not bandlimited.

However, if the envelope is constrained by compatibility to be of bandwidth W, it is clear that the envelope square is of bandwidth $2W$, hence the spectral width of a bandlimited hybrid wave must also be $2W$. In other words, if a given envelope is bandlimited to W, then the spectral width of a hybrid wave with any arbitrary phase modulating function must equal or exceed $2W$. Thus we have the important result that *absolute compatibility with an AM receiver employing a linear envelope detector cannot possibly be achieved with the spectral economy of conventional single-sideband.* Costas[2] came to this conclusion by simple reasoning with the case of single-tone modulation.

Although this result precludes the possibility of a truly compatible SSB system, it still includes the possibility of an SSB system employing a simple asynchronous receiver. If we accept the use of a square-law detector at the receiver, the theorem suggests that a technique may be found to generate an SSB signal whose envelope is the square root of the message function to be transmitted.

It is an inversion of Theorem 1 that is required for the existence of such a system. Given a message function $m(t)$ of bandwidth W cps, is it possible to generate a modulated wave occupying a spectral width W cps in such a way that $m(t)$ is the square of the envelope of that wave? The following considerations indicate an affirmative answer.

We shall write the intelligence signal $m(t)$ as $\alpha^2(t)$, since it is to be the square of the envelope of the modulated wave. The above question can be posed in the following manner. What are the conditions on $\alpha^2(t)$ that are both necessary and sufficient for the existence of a phase function $\phi(t)$ for which simultaneous amplitude and phase modulation of a carrier by $\alpha(t)$ and $\phi(t)$ produces a waveform with spectral width equal to that of $\alpha^2(t)$? Clearly a necessary condition is that $\alpha^2(t)$ be nonnegative. Further conditions, both necessary and sufficient, are found in the following theorem of Paley and Wiener:[7]

[5] S. O. Rice, "Mathematical analysis of random noise," *Bell Sys. Tech. J.*, vol. 24, pp. 125–131; January, 1945.

[6] J. Dugundji, "Envelopes and pre-envelopes of real waveforms," IRE Trans. on Information Theory, vol. IT-4, pp. 53–57; March, 1958.

[7] R. E. A. C. Paley and N. Wiener, "Fourier Transforms in the Complex Domain," Amer. Math. Soc. Colloquium Publications, New York, N. Y., vol. 19, pp. 16–20; 1934.

Theorem 2 (Paley and Wiener): Let $\alpha(t)$ be a real non-negative function, integrable square in $(-\infty, \infty)$. Then a necessary and sufficient condition for the existence of a complex function $\psi(t)$ whose Fourier transform vanishes for negative argument and $|\psi(t)| = \alpha(t)$, is that

$$\int_{-\infty}^{\infty} \frac{|\log \alpha(t)|}{1+t^2}\, dt < \infty.$$

It is clear that if $\alpha^2(t)$ is non-negative and satisfies the logarithmic integral condition, then $\alpha(t)$ will also satisfy the condition, and the theorem applies. The logarithmic integral converges 1) if $\alpha(t)$ does not vanish identically over any interval or, more generally, 2) if $\alpha(t)$ falls to zero slower than an exponential decay of arbitrary exponent. Condition 1) is required for the *existence* of a proper phase function. Obviously if the envelope of a modulated wave is zero during a finite interval, the phase of the wave during that interval is indeterminate. Satisfaction of condition 2) [which incidentally, implies satisfaction of condition 1)] is guaranteed if $\alpha(t)$ is the output of any physically realizable network. It is, in fact, guaranteed by the bandlimitedness of either $\alpha(t)$ or its square. Also, since the exponential decay can be as fast as one chooses with an arbitrary choice of exponent there is, in effect, no limit on the decay time of $\alpha(t)$ except as imposed by a limited bandwidth condition. Thus the only condition required by the theorem that has real physical significance is that $\alpha(t)$ be non-negative. Since an envelope is by definition non-negative, this condition is obvious.

The proof of the foregoing theorem can be found in the book of Paley and Wiener. We reproduce in the Appendix, however, a simple proof of the sufficiency part of the theorem since the desired relationship between the envelope $\alpha(t)$ and the phase $\phi(t)$ are brought out in the course of that proof. It is shown in the Appendix that if the phase $\phi(t)$ is chosen to be the Hilbert Transform of the logarithm of the envelope, that is, if

$$\phi(t) = \frac{1}{\pi}\int_{-\infty}^{\infty} \frac{\log \alpha(\mu)}{t-\mu}\, d\mu, \qquad (12)$$

then the function $\psi(t) = \alpha(t) \exp [i\phi(t)]$ is an analytic signal whose spectrum vanishes for negative argument.

From (10), we have

$$\frac{1}{2\pi}\int_{0}^{\infty} \Psi^*(\lambda)\Psi(\lambda+\omega)d\lambda = \int_{-\infty}^{\infty} \alpha^2(t)e^{-i\omega t}dt. \qquad (13)$$

If the spectrum of the intelligence signal $\alpha^2(t)$ is assumed to vanish for $|\omega| > 2\pi W$, then both sides of (13) must so vanish. This is possible only if the spectrum $\Psi(\omega)$ also vanishes for $\omega > 2\pi W$. We have thus established:

Theorem 3: Given a real non-negative function $\alpha^2(t)$ satisfying the conditions of Theorem 2, and, in addition, having a Fourier transform that vanishes for $|\omega| > 2\pi W$, the function

$$\psi(t) = \alpha(t) \exp \left[\frac{i}{\pi}\int_{-\infty}^{\infty} \frac{\log \alpha(\mu)}{t-\mu}\, d\mu \right]$$

has a Fourier transform that vanishes outside the interval $0 < \omega < 2\pi W$.

The implication of Theorem 3 is as follows. Given a non-negative message function (satisfying certain integrability conditions) of bandwidth W, and a carrier signal at some arbitrary frequency, then simultaneous amplitude and phase modulation of that carrier with an envelope which is the *square root* of the message function, and a phase that is in quadrature with the *logarithm* of that envelope, produces a hybrid wave occupying a band above the carrier frequency of spectral width exactly equal to that of the message function. Distortionless reception of such a signal is achieved with a square-law envelope detector. It is of interest to note that even though the squared envelope is bandlimited, neither the enevelope itself nor the phase function are necessarily bandlimited, yet the hybrid modulated wave generated by these is limited to the same bandwidth as the envelope square.

A Square-Law SSB System

A complete transmitter for generating a hybrid wave that conveys a message in the square of the envelope and yet has the spectral economy of conventional single sideband is shown in the block diagram of Fig. 1. The message function $m(t)$ is first made positive to produce the envelope-squared $\alpha^2(t)$. Although this operation can be accomplished simply by adding a constant equal in magnitude to the largest negative peak of the message function, a substantial improvement in efficiency is realized for speech transmission by the process depicted in Fig. 2. The speech signal is rectified and filtered to obtain a negative slowly-varying envelope. This envelope is inverted and added to (or subtracted from) the original speech waveform to yield the non-negative waveform of Fig. 2(b). No serious loss of intelligibility results since this process simply adds a low-frequency, virtually inaudible signal to the original speech waveform. In the transmitted wave, this non-negative process is equivalent to a controlled-carrier operation.

The non-negative signal $\alpha^2(t)$ is fed to a nonlinear device whose output is proportional to the logarithm of the input. With an appropriate gain adjustment, the signal $\log \alpha(t)$ is available for Hilbert transformation by means of a wide-band 90° phase-splitting network. The two outputs of the phase-splitter are conjugate harmonic functions, one a delayed replica of the input and the other lagging in phase from the first by a constant 90° at all frequencies. The conventional RC phase-splitters of the phasing system of SSB are not adequate for this system because of severe phase distortion. One of the outputs must preserve the input waveform because of the nonlinear operation that has been performed. A phase-splitting network that does preserve this waveform is described in detail below.

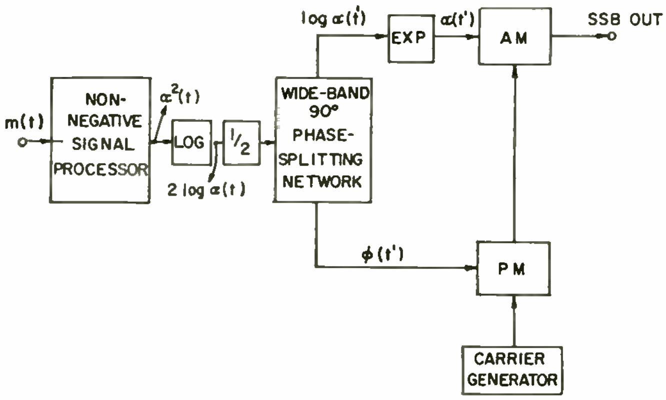
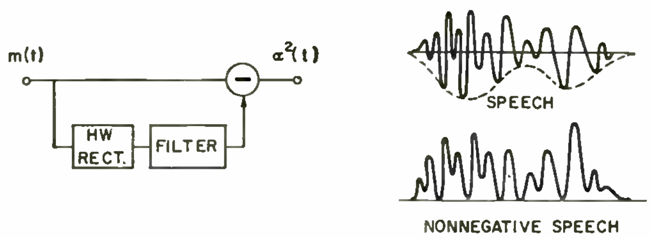


Fig. 1—The square-law SSB system.



(a) (b)

Fig. 2—Non-negative signal processor for speech.

The phase-shifted output $\phi(t')$ represents the instantaneous phase signal for the hybrid wave since it is the harmonic conjugate of the logarithm of the envelope. The delayed output log $\alpha(t')$ is passed through another nonlinear device with an exponential characteristic to recover the envelope $\alpha(t')$. An arbitrary carrier is first modulated in phase by the signal $\phi(t')$, and the PM wave is further modulated in amplitude by $\alpha(t')$ to produce the single-sideband output.

The Wide-Band 90° Phase-Shift Network

So far we have tacitly assumed that the wide-band 90° phase-shifting network is physically realizable. Although a constant amplitude response and constant phase shift is not a physically realizable characteristic exactly, it can be approximated arbitrarily closely with sufficient delay. If $s(t)$ is the input to such a filter, the output $\sigma(t)$ at any instant t is seen from (3) to depend upon both $s(t-\tau)$ and $s(t+\tau)$ for all positive τ. But $s(t+\tau)$ represents *future* values of the input. Thus the exact 90° phase shifter is a clairvoyant network. The dependence on future values is of lesser degree as τ increases because of the τ in the denominator of (3). We can thus make finite the upper limit of the integral, and can achieve as small an error as desired by setting that limit sufficiently large.

Now if the network input $s(t)$ is bandlimited, it is completely defined by its samples at the Nyquist interval, hence the output $\sigma(t)$ can be made to depend upon only a finite number of sample values of the input. These sample values can be obtained from taps on a delay line as in Fig. 3. The delay line is center tapped to

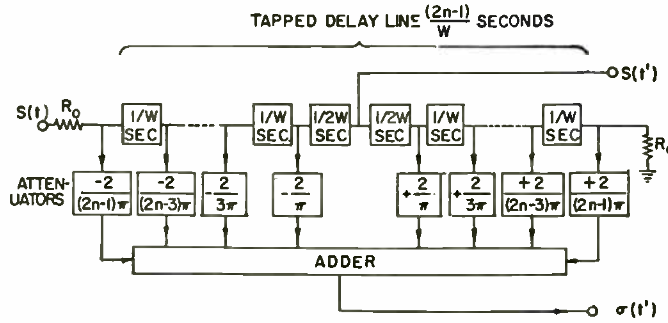


Fig. 3—The 90° phase-splitting network.

provide a delayed replica of the input. If we denote by $s(t')$ the present value of the signal at the center tap, then the signals at taps toward the input represent future values of $s(t')$ while those toward the termination represent past values. By the proper choice of a linear combination of the values at the delay line taps, the network of Fig. 3 provides a close approximation to the Hilbert transform relation of (3).

The network provides two output signals, the one at the delay line center tap representing a delayed replica of the input signal. The transfer characteristic from the input to the center tap is thus a unit amplitude response at all frequencies and a phase response that is linear with slope $(2n-1)/2W$, intercepting zero degrees at zero frequency. This response is shown by the dashed lines of Fig. 4. The transfer characteristic from the input to the adder output can be shown to be

$$H_n(\omega) = \exp\left\{-i\left[\frac{2n-1}{2W}\omega + \frac{\pi}{2}\right]\right\} \cdot \frac{4}{\pi}\sum_{k=1}^{n}\frac{1}{2k-1}\sin\left(\frac{2k-1}{2W}\right)\omega. \qquad (14)$$

The exponential term indicates a linear phase shift intercepting 90°, while the summation can be seen to be an amplitude response represented by a partial sum of the Fourier series expansion for the constant one. Thus the two output signals differ in phase by precisely 90° at all frequencies, while maintaining approximately equal magnitudes of their Fourier components. The amplitude and phase response for a network with $n=13$ is shown by the solid lines of Fig. 4.

It is important that the input signal be properly bandlimited, because the amplitude response of the phase shifter is periodic in frequency. The amplitude response between W and $2W$ is identical to that between 0 and W.

Conclusions

In the case of an average power limitation in the final RF stage of the transmitter, the square-law system operates at a disadvantage relative to conventional SSB. In fact, the average power in the "controlled-carrier" resulting from the slowly-varying envelope of the speech waveform is obviously excess power over that required for suppressed-carrier SSB operation.

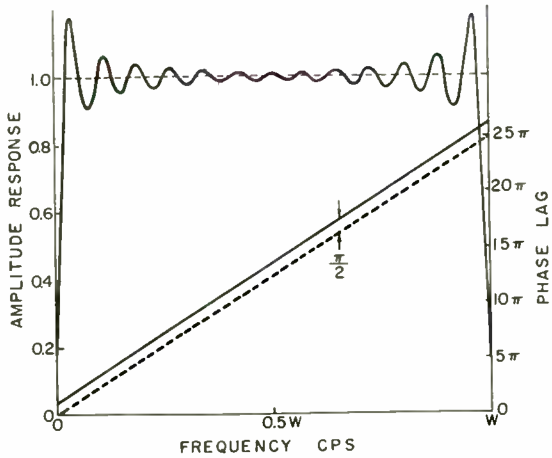


Fig. 4—Amplitude and phase response for 90° phase splitter.

However, if the final stage is peak-power limited, the square-law system can be used effectively, even when clipped speech is being transmitted. It is well known that SSB suffers from a pronounced peaking factor when operating with compressed waveforms. In the square-law system, the envelope is the *square root* of the intelligence and has an even lower peak-to-rms ratio than the original message waveform. Thus, for a given peak power, the final stage can be operated at a higher average power in the square-law system.

An additional advantage of the square-law system is its capability of transmitting analog data waveforms in the same bandwidth as conventional SSB, without the requirement of precise phase synchronization for demodulation at the receiver. In fact, to transmit such waveforms by conventional SSB would require the equivalent of a phasing system employing a phase distortionless phase splitter such as the one described above. At the present state of the art, the filter method is not promising for this application because of the requirement of a sideband filter with good phase linearity *measured from the frequency of the virtual carrier.* Although available sideband filters have a reasonable phase linearity in the center of the pass band, the linearity in the cutoff regions is particularly poor.

Distortionless detection of the square-law signal is achieved only with a square-law envelope detector at the receiver. In the case of speech transmission, however, the use of a linear envelope detector will result in no serious loss of intelligibility. Thus, in a sense, compatibility with standard AM receivers is achieved.

Appendix

Proof of Sufficiency of Theorem 2

Assuming the convergence of the logarithmic integral, let us set $z=t+i\tau$, and define the function $\Gamma(z)$ for $\tau>0$ by

$$\Gamma(z) = \Gamma(t+i\tau) = \frac{1}{i\pi}\int_{-\infty}^{\infty}\frac{\log\alpha(\mu)}{\mu-(t+i\tau)}\,d\mu$$

$$= \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{\tau\log\alpha(\mu)}{(t-\mu)^2+\tau^2}\,d\mu + i\,\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{(t-\mu)\log\alpha(\mu)}{(t-\mu)^2+\tau^2}\,d\mu. \tag{15}$$

If we take the limit as τ approaches zero from above, the function $\tau/\pi[(t-\mu)^2+\tau^2]$ approaches a unit impulse function centered at $\mu=t$ and

$$\lim_{\tau\to 0+}\Gamma(t+i\tau) = \log\alpha(t) + i\,\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{\log\alpha(\mu)}{t-\mu}\,d\mu. \tag{16}$$

The imaginary part of this expression is seen to be the Hilbert transform of the real part, $\log\alpha(t)$. If we form the function $\psi(z)=\exp\Gamma(z)$, and define $\psi(t)$ by

$$\psi(t) = \lim_{\tau\to 0+}\exp\Gamma(t+i\tau) = \alpha(t)\exp\left[\frac{i}{\pi}\int_{-\infty}^{\infty}\frac{\log\alpha(\mu)}{t-\mu}\,d\mu\right], \tag{17}$$

then

$$|\psi(t)| = \alpha(t), \tag{18}$$

and we have only to show that the Fourier transform of $\psi(t)$ vanishes for negative argument. From the inequality between the arithmetic and geometric means, we have

$$|\psi(t+i\tau)| = \exp\frac{1}{\pi}\int_{-\infty}^{\infty}\frac{\tau\log\alpha(\mu)}{(t-\mu)^2+\tau^2}\,d\mu \le \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{\tau\alpha(\mu)}{(t-\mu)^2+\tau^2}\,d\mu, \tag{19}$$

and from the Schwartz inequality,

$$|\psi(t+i\tau)|^2 \le \frac{1}{\pi}\int_{-\infty}^{\infty}\frac{\tau\alpha^2(\mu)}{(t-\mu)^2+\tau^2}\,d\mu. \tag{20}$$

Thus for $\tau>0$,

$$\int_{-\infty}^{\infty}|\psi(t+i\tau)|^2dt \le \frac{1}{\pi}\int_{-\infty}^{\infty}\alpha^2(\mu)\int_{-\infty}^{\infty}\frac{\tau dt}{(t-\mu)^2+\tau^2}\,d\mu = \int_{-\infty}^{\infty}\alpha^2(\mu)d\mu, \tag{21}$$

and $\psi(t+i\tau)$ is boundedly integrable square on every line $\tau>0$. Now since the Fourier transform of $\psi(t+i\tau)$ is given by $\Psi(\omega)e^{-\omega\tau}$, we obtain from Parseval's relation that

$$\frac{1}{2\pi}\int_{-\infty}^{\infty}|\Psi(\omega)|^2e^{-2\omega\tau}d\omega = \int_{-\infty}^{\infty}|\psi(t+i\tau)|^2\,dt \le \int_{-\infty}^{\infty}\alpha^2(\mu)d\mu \tag{22}$$

for every $\tau>0$. Since the left-hand side is bounded for every positive τ, it follows that the spectrum $\Psi(\omega)$ must vanish for negative ω, and the proof of sufficiency is complete.

On the Resolving Time and Flipping Time of Magnetoresistive Flip-Flops*

A. AHARONI† AND E. H. FREI‡, MEMBER, IRE

Summary—**The time constant of the exponential approach to the stable states is shown to be a fair approximation for the flipping time of the "bridge" magnetoresistive flip-flop. This time constant turns out to be at least of the order of milliseconds for the magnetoresistive materials known at present, which is too large. Similar time constants are used as an approximation for the resolving times of a general nonlinear network, in particular networks in which the nonlinear element is a magnetoresistor. It is shown that these time constants are the latent roots of a certain matrix whose elements can readily be calculated from the parameters of the network.**

The flipping time of the "bridge" magnetoresistive flip-flop is calculated as a function of the energy supplied by the incoming pulse. The calculation is made by the numerical solution of the nonlinear differential equation involved. The results show that for input pulses of conceivable amplitudes the linear time constant is a good measure of the flipping time.

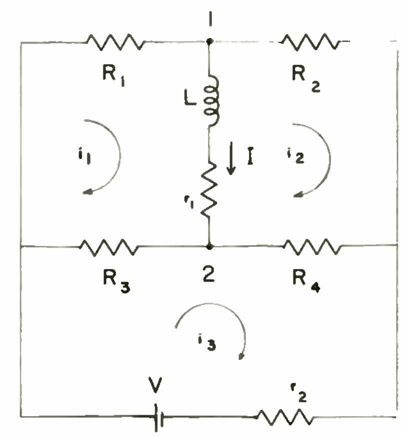


Fig. 1—Basic circuit of the "bridge" magnetoresistive flip-flop.

I. Introduction

In a previous paper[1] the authors discussed the possibility of constructing a magnetoresistive flip-flop using the "bridge" circuit of Fig. 1. Obviously the parameters must be chosen in such a way that the change of the magnetic field produced by the changing current and the resulting change of resistance in the magnetoresistive material produces an *amplified* change of current. In order to obtain an efficient element, the magnetoresistive effect must be as large as possible. An arrangement giving larger fields with larger currents will "drive" the element stronger, but will also give higher inductance and lower the speed. It was shown[1] that the current I in the coil L is the solution of

$$-L\frac{dI}{dt} = \frac{(r_1 + R - MV)I + [Nr_1 + R(2N - M^2)]I^3 + N^2RI^5}{1 + NI^2} \tag{1}$$

if one assumes for simplicity that r_2 of Fig. 1 can be neglected when compared to the resistors of this circuit, and that $R_4 = R_1$, $R_3 = R_2$. Here,

$$M = 2CKH_0/(1 + CH_0^2), \quad N = CK^2/(1 + CH_0^2),$$
$$R = R_0(1 + CH_0^2) \tag{2}$$

where R_0 is the resistance of R_1 and R_2 in zero magnetic field, C is the magnetoresistive coefficient, H_0 is a constant magnetic field applied to all the resistors and K is the constant of proportionality between the current I and the magnetic field it produces.

It was also shown that whenever

$$MV > r_1 + R \tag{3}$$

the circuit has (besides the unstable equilibrium $I = 0$) two stable equilibrium states, $\pm I_0$, which are the roots of

$$r_1 + R - MV + [Nr_1 + R(2N - M^2)]I_0^2 + N^2RI_0^4 = 0. \tag{4}$$

The flipping time, on the other hand, was not calculated, but was assumed to be of the order of L/R. It will be shown here that in many cases this is not a good approximation, and the actual flipping time is much larger. For the sake of a more rigorous calculation, another time constant is introduced in Part II and is shown (in Parts III, IX and X) to be a good approximation. In Part IV this time constant is referred to the power dissipated by R_1 and R_2 and is shown (in Part V) to be at least of the order of 10^{-3} seconds for conceivable values of this power. In Parts VI and VII a general network theorem concerning the evaluation of time constant of networks containing magnetoresistive elements will be proven. In Part VIII the dependence of the flipping time on the incoming pulse is considered.

* Received by the IRE, June, 1959; revised manuscript received, January 4, 1960. This paper is part of a thesis submitted by A. Aharoni to the Hebrew University, Jerusalem, in partial fulfilment of the requirements for the Ph.D. degree.

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[1] A. Aharoni, E. H. Frei, and G. Horowitz, "A new active circuit element using the magnetoresistive effect," *J. Appl. Phys.*, vol. 26, pp. 1411–1415; December, 1955.

For the purpose of this paper, the following notation will be used:

$$\sigma = (r_1 + R)/R; \qquad \lambda = 4CH_0^2/(1 + CH_0^2) \tag{5}$$

$$m = MV/R - \sigma; \qquad n = \sigma + 1 - \lambda \tag{6}$$

where m is a measure of the "amplification" property[2] of the element.

Using (2) and definition (5), one obtains

$$\begin{aligned} M^2 &= 4C^2K^2H_0^2/(1 + CH_0^2)^2 \\ &= 4NCH_0^2/(1 + CH_0^2) = N\lambda. \end{aligned} \tag{7}$$

Substituting (5), (6), and (7) in (1) yields

$$-(L/R)dI/dt = (-mI + NnI^3 + N^2I^5)/(1 + NI^2) \tag{8}$$

while substituting the same in (4) gives

$$-m + nNI_0^2 + N^2I_0^4 = 0. \tag{9}$$

Since the coefficient C is positive, as are all the resistances involved, it is clear from the definition (5) that

$$1 < \sigma < \infty; \qquad 0 < \lambda < 4 \tag{10}$$

while the substitution of (5) and (6) in (3) gives

$$m > 0 \tag{11}$$

and the substitution of (10) in (6) gives

$$-2 < n < \infty. \tag{12}$$

II. The Time-Constant τ

Consider the behavior of I near a stable state I_0, *i.e.*, when $I = I_0 + i$ and i is small enough for its higher powers to be neglected. In this case (8) reduces to

$$\begin{aligned} -\frac{L}{R}\frac{dI}{dt} &= -\frac{L}{R}\frac{di}{dt} \\ &\approx \frac{-m(I_0 + i) + nN(I_0^3 + 3I_0^2 i) + N^2(I_0^5 + 5I_0^4 i)}{1 + N(I_0^2) + 2I_0 i)} \end{aligned}$$

which together with (9) gives

$$-(L/R)di/dt \approx (2nNI_0^2 + 4N^2I_0^4)i/(1 + NI_0^2).$$

This means that i vanishes exponentially, $i = i_0 e^{-t/\tau}$, with the time constant

$$\tau = L(1 + NI_0^2)/2RNI_0^2(n + 2NI_0^2). \tag{13}$$

It should be noted that the time constant is always positive so that i vanishes and the element decays into the stable equilibrium. This is assured by (11) which together with (9) gives

$$n + NI_0^2 > 0,$$

the other terms in τ being positive by definition.

[2] A. Aharoni and E. H. Frei, "Remarks on magneto-resistive amplifiers," *Bull. Research Council of Israel*, vol. 5A, pp. 240–241; April, 1956.

When transferring from one stable state to the other, the terminal epoch is thus an exponential decay with the time constant given by (13). It therefore seems conceivable that the flipping time will be at least of the order of this τ. In the next part a more rigorous justification of this will be given.

III. The Flipping Time

Let

$$x = x(t) = NI^2 \tag{14a}$$

be substituted in (8). This yields

$$\begin{aligned} -(L/2R)dx/dt &= x(-m + nx + x^2)/(1 + x) \\ &= x(x - a)(x + b)/(1 + x) \end{aligned} \tag{14b}$$

where

$$\begin{aligned} a &= \tfrac{1}{2}[-n + (n^2 + 4m)^{1/2}] > 0, \\ b &= \tfrac{1}{2}[n + (n^2 + 4m)^{1/2}] > 0. \end{aligned} \tag{15}$$

Comparing (15) with (9), one can also write

$$a = NI_0^2 \tag{16}$$

Suppose now that the element is somehow brought to a point x_1, past the unstable equilibrium $x = 0$ and is allowed to decay to a point x_2 approaching the stable states $x = a$. By (14), the time this decay will take is

$$\begin{aligned} t &= \frac{L}{2R}\int_{x_1}^{x_2} \frac{(1 + x)dx}{x(x + b)(a - x)} \\ &= \frac{L}{2R}\left[\frac{1 - b}{b(a - b)}\log\frac{x}{x + b} + \frac{1 + a}{a(a + b)}\log\frac{x}{a - x}\right]_{x_1}^{x_2} \end{aligned}$$

Using (13), (15) and (16) one can rewrite this in the form

$$\frac{t}{\tau} = \left[\log\frac{x}{a - x} + \frac{a(1 - b)}{b(1 + a)}\log\frac{x}{x + b}\right]_{x_1}^{x_2}. \tag{17}$$

Figs. 2 through 5 are a plot of this t/τ as a function of the various parameters involved, namely m, n, x_1 and x_2. Figs. 2 and 3 give the time it takes the current I to reach a given fraction of its stable value I_0, with $I = 0.1\ I_0$ as a starting point. Fig. 2 is drawn for $n = 1$ and with m as a varying parameter, while Fig. 3 is drawn for $m = 1$ with n as a varying parameter. Fig. 4 shows the dependence of the time on the starting point, for $m = 1$, $n = 0$. Fig. 5 is a plot of the time T it takes the element to change I from 0.1 to 0.9 of its stable value. It will be noted that, whereas this time has a minimum as a function of n, it is a monotonic decreasing function of m. However, larger values for m than those given in the above figures are of little interest because a large m actually means a large amplification factor of the active element, and according to our present experience[2] even the value of $m = 100$ is much beyond what might be realized in practice. Besides this, for

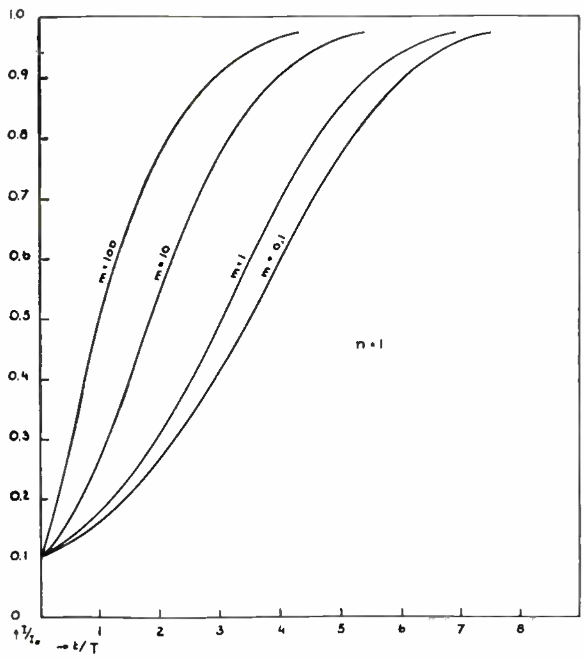

Fig. 2—Decay of the current I in the coil into its stable state I_0 for the case $n=1$ and various values of m. The time scale is in terms of the linear time constant τ, defined in the text. The starting point is $I=0.1\ I_0$.

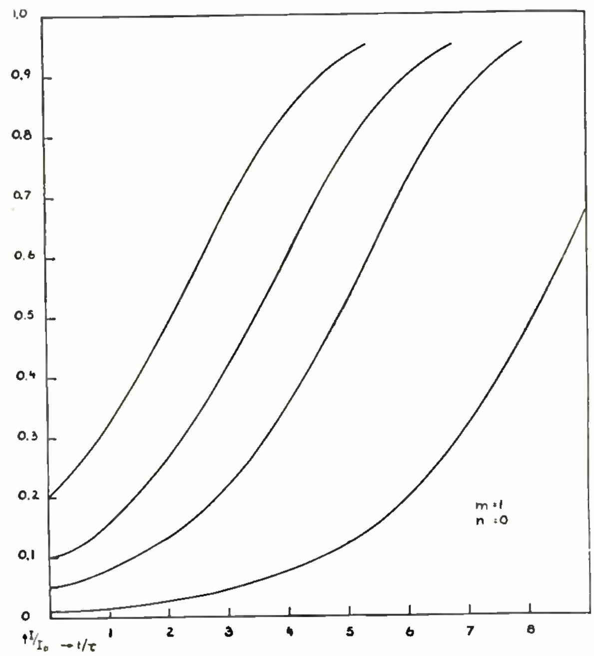

Fig. 4—Decay of the current I in the coil into its stable state I_0 for $m=1$, $n=0$ and for various starting points past the unstable state $I=0$. Same time scale as in Fig. 2.

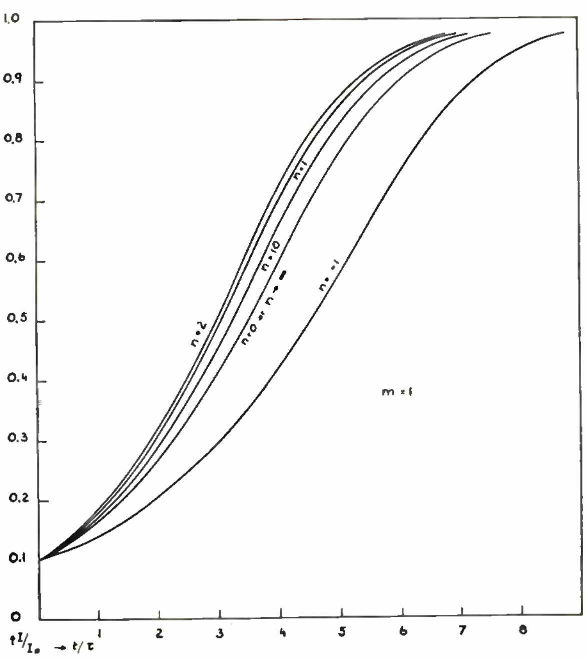

Fig. 3—Decay of the current I in the coil into the stable state I_0, for the case $m=1$ and various values of n. The time decreases when n is increased up to 2 and then increases again. For $n=\infty$ the curve coincides with that for $n=0$. Same time scale and starting points as in Fig. 2.

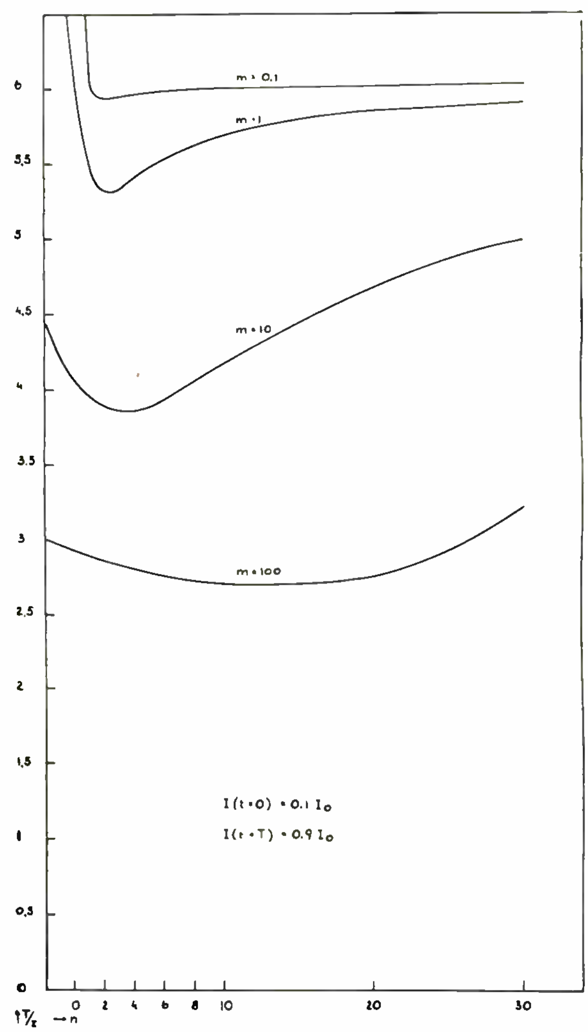

Fig. 5—The time it takes the bridge magnetoresistive flip-flop to decay from 0.1 to 0.9 of the stable state current, as function of the 2 parameters m and n defined in the text. The time is given in terms of the linear time constant, τ.

such large values of m, τ increases with m, usually much more rapidly than the decrease of t/τ (see Part V) so that t itself increases.

A study of Figs. 2 through 5 clearly shows that t/τ does not change appreciably with quite large variation of the parameters involved. This means that the time of transfer (without external pulse) from any value near the unstable states to any value near the stable point may be taken as practically independent of the design of the network. A value of, for example, 3–5 τ seems plausible from the graphs. This facilitates the handling since the formula for τ is much simpler than that for t.

The applicability of τ to expressing the time of transfer is more easily seen when one tries to compare it with some other time constant, e.g., L/R, which always appears in *linear* circuits. In Table I some numerical values [calculated from (17)] are given for the case $n=1$.

TABLE I

m	$\tau/(L/R)$	T/τ	$T/(L/R)$
0.1	5.0354	6.0151	30.288
1	0.5854	5.4428	3.185
10	0.1070	3.9479	0.422
100	0.0276	2.8898	0.080

It is seen from the values of $\tau R/L$ in the second column of the table that had the time been given in terms of L/R instead of in terms of τ, one would have had to magnify the scale of the first graph in Fig. 2 five times and decrease that of the last one 30 times. It is also seen from the last two columns of the table that while T (the time needed to transfer from 0.1 I_0 to 0.9 I_0), is increased only 2.1 times in terms of τ, for the given range of m, this time increases 380 times in terms of L/R for the same range of m.

It should be noted that the time calculated here is not the flipping time from one stable state to the other, an accurate definition of which should involve a definition of the incoming pulse[3] as well as the way by which the output of this element is applied to similar circuits. However, without considering these questions, which will be partly treated in Part X, the above time of transfer seems to be a good measure for the flipping time. This is due to the fact that T/τ is independent of the circuit parameters, so that it will not change when the element is loaded in the input and output circuits. It is also practically independent of the value past the unstable state where the input pulse stops, so that one can expect it to be independent of the shape of the pulse and of the form of its application to the coil.

One can therefore conclude that the flipping time is at least about 3τ, when τ is given by (13).

[3] M. Rubinoff, "Further data on the design of Eccles-Jordan flip-flops," *Electrical Engrg.*, vol. 71, pp. 905–910; June, 1952.

IV. τ as a Function of the Power Dissipated by the Magnetoresistors

In the stable state there is a constant current through each of the four magnetoresistors of Fig. 1. This dc is limited by practical considerations of heat dissipation. From the symmetry of the circuit, it is clear that the power dissipated by R_1 in one stable state will be dissipated by R_2 in the other stable state, and vice versa. It will therefore suffice to calculate the dc power for one stable state only, e.g., for $+I_0$.

From the analysis of the circuit of Fig. 1 in the stationary states (i.e., $\partial I/\partial t=0$) when r_2 is negligible, one obtains the power dissipated by R_1:

$$R_1 i_1^2 = (V - rI_0^2)/4R_1 \tag{18}$$

and similarly

$$R_2 i_2^2 = (V + rI_0)^2/4R_2. \tag{19}$$

For the state in which $I=+I_0$,

$$R_1 = R(1 - MI_0 + NI_0^2),$$
$$R_2 = R(1 + MI_0 + NI_0^2). \tag{20}$$

This means that each magnetoresistor should be designed so that it can dissipate at least the dc power

$$W = \mathrm{Max}\left[\frac{(V + r_1I_0)^2}{4R(1 + MI_0 + NI_0^2)}, \frac{(V - r_1I_0)^2}{4R(1 - MI_0 + NI_0^2)}\right]. \tag{21}$$

Substituting for V from (6), for M from (7), for I_0 from (16) and for r from (5), one gets

$$W = (R/4\lambda N) \max (P, Q) \tag{22}$$

with

$$P = [\sigma + m(\sigma - 1)(\lambda a)^{1/2}]^2/[1 + a + (\lambda a)^{1/2}], \tag{23}$$
$$Q = [\sigma + m(\sigma - 1)(\lambda a)^{1/2}]^2/[1 + a - (\lambda a)^{1/2}], \tag{24}$$

where a is given by (15).

It is evident from Fig. 1 that

$$V > r_1 I_0,$$

i.e.,

$$\sigma + m - (\sigma - 1)(\lambda a)^{1/2} > 0. \tag{25}$$

Also, $\lambda < 4$ by (10) so that

$$1 + a - (\lambda a)^{1/2} > 1 + a - 2a^{1/2} = (1 - a^{1/2})^2 \geq 0. \tag{26}$$

Consider now the following arrangement. The four magnetoresistors, each having the volume v cm³, are placed in small gaps in one or more toroids, made of ferromagnetic material with permeability μ and volume $\bar{V}$ cm³. If the magnetic field H is assumed to be uniform in the gaps, the energy stored in it is

$$E = H^2(4v + \bar{V}/\mu)/8\pi \text{ erg.}$$

On the other hand, the field is produced by the current of I ampere in the coil L, so that

$$E = \tfrac{1}{2}LI^2 \times 10^7 \text{ erg.}$$

Comparing the foregoing, one obtains

$$W = \frac{R(v + \overline{V}/4\mu)}{\lambda(4-\lambda)C\pi L10^7} \operatorname{Max}(P, Q)$$

and, finally, by using this relation to eliminate L/R from (13),

$$\tau = \frac{v + \overline{V}/4\mu}{2\pi 10^7 CW} \tau^* \tag{27}$$

with

$$\tau^* = (1 + a) \operatorname{Max}(P, Q)/[a\lambda(4-\lambda)(n^2 + 4m)^{1/2}]. \tag{28}$$

Here a, P and Q are functions of σ, λ and m, defined by (15), (23) and (24) respectively.

V. Estimation of τ

The expression for τ^* is rather complicated, involving three independent parameters, σ, λ, m, whose range is restricted only by (10) and (11). For the limiting case $\sigma = 1$, it was proved analytically[4] that τ^* has a *single* minimum, $\tau^*_{\min} = 1.677$ at $m = 1.069$, $\lambda = 1.67$. The general case was only numerically treated using the WEIZAC (the electronic computer of the Weizmann Institute of Science, Rehovot, Israel).

Typical results of these computations (the case $m = 1$) are plotted in Fig. 6. It is seen that an ellipsoid-like surface is obtained with a single minimum with respect to λ. It should be noted that the derivative with respect to λ is not continuous and some of the graphs "break" at a certain point. This is especially noticeable at $\lambda = 3$ on the curve $\sigma = 2$. These discontinuities are caused by the switchover from P to Q in (28).

Additional computations seemed to confirm that the functional dependence on λ, with a single minimum as illustrated by Fig. 6, was typical. Some of the results for these minima are plotted in Fig. 7 as a function of σ, with m as a parameter. Because of the overlap of the curves, the one for $m = 16$ is drawn dashed whereas $m = 32$ is omitted.

In Fig. 8, the same results are plotted as a function of m with σ as a parameter. It shows that τ^* has a single minimum as a function of m also. The minima are plotted in Fig. 9 as a function of σ.

The results thus obtained show that τ^* is never lower than the value 1.677 obtained for $\sigma = 1$. Substituting this value in (27) one obtains

$$\tau \geq \frac{v + \overline{V}/4\mu}{CW} \times 2.67 \times 10^{-8}. \tag{29}$$

[4] A. Aharoni, "Magneto-Resistive Memory," a Ph.D. thesis submitted to Hebrew University, Jerusalem, in 1957 (in Hebrew).

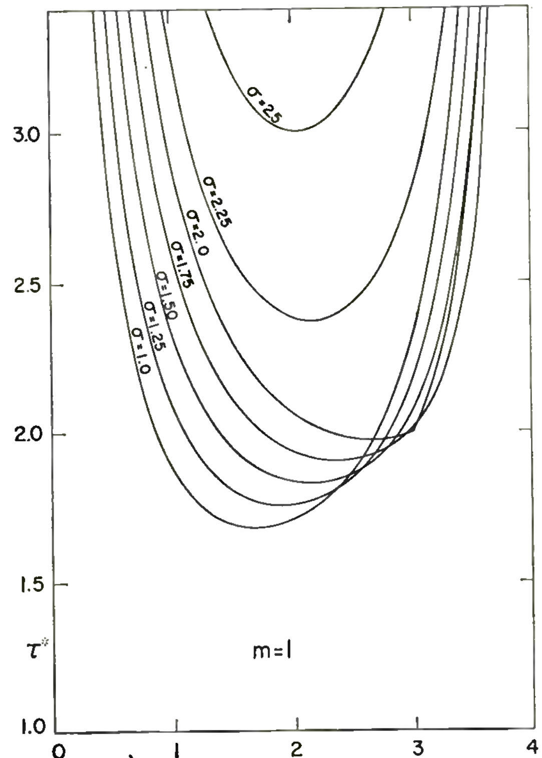

Fig. 6—The reduced time constant τ^* defined by (28) for $m = 1$, as a function of σ and λ.

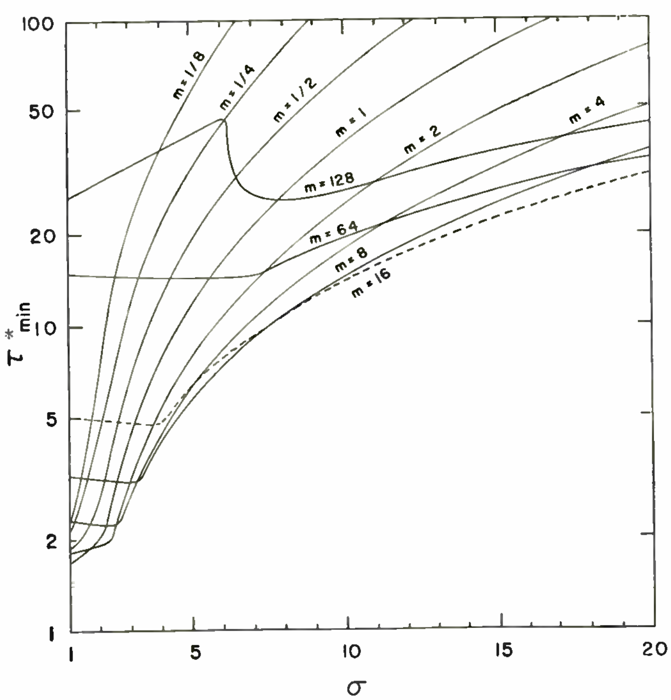

Fig. 7—The minima with respect to λ of the reduced time constant τ^* defined by (28), as a function of σ, for various values of m.

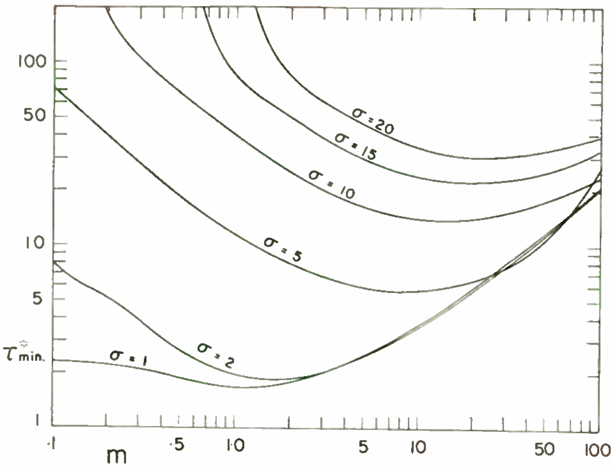

Fig. 8—The minima of Fig. 6 as a function of m, for various values of σ.

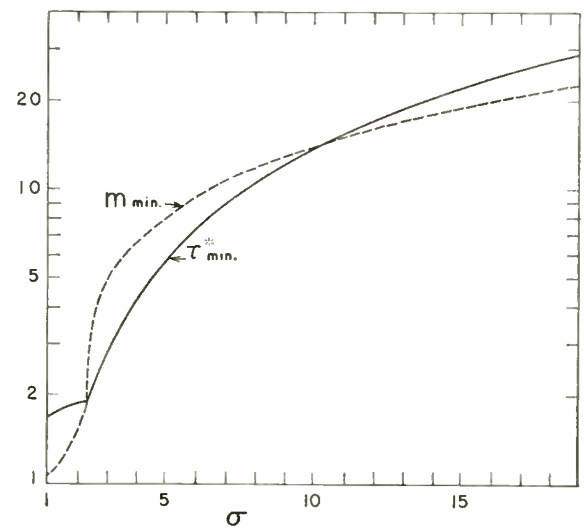

Fig. 9—The minima of Fig. 8 with respect to m and the values of m of which these minima are realized, plotted as functions of σ.

According to the values[1] of C for the known magnetoresistive materials at temperatures which are not lower than that of liquid air and for conceivable values of the power density W/v which the magnetoresistors can dissipate, the flipping time cannot be smaller in order of magnitude than 10^{-3} seconds, which is too large for practical use as a flip-flop.

VI. General Network Considerations

The method of calculation of the flipping time will be extended to more general networks. Consider a network composed of ohmic resistors, of magnetoresistors or other nonlinear resistors of known characteristics, and of coils. (Capacitors will be neglected here.) The voltage source in the kth loop is denoted by V_k, and the current in this loop by i_k. Suppose that the network is such that its behavior can be described by the equation

$$V_k = A_k(i_1, \cdots, i_n) + \sum_{m=1}^{n} B_{km} di_m/dt \tag{30}$$

with known constants V_k, B_{km} and known A_k as functions of the currents (but not as functions of the time). These A_k and B_{km} can be found for a general *magnetoresistive* network. It is obvious that for any *specified* network these expressions can be found from Kirchhoff's laws, if the characteristics of the nonlinear resistors are known.

Suppose further that the network has at least two stable equilibriums in addition to a number of unstable ones. If the currents in an equilibrium state are denoted by I_m, (30) implies that these values should satisfy the equation

$$V_k = A_k(I_1, \cdots, I_n). \tag{31}$$

Consider now the system in the vicinity of a stable state, *i.e.*, for

$$i_k = I_k + \delta I_k(t)$$

where δI_k is small enough, so that its orders higher than the first can be neglected. For a first approximation, one can express (30) in the form

$$V_k = A_1(I_1, \cdots, I_n) + \sum_m \delta I_m \partial A_k/\partial I_m + \sum_m B_{km} d\delta I_m/dt. \tag{32}$$

Substituting from (31) one can write (32) using matrix notations $(F) = (F_{km})$, $(B) = (B_{km})$, $\boldsymbol{\delta I} = (\delta I_m)$, in the form

$$F\boldsymbol{\delta I} + Bd\boldsymbol{\delta I}/dt = \mathbf{0} \tag{33}$$

where

$$F_{km} = \partial V_k/\partial I_m \tag{34}$$

and the derivatives are calculated from (31).

The solution of (33) is clearly a sum of exponentials, for if one substitutes in (33)

$$\boldsymbol{\delta I} = \boldsymbol{\delta I}_0 e^{-t/\tau} \tag{35}$$

one has

$$(\tau F - B)\boldsymbol{\delta I}_0 = \mathbf{0}. \tag{36}$$

A necessary and sufficient condition for the existence of solutions to this set of equations is the vanishing of the determinant of coefficients. That means that (35) is the solution of (33) if, and only if, the time constant τ is one of the roots of

$$|\tau F - B| = 0. \tag{37}$$

It should be noted that if F is a regular matrix, one can write (36) in the form

$$F^{-1}B\boldsymbol{\delta I}_0 = \tau\boldsymbol{\delta I}_0$$

which means that $\boldsymbol{\delta I}_0$ is a characteristic vector, and τ is a characteristic root of the matrix $F^{-1}B$. Eq. (37) is a more general definition for τ including the case $|F| = 0$ as well.

The time constants defined by (37) then determine the decay of the system into a stable equilibrium at the end of the process of flipping from one stable state to the other. It is thus reasonable to assume that the flipping time to this stable state is at least of the order of the largest root τ of (37). Eq. (37) then provides a practical and convenient first approximation to the flipping time in a wide class of general nonlinear networks, or at least a lower limit, which in most cases is a sufficient approach. It should be noted that the treatment of nonlinear resistors with the rigorism of Part VIII is not common in the literature beyond a one-loop network.[5] Even in the Eccles-Jordan bistable circuits which have already been thoroughly investigated, one does not introduce the nonlinear element (the vacuum tube) into the calculation, but rather one calculates the time-constant of the condenser's decay.[3] It should also be noted that the roots of (37) are generally not required and an estimation of the largest root usually suffices.

Another use of (37) is in the determination of the stability of equilibriums. Suppose the roots of (31) are known for a certain problem, but it is not yet known which of them are stable and which are not. In this case, one can construct (37) for each state in turn and determine the *signs* of the roots. It is clear from (35) that only positive roots give a decay into the considered state, while, if at least one of the roots is negative, the system will tend to withdraw from that state.

Though B, a matrix of coils and their couplings, is always symmetric, as will be proved below, the matrix F is not symmetric for most magnetoresistive cases. It might therefore occur that the roots of (37) are not real. This is no drawback since by (35) only the real part of τ determines both the flipping time and the stability, while the case of pure imaginary roots gives sustained oscillations and does not occur when multistable systems are considered. One can therefore state that a necessary and sufficient condition for an equilibrium to be stable is that the real parts of all the nonzero roots of (37) be positive.

The problem now is to define more specifically the matrices B and F involved in (37) for the case where the nonlinear resistors are magnetoresistive ones. For simplicity of notation, it will be assumed here that not more than one magnetoresistor is placed in each branch of the network. Many other cases not obeying this restriction may also be included in the following argument, but it would unnecessarily complicate the notations if the restriction were to be removed.

According to the above assumption, every magnetoresistor is uniquely determined by the current in two adjacent loops. The resistor defined thus by the kth and the mth current will be denoted by R_{km} (see Fig. 10). The ohmic resistance in the same branch will be denoted by r_{km}. This notation implies no restrictions since ohmic resistors in series may be taken as one for any

[5] L. A. Pipes, "Analysis of electric circuits containing nonlinear resistance," *J. Franklin Inst.*, vol. 263, pp. 47–55; January, 1957.

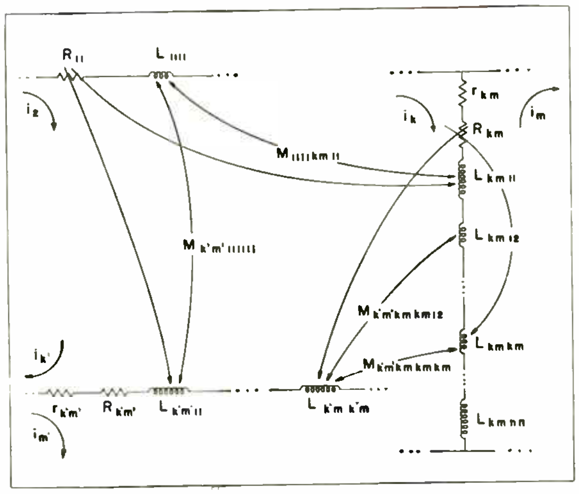

Fig. 10—Schematic representation of the notation used for the general magnetoresistive networks. An arrow across a resistor directed towards a coil indicates that the former is a magnetoresistor whose resistance depends on the current in the latter. A double arrow designates a coupling between coils.

network calculations. From these definition, it is evident that

$$R_{mk} = R_{km}, \qquad r_{mk} = r_{km}. \tag{38}$$

In each branch of the network only $\frac{1}{2}n(n+1)$ coils will be allowed in this treatment (n being the number of loops) each being wound on a different magnetoresistor. The inductance of the coil placed between the current i_k and i_m and wound on the R_{pq} will be denoted by L_{kmpq} (see Fig. 10, in which the arrow directed from a resistor to a coil designates that the former is a magnetoresistor and is influenced by the current in the latter). The coupling between L_{kmpq} and L_{rstj} (as referred to the current i_p and i_r) will be denoted by $M_{kmpqrstj}$. It is clear from the definition that

$$M_{kmpqrstj} = M_{rstjkmpq} = M_{kmpqrsjt} = -M_{mkpqrstj}. \tag{39}$$

It is assumed that only the coils thus defined are those whose magnetic fields change the resistance of the magnetoresistors. A constant magnetic field may be included by adding a loop made of a coil in series with a current source. However, this assumption excludes coils which are not wound on magnetoresistors, *e.g.*, coupling transformers or chokes, unless at least one of the R_{km} in the network is zero and the related coils are not. The definition of the L's also implies that only coils wound on the same resistor may be mutually coupled, *i.e.*,

$$M_{kmpqrstj} = 0 \text{ if } (p, q) \neq (t, j).$$

Using these definitions, one can write Kirchhoff's law for the kth loop in the form

$$V_k - \sum_m A_{km} i_m = \sum_{mpq} L_{kmpq} d(i_k - i_m)/dt + \sum_{pq} L_{kkpq} di_k/dt + \sum_{mpqrs} M_{kmpqrspq} di_r/dt \tag{40}$$

where

$$A_{km} = \begin{cases} \sum_p (R_{kp} + r_{kp}), & m = k \\ -R_{km} - r_{km}, & m \neq k. \end{cases} \tag{41}$$

It should be noted that the M's were defined for *different* coils only, so that the last summation in (40) contains only terms for which $(k, m) \neq (r, s)$. Now, it is readily seen that, if one defines

$$M_{kmpqkmpq} = L_{kmpq}, \tag{42}$$

(39) still applies for the new terms and the definitions of M can now be collected into the following relation (assuming ideal coupling):

$$M_{kmpqrstj} = \pm (L_{kmpq}L_{rstj})^{1/2}(\delta_{pt}\delta_{qj} + \delta_{pj}\delta_{qt} - \delta_{pq}\delta_{pt}\delta_{pj}). \tag{43}$$

Here δ_{pt} is the usual *Kronecker delta* which has the value 1 if $p=t$ and zero otherwise. According to these notations,

$$\sum_{mpq} L_{kmpq} d(i_k - i_m)/dt = \sum_{\substack{mpq \\ m\neq k}} M_{kmpqkmpq} di_k/dt + \sum_{mpq} M_{kmpqmkpq} di_m/dt.$$

It is seen that the first two summations of (40) contain the terms which are needed to complete the summation of the last sum, so that one obtains

$$V_k - \sum_m A_{km} i_m = \sum_{mpqrs} M_{kmpqrspq} di_r/dt,$$

i.e.,

$$V_k = \sum_m A_{km} i_m + \sum_m B_{km} di_m/dt \tag{44}$$

where

$$B_{km} = \sum_{rpqs} M_{krpqmspq}. \tag{45}$$

By comparing (44) to (30) it is seen that (45) is the required definition for B to be substituted in (37). It is a symmetric matrix since by (45) and (39)

$$B_{mk} = B_{km}. \tag{46}$$

Using (45) one can evaluate B from the network data concerning the coils and their couplings. To use (37) one still has to have a convenient definition for the matrix F. In order to do this, consider the geometrical arrangement in which L_{pqkm} consists of N_{pqkm} turns wound on an almost closed toroid of length s_{km} centimeters (Fig. 11). The toroid is made of a material having high permeability, μ_{km}, and its cross section is S_{km} cm². The magnetoresistor R_{km} is placed in a small air gap s_{km}' centimeters wide. If one applies a plus or minus sign to N according to the relative direction of winding, one has for this geometry, according to (43),

$$M_{pqmabkm} = \frac{4\pi S_{km} N_{pqkm} N_{abkm}}{s_{km}' + s_{km}/\mu_{km}} 10^{-9} \text{ Henry}. \tag{47}$$

On the other hand, the field produced by such a coil L_{pqkm} at R_{km}, due to the current i_p amperes, is

$$H_{km} = 4\pi N_{pqkm} i_p / 10(s_{km}' + s_{km}/\mu_{km}) \text{ gauss}.$$

It follows that the whole field at R_{km} is

$$H_{km} = \frac{4\pi}{10(s_{km}' + s_{km}/\mu_{km})} \sum_{pq} N_{pqkm} i_p, \tag{48a}$$

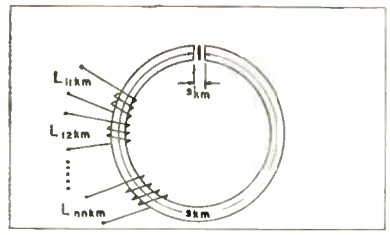

Fig. 11—Schematic representation of the geometric arrangement of coils and magnetoresistors.

it being assumed as before that the N's have the proper signs according to the direction of the windings.

Assuming that the magnetoresistive effect is essentially quadratic, as it is in most practical cases, one can write

$$R_{km} = R_{km}^{(0)}(1 + CH_{km}^2). \tag{48b}$$

The linear term may be added to the appropriate r_{km}, so that

$$R_{km} = R_{km}^{(0)} C H_{km}^2.$$

Substituting into this from (48)

$$R_{km} = R_{km}^{(0)} C \frac{(4\pi)^2}{10^2(s_{km}' + s_{km}/\mu_{km})^2} \sum_{pqab} N_{pqkm} \cdot N_{abkm} i_p i_a$$

or, according to (47),

$$R_{km} = \alpha_{km} \sum_{pqab} M_{pqkmabkm} i_p i_a \tag{49}$$

where

$$\alpha_{km} = \frac{4\pi 10^7 C R_{km}^{(0)}}{S_{km}(s_{km}' + s_{km}/\mu_{km})}. \tag{50}$$

Substituting (49) in (41) and using (44) one obtains

$$V_k = \sum_m E_{km} i_m + \sum_{mpq} D_{kmpq} i_m i_p i_q + \sum_m B_{km} di_m/dt \tag{51}$$

with

$$E_{km} = E_{mk} = \begin{cases} \sum_s r_{ks}, & \text{if } m = k \\ -r_{km} & \text{if } m \neq k \end{cases} \tag{52}$$

and

$$D_{kmpq} = \begin{cases} \sum_{abs} \alpha_{ks} M_{paksqbks}, & \text{if } m = k \\ -\alpha_{km} \sum_{ab} M_{pakmqbkm}, & m \neq k. \end{cases} \tag{53}$$

Definition (53) and the relations (39) imply

$$D_{kmpq} = D_{kmpq} = D_{mkpq}. \tag{54}$$

The equilibriums of the general magnetoresistive network discussed are then the roots of the set of cubic equations:

$$V_k = \sum_m E_{km} I_m + \sum_{mpq} D_{kmpq} I_m I_p I_q. \tag{55}$$

Hence, according to (34),

$$F_{km} = E_{km} + \sum_{pq} D_{kmpq} I_p I_q + \sum_{sq} D_{ksmq} I_s I_q + \sum_{sp} D_{ksrm} I_s I_p$$

or by (54)

$$F_{km} = E_{km} + \sum_{pq} (D_{kmpq} + 2D_{kpmq}) I_p I_q. \tag{56}$$

Eq. (56) is then the required relation for the matrix F to be substituted in (37) for the calculation of the time constants. D and E are readily evaluated using the network data and (50), (52) and (53).

VII. Properties of the Parameters

For the numerical analysis of a specified network, it is sufficient to solve (37). However, if one wants to investigate the time for a more general group of such networks, one has to have a measure for the time constants expressed by the parameters involved in this treatment.

Consider a certain magnetoresistor R_{km}. In a stationary state the current which flows through it is $I_k - I_m$, while its resistance is at least $R_{km}^{(0)}$, according to (48b). This resistor should therefore be designed so that it can dissipate at least the stable state power

$$W_{km} = R_{km}^{(0)}(I_k - I_m)^2.$$

Substituting this in (50) one obtains

$$\frac{1}{\alpha_{km}(I_k - I_m)^2} = \frac{S_{km}(s'_{km} + s_{km}/\mu_{km})}{4\pi 10^7 C W_{km}} = \frac{v_{km} + \overline{V}_{km}/\mu_{km}}{4\pi 10^7 C W_{km}} \tag{57}$$

But, according to Part V, the expression on the right hand side is at least of the order of 10^{-3} seconds. One can therefore conclude that, for the network to be a practical flip-flop, *the largest root of (37) should be at least 2 orders of magnitude smaller than the smallest of the expressions* $\alpha_{km}^{-1}(I_k - I_m)^{-2}$ *of the network.*[6]

Using the measure, certain groups of magnetoresistive bistable networks were analyzed,[4] but all those tried could be proved to have time constants of at least the order of one of the $\alpha_{km}^{-1}(I_k - I_m)^{-2}$.

VIII. Flipping by an External Pulse

In the following the dependence of the flipping time on the magnitude of the incoming pulse, which transfers the elements from a stable state to the vicinity of the unstable state, will be discussed.

The circuit of Fig. 12 is assumed. The flipping pulse $E(t)$ with an internal resistance kR is applied via a coupling transformer. Here $k > 0$, and R is the resistance of each of the magnetoresistors R_1 and R_2 in the bias-

[6] For $k = m$ one should take $\alpha_{kk}^{-1} I_k^{-2}$ since I_k is the current through R_{kk}.

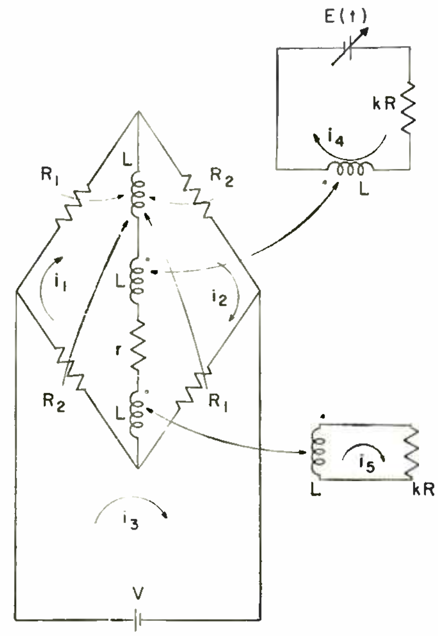

Fig. 12—Schematic representation of the bridge magnetoresistive flip-flop is discussed. The resistance of R_1 and R_2 depends on the current through L. A pulse $E(t)$ serves to flip over from one stable state of the current in the coil to the other. A linear load kR is assumed to be transformer coupled.

ing field H_0. Since in most applications, a flip-flop circuit has to operate a similar cascaded element, a load kR is added in Fig. 12. The load is made equal to the pulsing internal resistance since this is usually the most favorable condition to utilize the power involved. The internal resistance of the dc source V is assumed to be zero. For simplicity, all the coils are assumed to be equal, each having an inductance L. The resistance of the magnetoresistors is assumed as in (20). Analysis of the circuit gives

$$kRi_4 + L di_4/dt = L dI/dt + E(t) \tag{58}$$

$$kRi_5 + L di_5/dt = L dI/dt \tag{59}$$

$$\frac{-mI + nXI^3 + X^2 I^5}{1 + XI^2} = \frac{L}{R}\frac{d}{dt}(i_4 + i_5 - 3I). \tag{60}$$

From (58), (59), (60) and (30) it is seen that the B matrix for this case is

$$B = \begin{vmatrix} 3 & -1 & -1 \\ -1 & 1 & 0 \\ -1 & 0 & 1 \end{vmatrix} \frac{L}{R}.$$

Hence, using (34) and (47) the linear time constant, τ, is the largest root of

$$\begin{vmatrix} \tau F_{11} - 3L/R & L/R & L/R \\ L/R & \tau k - L/R & 0 \\ L/R & 0 & \tau k - L/R \end{vmatrix} = 0 \tag{61}$$

with

$$F_{11} = 2a(n + 2a)/(1 + a) \tag{62}$$

where a is as defined in (15).

Subtracting the second column of (61) from the last one, and adding the second row to the last one, it is readily seen that one of its roots is

$$\tau_1 = L/kR \tag{63}$$

while the other two are the roots of

$$2ak(n+2a)\alpha^2 - \alpha\{3k(1+a) + 2a(n+2a)\} + (1+a) = 0 \tag{64}$$

with

$$\alpha = \tau R/L. \tag{65}$$

The time constant for this element without the input and output circuits has already been calculated above. The result (for a coil of $3L$ used here) was

$$\tau_0 = \alpha_0 R/L$$

with

$$\alpha_0 = \frac{3(1+a)}{2a(n+2a)}. \tag{66}$$

Substituting for a from (66), the largest root of (64) is

$$\alpha = (1/2k)[\alpha_0 k + 1 + (\alpha_0^2 k^2 + 1 + 2k\alpha_0/3)^{1/2}]. \tag{67}$$

It should be noted that the time constant given by (65) and (67) is larger than τ_0 and tends to this value only for $k \to \infty$. It is also evident that the square root in (67) is larger than 1, so that $\alpha > 1/k$. This means that the solution of this equation is larger than the one determined by (63). Since only the largest solution is of interest, the linear time constant, τ, is the one determined by (65) and (67).

IX. The Differential Equation for the Time

Using the above mentioned τ, let the following notations

$$T = t/\tau, \quad y = I/I_0, \quad z_1 = i_4/I_0, \quad z_2 = i_5/I_0 \tag{68}$$

be substituted in (58), (59) and (60). Using (66), these equations reduce to

$$d(z_1 - y)/dT = -\alpha k z_1 + \alpha E(T)/RI_0, \tag{69}$$

$$d(z_2 - y)/dT = -\alpha k z_2, \tag{70}$$

$$\frac{d}{dT}(3y - z_1 - z_2) = \alpha a y \frac{(1-y^2)(ay^2+b)}{1+ay^2}. \tag{71}$$

Here a and b are given by (15).

Now it can be verified by substitution that the solutions of (69) and (70), which satisfy the initial conditions $z_1(0) = z_2(0) = 0$, are

$$z_1 = e^{-k\alpha T}\int_0^T (\alpha E/RI_0 + dy/dt)e^{k\alpha T}dT, \tag{72}$$

$$z_2 = e^{-k\alpha T}\int_0^T (dy/dT)e^{k\alpha T}dT. \tag{73}$$

Substituting these solutions in (71) one obtains

$$\frac{dy}{dT} + k\alpha e^{-k\alpha T}\int_0^T\left(\frac{\alpha E}{RI_0} + 2\frac{dy}{dT}\right)e^{k\alpha T}dT - \frac{\alpha E}{RI_0} = \frac{\alpha a y(1-y^2)(ay^2+b)}{1+ay^2}. \tag{74}$$

One can differentiate (74) after multiplying both sides of this equation by $e^{k\alpha T}$, thus getting rid of the integral and obtaining a nonlinear differential equation of the second order. Here we shall consider only the simplest case

$$m = 1, \qquad n = 0. \tag{75}$$

In this case (67) reduces to

$$\alpha = (1/2k)[3k/2 + 1 + (9k^2/4 + 1 + k)^{1/2}] \tag{76}$$

and the differential equation becomes

$$\frac{d^2y}{dT^2} + \alpha\frac{dy}{dT}(3k - 1 + 3y^2) - k\alpha^2 y(1-y^2) = G\alpha^2 \tag{77}$$

with

$$G = (1/\alpha RI_0)(dE/dT). \tag{78}$$

Let the pulse be initiated at $T=0$ when the element is in one of its stable states ± 1, for example in the state $y = -1$. The pulse voltage is either stepped in zero time to a value $E(0)$ or rises continuously from zero, in which case $E(0) = 0$. In either case, the currents cannot change discontinuously, due to the presence of inductance, so that (58) and (59) imply

$$(di_4/dt)_{t=0} - (dI/dt)_{t=0} = E(0)/L, \qquad (di_5/dt)_{t=0} = 0.$$

Substituting these relations in (60) and taking into account that at $t=0$ the element is in a stable state, one obtains

$$(dI/dt)_{t=0} = E(0)/2L,$$

i.e.,

$$(dy/dT)_{T=0} = \alpha E(0)/RI_0. \tag{79a}$$

Eq. (79a) and

$$y(0) = -1 \tag{79b}$$

are the necessary initial conditions for the solution of (77).

X. Numerical Computations and Results

Eq. (77) is nonlinear but does not contain the independent variable explicitly and can therefore be reduced to a first order equation. By substituting

$$u = dT/dy \tag{80}$$

it is transformed into

$$du/dy = \alpha u^2(3k - 1 + 3y^2) - k\alpha^2 u^3 y(1-y^2) - G\alpha^2 u^3. \tag{81}$$

This is a special case of Abel's equation[7] whose solution cannot be expressed in terms of tabulated functions. Therefore a numerical computation was undertaken.

The first case considered was a ramp pulse (*i.e.*, E increases linearly up to a certain value and then remains constant). According to (78), this assumption implies a constant value for G to be substituted in (77), for $0 \le T \le T^*$ and $G=0$ for $T \ge T^*$. For the solution of (77) one thus has the 3 parameters k, G, T^*. According to (79) the initial conditions for this case are

$$y(0) = -1, \; y'(0) = 0.$$

Using these values and the Fourth order Runge-Kutta Method,[8] (77) was solved numerically for some values of the parameters involved. The results for $k=1$, $G=1$ are plotted in Fig. 13. Curve 1 of this figure is for an unterminated pulse. In curve 2 the pulse is terminated at the unstable state ($G=0$ from $y=0$ onwards). In the other curves the pulse is terminated at the points designated by the arrows. It is clear that the higher the energy supplied to the element, the shorter the flipping time. However, the most interesting case is that in which the energy supplied by the pulse does not exceed the energy which the element needs to flip over a cascaded element of similar type. Since it is difficult to estimate this energy, it was assumed that the power supplied by the pulse when it is terminated equals RI_0^2 which is of the order of the internal power. Curve (4) of Fig. 13 comes roughly within this limit; but it is seen that the energy supplied in this case does not suffice to flip-over the element, and it falls back to the initial state.

Similar results for $k=1$, $G=3$ are plotted in Fig. 14. In this figure, also, curve 1 represents an unterminated pulse and the arrows designate the points of termination for the other pulses (T^*). In this case again the terminations should be such that the power supplied is somewhat higher than RI_0^2, otherwise the element falls back to the initial state. Of special interest is curve 4, where the element almost reaches the unstable state $y=0$, and remains in its vicinity for rather a long time.

The results for the derivative of y in the case $k=G=1$ were used to calculate z_1 and z_2 from (72) and (73). The results are plotted in Figs. 15 and 16, respectively. The indexes 1 and 2 in these figures refer to the appropriate curves in Fig. 13. It is seen that in this case at least the output current is considerably smaller than the input current, which prevents cascading elements. No attempt was made to look for output amplitudes which would be at least equal to the input.

[7] E. Kamke, "Differentialgleichungen," Chelsea Publishing Co., New York, N. Y., Band 1, p. 24; 1948.

[8] J. B. Scarborough, "Numerical Mathematical Analysis," The Johns Hopkins Press, Baltimore, Md., 2nd Edition, pp. 299–303; 1950.

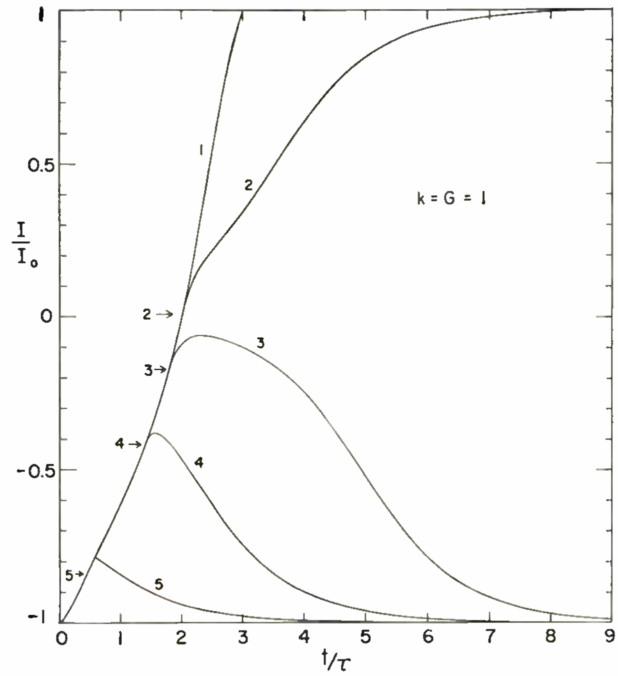

Fig. 13—The transfer of the current I in the coil from the stable state $-I_0$ to the stable state $+I_0$, for linearly rising pulse with $G=1$ and $k=1$. The arrows show the termination of the pulse and are indexed according to the indexing of the curves. Curve 1 is for an unterminated pulse. The time scale is referred to the linear time constant τ.

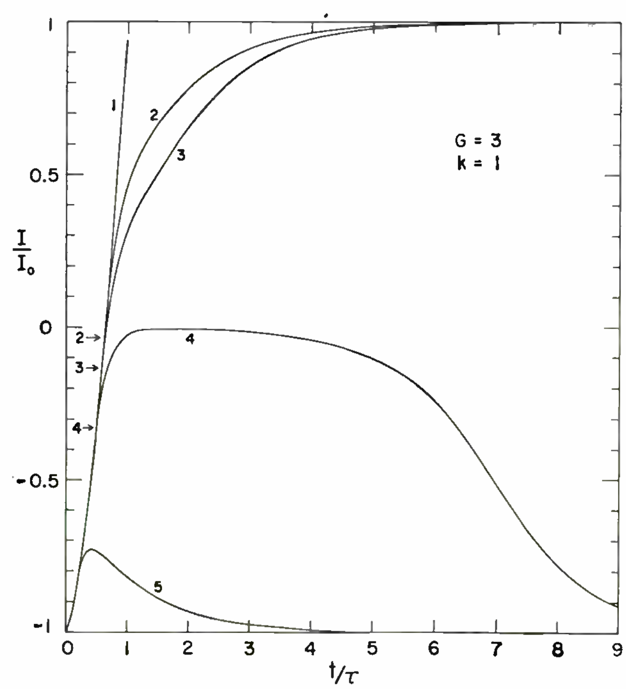

Fig. 14—Same as Fig. 13 with $G=3$.

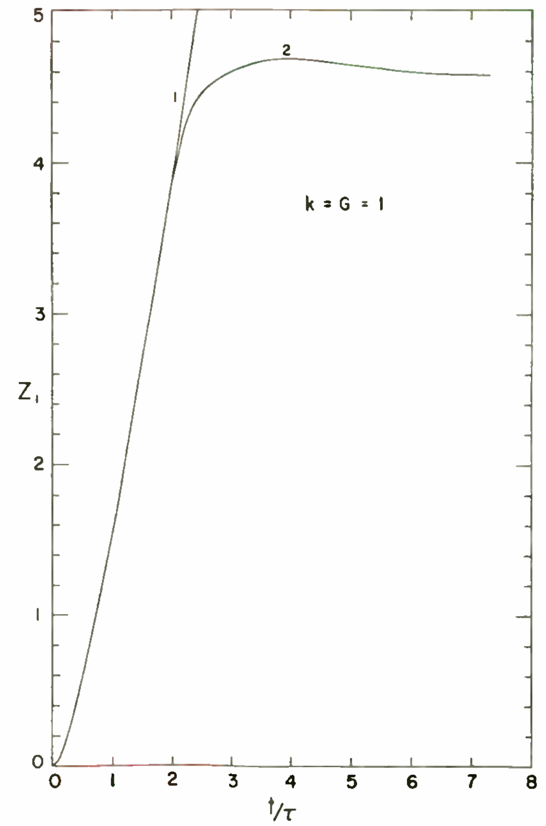

Fig. 15—The input pulse current of the curves 1 and 2 of Fig. 13 as a function of the time.

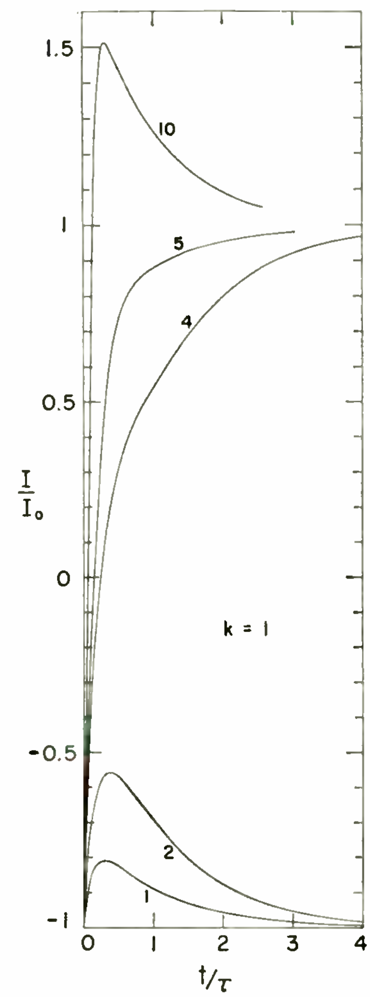

Fig. 17—The transfer of the current I in the coil from the stable state $-I_0$ to the stable state $+I_0$ for the step function pulse and for $k=1$. The values of E/RI_0 are indicated in the figure. Same time scale as in Fig. 13.

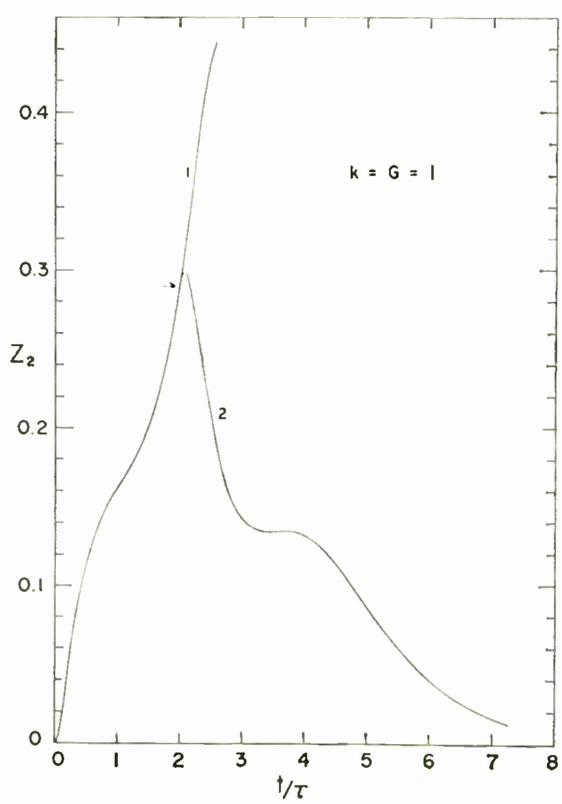

Fig. 16—The output current of the curves 1 and 2 of Fig. 13 as a function of the time.

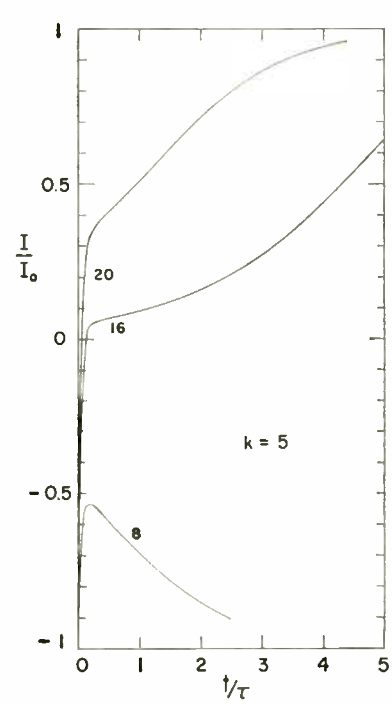

Fig. 18—Same as Fig. 17 for $k=5$.

The second type of flipping pulse considered was a step function for E, *i.e.*, a sudden rise from 0 to E of the voltage at $T=0$ and a constant voltage E for $T>0$. In this case one has to substitute $G=0$ in (77) and use E in the initial conditions (79) only.

Since the derivative of y is rather large, for large values of E, the Runge-Kutta method cannot be applied to (77) in this case. Therefore, the transformation (80) was used and the fourth order Runge-Kutta method was applied to (81). Computations from this equation were made until u was large enough. The last values were then used as initial values for the numerical solution of (77).

The results of the numerical computations for $k=1$ and $k=5$ are plotted in Figs. 17 and 18, respectively, for various values of E. The numbers indicated in the figures are the values of E/RI_0.

In the results plotted in Figs. 13, 14, 17, and 18 the time scale is referred to the linear time constant τ. It can, therefore, be seen that τ is a very good approximation for the flipping time for conceivable values of the power involved.

TABLE OF NOTATIONS

(MOST COMMONLY USED)

$C \cdots$ Magnetoresistive coefficient
(Bismuth from 3.1×10^{-6} Gauss^{-2} at liquid air temperature to 1.7×10^{-8} Gauss^{-2} at room temperature; slightly higher with InSb and some other semiconductors).

$K \cdots$ Constant of proportionality of magnetic field, produced by current I, Fig. 1.

$L \cdots$ Inductance of coil.

$R_1 \cdots R_4$, $r_1 r_2$ Resistances according to Figs. 1 and 12. $R_1=R_4$ and $R_2=R_3$.

$R_0 \cdots$ Resistance of $R_1 \cdots R_4$ in zero field.

$I_0 \cdots$ Current in a stable state.

$M=2CKH_0/(1+CH_0^2)$.

$N=CK^2/(1+CH_0^2)$.

$R=R_0(1+CH_0^2)$.

$\sigma=(r_1+R)/R$.

$\lambda=4CH_0^2/(1+CH_0^2)$.

$m=MV/R-\sigma$, characteristic amplification of element.

$n=\sigma+1-\lambda$.

$\tau \cdots$ Time Constant

$a=\frac{1}{2}[-n+(n^2+4m)^{1/2}]$.

$b=\frac{1}{2}[n+(n^2+4m)^{1/2}]$.

$P=[\sigma+m+(\sigma-1)(\lambda a)^{1/2}]^2/[1+a+(\lambda a)^{1/2}]$.

$Q=[\sigma+m-(\sigma-1)(\lambda a)^{1/2}]^2/[1+a-(\lambda a)^{1/2}]$.

$\tau^*=(1+a)\max(P, Q)/[a\lambda(4-\lambda)(n^2+4m)^{1/2}]$.

IRE Standards on Nuclear Techniques: Definitions for the Scintillation Counter Field, 1960*

60 IRE 13. S1

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Accelerating Electrode. An electrode to which a potential is applied to increase the speed of the electrons which constitute the space current.

Accelerating Grid. See *Accelerating Electrode.*

After Pulse. A spurious pulse induced in a *photomultiplier* by a previous pulse.

Anode (Electron Tubes). An electrode through which a principal stream of electrons leaves the interelectrode space.

Anti-Coincidence Circuit. A circuit that produces a specified output pulse when one (frequently pre-designated) of two inputs receives a pulse and the other receives no pulse within an assigned time interval.

Background Counts (in Radiation Counters). *Counts* caused by *ionizing radiation* coming from sources other than that to be measured.

Background Response (in Radiation Detectors). Response caused by *ionizing radiation* coming from sources other than that to be measured.

Coincidence Circuit. A circuit that produces a specified output pulse when and only when a specified number (two or more) or a specified combination of input terminals receives pulses within an assigned time interval.

Collector (Electron Tubes). An electrode that collects electrons or ions which have completed their functions within the tube.

Count (in Radiation Counters). A single response of the counting system.

Counting Efficiency (Scintillation Counters). The ratio of 1) the average number of photons or particles of *ionizing radiation* that produce *counts* to 2) the average number incident on the sensitive area. The operating conditions of the counter and the conditions of irradiation must be specified.

Counting-Rate Meter. A device that indicates the time rate of occurrence of input pulses averaged over a time interval.

Current Amplification (Photomultipliers). The ratio of 1) the signal output current to 2) the photoelectric signal current from the *photocathode.*

Dark Current (Phototubes). See *Electrode Dark Current.*

Dark Current Pulses (Phototubes). *Dark current* excursions that can be resolved by the system employing the *phototube.*

Delay Circuit. A circuit that produces an output signal that is delayed intentionally with respect to the input signal.

Delay Line. A real or artificial transmission line designed to introduce delay.

Delay Coincidence Circuit. A *coincidence circuit* that is actuated by two pulses, one of which is delayed by a specified time interval with respect to the other.

Discriminator, Amplitude. See *Discriminator, Pulse-Height.*

Discriminator, Pulse-Height. A circuit that produces a specified output pulse if and only if it receives an input pulse whose amplitude exceeds an assigned value.

Dynode. An electrode which performs a useful function by means of secondary emission.

Electrode Dark Current (Phototubes). The component of electrode current remaining when *ionizing radiation* and optical photons[1] are absent.

Focusing Electrode. An electrode to which a potential is applied to control the cross-sectional area of the electron beam.

Focusing Grid. See *Focusing Electrode.*

Full Width at Half Maximum (FWHM). The full width of a distribution measured at half the maximum ordinate. (See Fig. 1.) For a normal distribution, it is equal to $2\,(2\ln 2)^{1/2}$ times the standard deviation (σ).

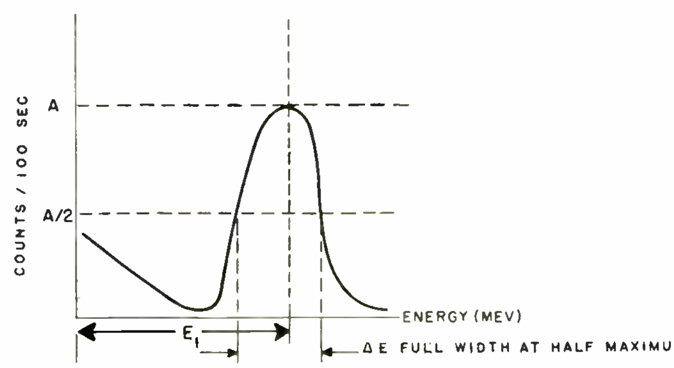

Fig. 1.

Note: The expression "full width at half maximum" given either as an absolute value or as a percentage of the value of the argument at the maximum of the distribution curve is frequently used in nuclear physics as an approximate description of a distribution curve. Its significance can best be made clear by reference to a typical distribution curve, shown in Fig. 1, of the measurement of the energy of the gamma rays from Cs^{137} with a scintillation counter spectrometer. The measurement is made by determining the number of gamma ray photons detected in a prescribed interval of time, having measured ener-

[1] Optical photons, for the purpose of this Standard, are photons with energies corresponding to wavelengths between 2000–15,000 Å.

gies falling within a fixed energy interval (channel width) about the values of energy (channel position) taken as argument of the distribution function. The abscissa of the curve shown is energy in million electron volt (mev) units and the ordinate is counts per given time interval per mev energy interval.

The maximum of the distribution curve shown has an energy E mev. The height of the peak, is A_1 counts/100 seconds/mev. The full width at half maximum ΔE is measured at a value of the ordinate equal to $A_1/2$. The percentage full width at half maximum is $100\ \Delta E/E_1$. It is an indication of the width of the distribution curve, and where (as in the example cited) the gamma ray photons are monoenergetic, it is a measure of the resolution of the detecting instrument. When the distribution curve is a Gaussian curve, the percentage full width at half maximum is related to the standard deviation σ by

$$100 \frac{\Delta E}{E_1} = 100 \times 2(2 \ln 2)^{1/2} \times \sigma.$$

Gain (Photomultipliers). See *Current Amplification.*

Ionizing Radiation. a) Particles or photons of sufficient energy to produce ionization in their passage through air. b) Particles that are capable of nuclear interactions with the release of sufficient energy to produce ionization in air.

Kick-Sorter (British). See *Pulse-Height Analyzer.*

Light Pipe. An optical transmission element that utilizes unfocused transmission and reflection to reduce photon losses.

Note: Light pipes have been used to distribute the light more uniformly over a *photocathode.*

Linear Pulse Amplifier. A *pulse amplifier* in which the peak amplitude of the output pulses is directly proportional to the peak amplitude of the corresponding input pulses, if the input pulses are alike in shape.

Multiplier Phototube. A phototube with one or more dynodes between its *photocathode* and output electrode.

Noise (Phototubes). The random output which limits the minimum observable signal from the *phototube.*

Photocathode. An electrode used for obtaining photoelectric emission.

Photocathode Blue Response. The photoemission current produced by a specified luminous flux from a tungsten filament lamp at 2870°K color temperature when the flux is filtered by a specified blue filter.

Photocathode Luminous Sensitivity. See *Sensitivity, Cathode Luminous.*

Photocathode Radiant Sensitivity. See *Sensitivity, Cathode Radiant.*

Photocathode Spectral Sensitivity Characteristic. See *Spectral Sensitivity Characteristic (Photocathode).*

Photocell. A solid state photosensitive electron device in which use is made of the variation of the current-voltage characteristic as a function of incident radiation.

Photomultiplier. See *Multiplier Phototube.*

Photon Emission Spectrum, Scintillator Material. The relative numbers of optical photons[1] emitted per unit wavelength as a function of wavelength interval. The emission spectrum may also be given in alternative units such as wave number, photon energies, frequency, etc.

Phototube. An electron tube that contains a *photocathode* and has an output depending on the total photoelectric emission from the irradiated area of the *photocathode.*

Pulse Amplifier. An amplifier designed specifically for the amplification of electrical pulses.

Pulse Counter. A device that indicates or records the total number of pulses that it has received during a time interval.

Pulse Decay Time. The interval between the instants at which the instantaneous amplitude last reaches specified upper and lower limits, namely, 90 per cent and 10 per cent of the peak pulse amplitude unless otherwise stated.

Pulse Duration. The time interval between the first and last instants at which the instantaneous amplitude reaches a stated fraction of the peak pulse amplitude

Pulse-Height Analyzer. An instrument capable of indicating the number or rate of occurrence of pulses falling within each of one or more specified amplitude ranges.

Pulse-Height Resolution, Electron (Photomultiplier). A measure of the smallest change in the number of electrons in a pulse from the *photocathode* that can be discerned as a change in height of the output pulse. Quantitatively, it is the fractional standard deviation (σ/A_1) of the pulse-height distribution curve for output pulses resulting from a specified number of electrons per pulse from the *photocathode.*

Note: The fractional *full width at half maximum* of the pulse-height distribution curve, (FWHM/A_1) is frequently used as a measure of this resolution, where A_1 is the pulse height corresponding to the maximum of the distribution curve.

Pulse-Height Resolution Constant, Electron (Photomultipliers). The product of the square of the *electron (photomultiplier) pulse-height resolution* expressed as the fractional *full width at half maximum* (FWHM/A_1), and the mean number of electrons per pulse from the *photocathode.*

Pulse Rise Time. The interval between the instants at which the instantaneous amplitude first reaches specified lower and upper limits, namely 10 per cent and 90 per cent of the peak pulse amplitude unless otherwise stated.

Pulse Shaper. Any transducer used for changing one or more characteristics of a pulse.

Note: This term includes pulse regenerators.

Pulse Stretcher. A *pulse shaper* that produces an output pulse whose duration is greater than that of the input pulse and whose amplitude is proportional to that of the peak amplitude of the input pulse.

Register, Mechanical. An electro-mechanical indicating *pulse counter.*

Resolving Time (Radiation Counters). The minimum achievable pulse spacing between *counts.*

Note: This quantity is a property of the combination of the tube and recording circuit.

Scaler, Pulse. A device that produces an output signal whenever a prescribed number of input pulses has been received. It frequently includes indicating devices for interpolation.

Scintillation. The optical photons[1] emitted as a result of the incidence of a particle or photon of *ionizing radiation* on a *scintillator.*

Scintillation Counter. The combination of *scintillation counter heads* and associated circuitry for detection and measurement of *ionizing radiation.*

Scintillation-Counter Cesium Resolution. The *scintillation-counter energy resolution* for the gamma ray or conversion electron from Cesium-137.

Scintillation-Counter Energy Resolution. A measure of the smallest difference in energy between two particles or photons of *ionizing radiation* that can be discerned by the *scintillation counter.* Quantitatively it is the fractional standard deviation (σ/E_1) of the energy distribution curve.

Note: The fractional *full width at half maximum* of the energy distribution curve ($FWHM/E_1$), is frequently used as a measure of the *scintillation-counter energy resolution* where E_1 is the mode of the distribution curve. (See Fig. 2.)

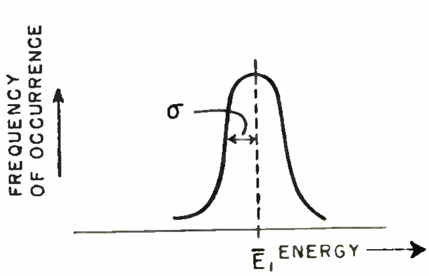

Fig. 2.

Scintillation-Counter Energy Resolution Constant. The product of the square of the *scintillation-counter energy resolution*, expressed as the fractional *full width at half maximum* ($FWHM/E_1$), and the specified energy.

Scintillation Counter Head. The combination of *scintillators* and *phototubes* or *photocells* which produces electrical pulses or other electrical signals in response to *ionizing radiation.*

Scintillation-Counter Time Discrimination. A measure of the smallest interval of time between two individually discernible events. Quantitatively it is the standard deviation (σ) of the time interval curve.

Note: The *full width at half maximum* of the time interval curve is frequently used as a measure of the time discrimination. (See Fig. 3.)

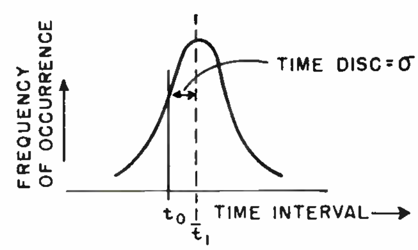

Fig. 3.

Scintillation Decay Time. The time required for the rate of emission of optical photons[1] of a *scintillation* to decrease from 90 per cent to 10 per cent of its maximum value.

Scintillation Duration. The time interval from the emission of the first optical photon[1] of a *scintillation* until 90 per cent of the optical photons[1] of the *scintillation* have been emitted.

Scintillation Rise Time. The time required for the rate of emission of optical photons[1] of a *scintillation* to increase from 10 per cent to 90 per cent of its maximum value.

Scintillator. The body of *scintillator material* together with its container.

Scintillator Conversion Efficiency. The ratio of the optical photon[1] energy emitted by a *scintillator* to the incident energy of a particle or photon of *ionizing radiation.* (The efficiency is generally a function of the type and energy of *ionizing radiation.*)

Scintillator Material. A material which emits optical photons[1] in response to *ionizing radiation.*

Note: There are five major classes of *scintillator materials,* namely,

1) inorganic crystals [*e.g.,* NaI(Tl) single crystals, ZnS(Ag) screens],

2) organic crystals (*e.g.*, anthracene, trans-stilbene),
3) solution scintillators: a) liquid, b) plastic, c) glass,
4) gaseous scintillators,
5) Cerenkov scintillators.

Scintillator Photon Distribution (in Number). The statistical distribution of the number of optical photons[1] produced in the scintillator by total absorption of mono-energetic particles. (See Fig. 4.)

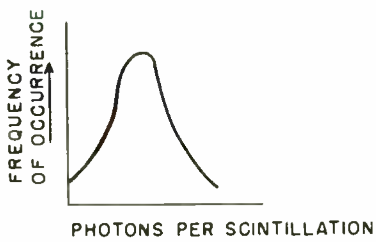

Fig. 4.

Scintillator Material Total Conversion Efficiency. The ratio of the optical photon[1] energy produced to the energy of a particle or photon of *ionizing radiation* which is totally absorbed in the *scintillator material.* (The efficiency is generally a function of the type and energy of the *ionizing radiation.*)

Selector, Amplitude. See *Selector, Pulse-Height.*

Selector, Pulse-Height. A circuit that produces a specified output pulse when and only when it receives an input pulse whose amplitude lies between two assigned values.

Semi-Transparent Photocathode. A *photocathode* in which radiant flux incident on one side produces photoelectric emission from the opposite side.

Sensitivity. The least signal input capable of causing an output signal having desired characteristics.

Sensitivity (of a Measuring Device). The ratio of the magnitude of its response to the magnitude of the quantity measured. It may be expressed directly in divisions per volt, millimeters per volt, milliradians per microampere, etc., or indirectly by stating a property from which sensitivity can be computed (*e.g.*, ohms per volt for a stated deflection).

Note: In the case of mirror galvanometers it is customary to express sensitivity on the basis of a scale distance of one meter.

Sensitivity, Cathode Luminous (Photocathodes). The quotient of photoelectric emission current from the *photocathode* by the incident luminous flux under specified conditions of illumination.

Note 1): Since *cathode luminous sensitivity* is not an absolute characteristic but depends on the spectral distribution of the incident flux, the term is commonly used to designate the sensitivity to radiation from a tungsten filament lamp operating at a color temperature of 2870°K.

Note 2): *Cathode luminous sensitivity* is usually measured with a collimated beam at normal incidence.

Sensitivity, Cathode Radiant (Photocathodes). The quotient of the photoelectric emission current from the *photocathode* by the incident radiant flux at a given wavelength under specified conditions of irradiation.

Note: Cathode radiant sensitivity is usually measured with a collimated beam at normal incidence.

Sensitivity, Incremental (of an Instrument). A measure of the smallest change in stimulus that produces a statistically significant change in response. Quantitatively it is usually expressed as the change in the stimulus which produces a change in response equal to the standard deviation of the response.

Sensitivity, Threshold (of an Instrument). A measure of the smallest stimulus that produces a significant response.

Spectral Quantum Yield (Photocathode). The average number of electrons photoelectrically emitted from the *photocathode* per incident photon of a given wavelength.

Note: The *spectral quantum yield* may be a function of the angle of incidence and of the direction of polarization of the incident radiation.

Spectral Sensitivity Characteristic (Photocathode). The relation between the *radiant sensitivity* and the wavelength of the incident radiation under specified conditions of irradiation.

Note: *Spectral sensitivity characteristic* is usually measured with a collimated beam at normal incidence.

Spurious Count. A count from a *scintillation counter* other than, a) one purposely generated or, b) one due directly to *ionizing radiation.*

Spurious Pulse. A pulse in a *scintillation counter* other than, a) one purposely generated or, b) one due directly to *ionizing radiation.*

Time Distribution Analyzer. An instrument capable of indicating the number or rate of occurrence of time intervals falling within one or more specified time interval ranges. The time interval is delineated by the separation between pulses of a pulse pair.

Time-Interval Selector. A circuit that produces a specified output pulse when and only when the time interval between two pulses lies between specified limits.

Time-Sorter. See *Time Distribution Analyzer.*

Wavelength Shifter (Scintillator). A photofluorescent compound used with a *scintillator material* to absorb photons and emit related photons of a longer wavelength.

Note: The purpose is to cause more efficient use of the photons by the *phototube* or *photocell.*

Noise in Oscillators*

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***Summary*—Noise affects the behavior of oscillators in at least two important ways. During sustained oscillation, noise creates undesired perturbations or modulation in both the amplitude and the phase of the wave. The amplitude perturbations produce a continuous spectrum which in typical situations is quite weak and broader than the bandwidth of the resonator. The phase perturbations disperse the nominal frequency into a continuous distribution which is of the same form but much stronger and narrower than for the amplitude perturbations.**

During the initiation of oscillation, noise constitutes the starting voltage and therefore affects the time required for the wave to reach some pre-established amplitude. The resulting jitter in the starting time of pulsed oscillators is objectionable because it degrades the signal-to-noise ratio in systems employing super-regenerative receivers or pulse-time modulation. The time and spectral distributions of noise effects in typical oscillators are derived and discussed in the following sections.

List of Symbols

A = amplitude of oscillation
C = capacitance
F = frequency
G = conductance
I = a current
K = constant of integration
L = inductance
P = power
Q = selectivity
R = the autocorrelation
T = temperature °K
U = mean noise amplitude
V = a voltage
ω = angular velocity
σ = probable error
ρ = an amplitude function
ϕ = a phase function
δ = logarithmic decrement
τ = a time interval
α = expansion coefficient
β = a variable related to decrement
γ = a dimensionless multiplier
π = the number 3.14159
μ = a dimensionless multiplier
a = a variable
b = a variable
d = the differential operator
e = the number 2.71828
f = frequency
h = a variable
i = a variable
j = a variable
k = Boltzmann's constant 1.38×10^{-23}
n = number of half-cycles
p = a variable, also probability
q = a variable, also cumulative probability
r = a variable
s = the saturation coefficient
t = time
u = an amplitude coeffcient
v = a voltage
x = an amplitude coefficient

Introduction

Of THE many publications concerned with noise, [2], [13], [15], a considerable number, [4], [7], [14], concern themselves with the more special problem of noise in oscillators. A variety of approaches has been taken, and results of undoubted value have been obtained. However, it appears that many important questions remain to be answered. In particular, it is believed that no previous treatment adequately describes either the distribution during build-up of oscillation or the spectrum during sustained oscillation. Conventional noise formulas are directly applicable only to steady-state situations in linear systems. Unfortunately, the build-up problem involves the transient state, and sustained oscillation involves nonlinearity. Thus, normal methods are inapplicable, and it has been necessary to return to basic considerations.

Only white noise of thermal or electronic origin is considered in this treatment. While microphonics, element drift and other effects contribute to the instability of oscillators, their action can be calculated by these or other methods and no specific treatment is given here. The analysis indicates that low frequency abnormalities such as the flicker effect may affect the amplitude distribution of the output but should have only a second order effect upon the frequency.

Narrow Band Noise

Oscillators are characterized by relatively selective circuits and the fact that all the effects of interest are concentrated in a relatively narrow band of frequencies. Therefore, we may restrict our attention to noise within fractionally narrow bandwidths. The voltage which results when white noise is passed through any relatively narrow band-pass filter may be described by

$$v(t) = x_1(t)\cos\omega_0 t + x_2(t)\sin\omega_0 t \qquad (1)$$

where ω_0 is the midband angular frequency and $x_1(t)$

* Original manuscript received by the IRE, June 9, 1959; revised manuscript received, Janaury 6, 1960.
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and $x_2(t)$ are *uncorrelated* functions of time which vary slowly and randomly about zero in a manner described by the normal or Gaussian probability function shown as Fig. 1. That is, the most probable value is zero and during half the time, the function lies within the range of probable error $\pm 0.675\sigma$, where σ, the *standard deviation*, is the rms value. The numerical value of σ depends upon the temperature, impedance, and bandwidth of the circuit and is discussed later.

Although zero is the most probable value for $x_1(t)$ and $x_2(t)$ separately, it is very improbable that the two coefficients are simultaneously zero. That this is true may be seen from the expression equivalent to (1), also given by Davenport and Root, [5],

$$v(t) = \rho(t) \cos [\omega_0 t + \phi(t)]. \tag{2}$$

Again ω_0 is the midband radian frequency and $\rho(t)$ and $\phi(t)$ are slowly varying functions. However, entirely new statistical properties are involved. The phase varies randomly and with equal probabilities in all quadrants: therefore, $\phi(t)$ has a rectangular distribution, uniform throughout the interval $\pm\pi$ radians. The amplitude $\rho(t)$ is now a magnitude or envelope function and has the Rayleigh distribution shown in Fig. 2. The standard deviation has the same significance as before, and during half the total time the envelope amplitude equals or exceeds 1.177 of the rms value. The fact that the same physical phenomenon is correctly described by both Gaussian and Rayleigh distributions is important to all the following discussions.

I. Initiation of Oscillation

In the interval which follows the instant that an oscillator is first turned on, the amplitude builds up from a small initial value until it becomes large enough to produce limiting at some point in the system. During this period, the behavior of typical oscillators is substantially linear. Therefore, it is possible to treat this part of the problem without recourse to nonlinear mathematics.

In low-frequency oscillators, it is very difficult to achieve the turn-on process without simultaneously generating a transient which, within the bandwidth of interest, is large compared with background noise. Under such circumstances, the oscillation starts from this injected transient rather than from noise, and the phase of oscillation is uniquely related to the turn-on signal. Such oscillations are referred to as coherent and are desirable in many situations. However, they are beyond the scope of the present treatment, which is limited to situations in which the starting transient is negligible and oscillation builds up from the white noise signal which is inherent in all physical circuits. Because pure noise has no correlation with respect to the turn-on signal, there is no phase relationship between the phase of the growing oscillation and the turn-on signal. Such oscillations are referred to as incoherent. Phase coherence is a convenient practical test for the existence and relative magnitude of a starting transient.

At best, the problem to be analyzed is complicated and difficult. For this reason the following treatment is divided into several steps of increasing complexity, each introducing some additional consideration and, therefore, being a closer approximation to a physical system.

A. Noise Build-Up in a Passive Circuit

Referring to Fig. 3, assume that L, C, and G are positive fixed elements and that the switch has been closed for a long time. Noise effects identifiable with the conductance G are represented by the current I. The sys-

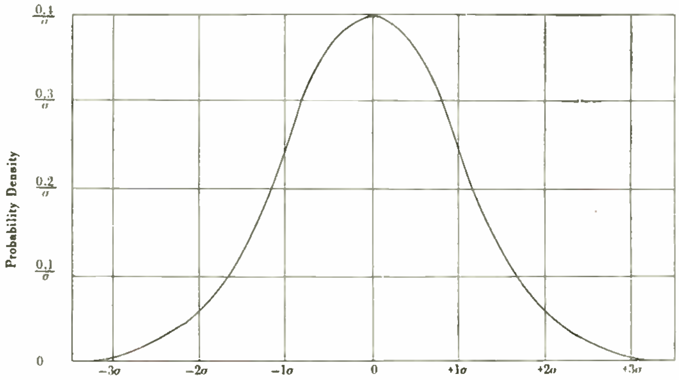

Fig. 1—The Gaussian probability density function

$$p(x) = \frac{1}{\sigma\sqrt{2\pi}} \epsilon^{-x^2/2\sigma^2}.$$

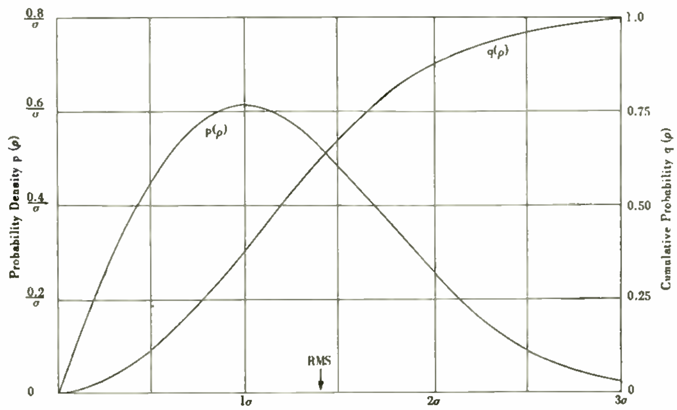

Fig. 2—The Rayleigh probability density function

$$p(\rho) = \frac{\rho}{\sigma^2} \epsilon^{-\rho^2/2\sigma^2}.$$

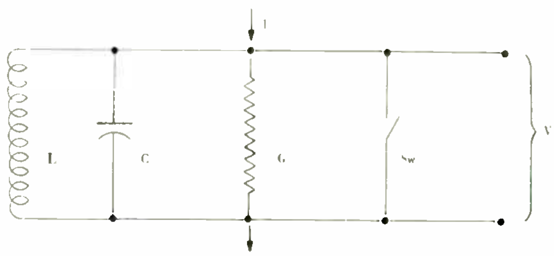

Fig. 3—First circuit model.

tem, which is initially at rest, is driven by the current I in the interval which follows the opening of the switch at time $t=0$.

Normal tube or thermal noise consists of a great many small impulses, randomly distributed in time. However, as already shown, the energy storage in a tuned circuit reduces the rapidity of response which is possible. Therefore, no loss of generality results if the impulses which occur within some finite period are collected or grouped together. For the present purpose, it is convenient to use one-quarter cycle as the collecting interval. On page 31 of Woodyard [17], it is shown that this procedure is correct and that it meets the requirements of sampling theory; this discussion also accounts for the sine and cosine components of (1).

In the present section, it is assumed that the sampling period is perfectly uniform. However, in the final section on phase perturbations this period is modified to conform with the existing phase of the wave as perturbed by preceding impulses. No appreciable error is introduced because the perturbation per cycle is very small.

The situation is best visualized by imagining that a very large number of identical oscillatory systems are subjected simultaneously to the same experiment. It is now possible to make meaningful statements as to the statistical behavior of this ensemble. In the following paragraphs the amplitudes u and v are to be interpreted in this sense.

During the first quarter-cycle after the switch is opened, the noise current will excite in the system a voltage of sine phase with an amplitude identifiable with some mean-square value u^2. In the ensuing quarter-cycle, the noise current excites a voltage of cosine phase and the same mean square amplitude. The initially excited sine term is not affected because sine and cosine terms are orthogonal. Since the sine and cosine components experience similar influences, it is sufficient to explore either. The cosine term is selected because it offers minor simplification of notation.

During the last quarter of the initial cycle, two opposing effects take place. The cosine oscillation excited during the second quarter cycle is decreased in amplitude by a factor β associated with the positive damping of the passive circuit. However, this effect is more than compensated by the additional impulse received from the driving current I. Thus, at the end of the second half-cycle, the amplitude of the cosine component is representable by

$$v_2^2 = u^2(1 + \beta) \tag{3}$$

where β is related to the logarithmic decrement δ by

$$\beta = e^{-\delta} \tag{4}$$

derived from

$$\beta^2 = e^{-2\delta} \tag{5}$$

which more closely represents the actual physical situation. For future reference it is noted that, because δ is small in the cases of interest, (4) may be approximated by

$$1 - \beta = \delta. \tag{6}$$

Consistent with ordinary usage, δ is taken as positive for a passive circuit.

The validity of (3) depends upon the fact that the noise impulses are statistically independent so that the square of a deviation is equal to the sum of the squares of all component deviations.

Iteration of the above reasoning shows that the mean square of the coefficient of the cosine voltage at the end of n half-cycles is represented by the finite sum

$$v_n^2 = u^2(1 + \beta + \beta^2 + \beta^3 + \cdots \beta^{n-1}). \tag{7}$$

The stepwise derivative of this series with respect to n is obtained by noting that, because $dn = 1$,

$$v_{n+1}^2 - v_n^2 = dv_n^2 = u^2\beta^n dn = u^2 e^{-n\delta} dn. \tag{8}$$

Integration gives

$$v_n^2 = K - u^2(e^{-n\delta})/\delta. \tag{9}$$

The constant of integration K is evaluated by fitting the initial condition,

$$v_n = 0 \quad \text{when } n = 0 \text{ with the result that}$$

$$v_n^2 = u^2(1 - e^{-n\delta})/\delta = u^2(1 - \beta^n)/\delta.^1 \tag{10}$$

The noise build-up represented by (7) or (10) is plotted as curve (A) in Fig. 4. It is seen to reach the asymptotic value

$$U^2 = u^2/\delta. \tag{11}$$

However, Pierce, [12], has shown that this equilibrium amplitude is given by

$$U^2 = kT/C \tag{12}$$

where $k = 1.38 \times 10^{-23}$ is Boltzmann's constant, and T is the effective absolute temperature of G. The coefficient in (12) is correct because U corresponds to the peak value of the cosine wave, which in turn is equal to the

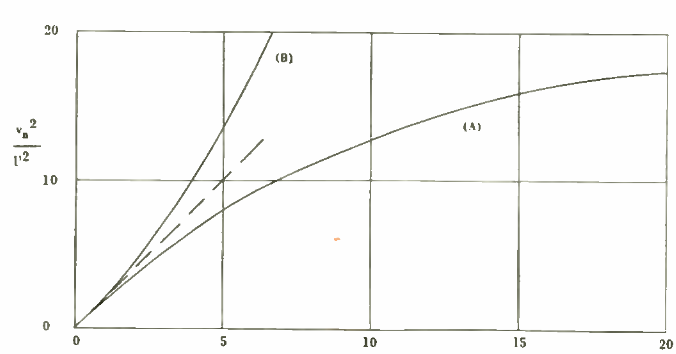

Fig. 4—Growth of the cosine component of voltage: (A) Passive circuit, $Q = +20\pi$; $\delta = +0.05$. (B) Active circuit, $Q = -20\pi$; $\delta = -0.05$.

[1] An alternate derivation of this result is given in Appendix I.

rms value of the total wave. This expression can be put in several equivalent forms by use of the following well-known relationships:

$$\omega_0 = 1/\sqrt{LC}, \tag{13}$$

$$Q = \omega_0 C/G, \tag{14}$$

and

$$\delta = \pi/Q = \pi G/\omega_0 C. \tag{15}$$

The magnitude of the mean-square noise impulse delivered per quarter-cycle is given by eliminating U^2 between (11) and (12) to obtain

$$u^2 = \pi kT/CQ = \pi kTG/\omega_0 C^2 \quad \text{peak volts}^2. \tag{16}$$

This relation is basic to all that follows, and may be taken as the principal result of this section.[2]

B. Noise Build-Up in an Active Circuit

Now assume that G is negative rather than positive but that no other change is made in the system previously described. Because numerical relationships between voltage and current are independent of whether G is positive or negative, it follows that the equations already developed can be used. The essential new fact is that disturbances caused by noise or other influences increase rather than decrease with the passage of time.

Specifically, δ is now inherently negative and β is larger (rather than smaller) than unity. Eqs. (7) and (10) are valid, but now diverge rather than converge. The situation corresponding to the negative selectivity, $Q = -20\pi$, is plotted as curve (B) in Fig. 4. As before, this curve represents the mean square value of the coefficient of the cosine component of voltage, and deviations from this value are given by the probability distribution of Fig. 2.

C. Change of Circuit Parameters

Let us now extend the analysis to include a change of circuit parameters as shown in Fig. 5. It is assumed that L, C and G are all positive and that an initial statistical equilibrium has been reached with respect to the noise current I identified with G at temperature T. The added conductance γG associated with an additional noise current μI is negative and numerically larger than G in the cases of interest, but the analysis is not so limited. In view of the preceding sections, and recalling that a conductance is incapable of energy storage, it is anticipated that subsequent to closing the switch at $t=0$, the voltage across the system will increase with time from some finite initial value.

Evidently, U_0^2, the value of the cosine voltage at time $t=0$, is correctly given by (12), and the parameter β which identifies the rate of growth is given by

$$\beta = e^{-\pi(1+\gamma)/Q}. \tag{17}$$

[2] Here and throughout the remainder of this paper, noise power is identified with positive frequencies, and no use is made of the negative frequency concept.

The total noise impulse received per quarter-cycle is, from (16),

$$u^2 = (\pi kT/CQ)(1 + \mu^2). \tag{18}$$

Therefore, the cosine component of voltage which exists at the end of n half-cycles after the switch is closed is

$$v_n^2 = U_0^2\beta^n + u^2(1 + \beta + \beta^2 + \cdots + \beta^{n-1}). \tag{19}$$

This equation may be converted with use of (10) to

$$v_n^2 = U_0^2\beta^n + \frac{u^2}{\delta}(1 - \beta^n) = \frac{u^2}{\delta} + \left(U_0^2 - \frac{u^2}{\delta}\right)\beta^n \tag{20}$$

where β, u, and δ have the values given in (17) and (18).

To illustrate the application of these results, assume that in Fig. 5, $\gamma = -2$ and $\mu = 1$. Further assuming that the passive circuit has a selectivity of $Q = 20\pi$, it can be determined from (17) that $\delta = -0.05$ and $\beta = 1.0514$. Consistent with (18), $u^2 = 0.1\,U_0^2$. The result of substituting these values in (20) yields the expression $v_n^2/U_0^2 = -2 + 3(1.0514)^n$, shown as curve (A) in Fig. 6. The slope is relatively steep for small values of n because noise impulses add to the exponential growth due to the negative conductance. After about ten cycles, the effect of additional noise impulses becomes negligible and the plot becomes essentially straight on the semilog coordinates used.

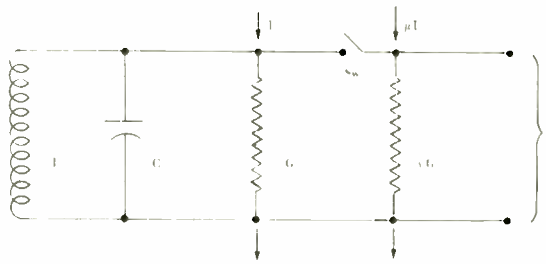

Fig. 5—Second circuit model.

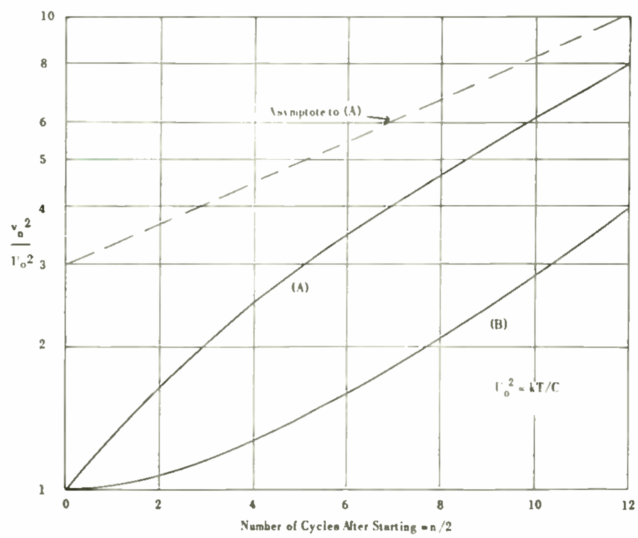

Fig. 6—Growth of the cosine component of voltage: (A) Parameters changed discontinuously. (B) Parameters changed continuously.

D. Build-Up with Continuously Variable Parameters

The analyses of the foregoing sections have in common the underlying assumption that the parameters of the system are changed discontinuously, that is, in a time which is short compared with one cycle of the oscillation. In practice, such a sudden change would be associated with transients which are large compared with the level of noise. Moreover, no means is known for producing switching rapid enough to meet this criterion at microwave frequencies. Therefore, the concern here is with systems in which the total conductance is changed in some orderly manner from a positive to a negative value. Because such a change is ordinarily made by means of an electronic device and because appreciable noise contributions are associated with such devices, it is quite desirable to obtain an analysis which permits time variation of the noise current.

The system under consideration is shown in Fig. 7 which differs from Fig. 3 only in that G and I are both functions of time. At time $t=0$, the voltage has an initial equilibrium value U_0 identifiable with the system comprised of L, C, G_0 and I_0, where G_0 and I_0 refer to the values of these parameters during all negative values of time. Starting at time $t=0$, both G and I change, ordinarily in such a way as to increase I and to decrease G from a positive to a negative value. No serious restriction of generality results from the assumption that neither G nor I changes appreciably during any one cycle of the oscillation. Therefore, the convolution integral and methods of linear, time-variable, systems may be used. The analysis of superregeneration by Wheeler [16], and others has much in common with the following discussion.

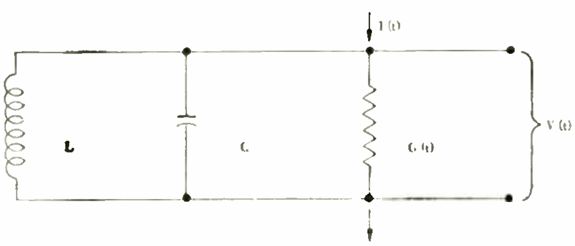

Fig. 7—Third circuit model.

Drawing on the results derived in preceding sections, the following expression can be written for the cosine component of noise in Fig. 7 at the end of n half-cycles:

$$\begin{aligned} v_n^2 = {} & U_0^2(\beta_1\cdot\beta_2\cdot\beta_3\cdot\beta_4 \cdots \beta_n) \\ & + u_1^2(\beta_2\cdot\beta_3\cdot\beta_4 \cdots \beta_n) \\ & + u_2^2(\beta_3\cdot\beta_4 \cdots \beta_n) \\ & + u_3^2(\beta_4 \cdots \beta_n) \\ & + \cdots\cdots\cdots \\ & + \cdots\cdots \\ & + u_n^2 \end{aligned} \tag{21}$$

where u_n and β_n refer to the noise impulse and the expansion factor applicable to the nth half cycle.

This expression may be written more compactly as

$$v_n^2 = \sum_{i=0}^{i=n} u_i^2 \prod_{j=i+1}^{j=n} \beta_j \tag{22}$$

where the initial condition is represented by the special convention $u_0^2 = U_0^2$.

Computation is facilitated by using an expression equivalent to (21) which has the form

$$v_n^2 = \beta_n v_{n-1}^2 + u_n^2. \tag{23}$$

The net change between v_n^2 and v_{n-1}^2 is

$$\frac{dv_n^2}{dn} = (1 - \beta_n)v_{n-1}^2 + u_n^2. \tag{24}$$

Because this equation is not readily integrated in its general form, a typical numerical example will be used. Assume that the initial conditions are the same as those in Fig. 6 and that the noise and negative conductance both increase (almost) linearly with time so that ten full cycles are required to carry out the transition which is achieved stepwise by switching in Fig. 5.

Consistent with the above,

$$u_n^2 = 0.05U_0^2\left(1 + \frac{n}{20}\right) \tag{25}$$

and

$$\beta_n = 0.95 + n/200. \tag{26}$$

By definition,

$$v_0^2 = U_0^2. \tag{27}$$

Therefore, iterative use of (23), (25) and (26) permits evaluation of v_n^2 for any desired value of n. The results obtained by this method are plotted as curve (B) in Fig. 6. It is noted that the slope of curve (B) at the end of ten cycles is the same as that of curve (A) for the same ordinate, and barring further changes in u and β, the two are simply displaced in time for all larger amplitudes.

E. Generalizations on Build-Up

Each of the foregoing analyses has led to a Gaussian distribution for the separate sine and cosine components of voltage. Hence, we conclude that for all oscillators of this type and at all times during the build-up interval, the distribution of the separate components is Gaussian and the distribution of the total amplitude is Rayleigh. This result, which follows from the fact that successive impulses are statistically independent, is quite general, and is later used to determine a distribution function which describes the jitter in the starting time of an ensemble of oscillators.

An additional feature of oscillation build-up is worthy of mention. Prior to turn-on, the noise voltage of a given oscillator remains small and changes randomly throughout a range comparable with its greatest magnitude. Likewise, the phase is subject to large random excursions. After turn-on, the level increases in a generally exponential manner to a relatively large value. Because the perturbing effects of noise remain nearly constant, as growth occurs it becomes increasingly less probable that the phase will be greatly shifted or that an oscillator which started with a smaller-than-normal amplitude will be able to advance into a larger-than-normal amplitude class. Explicitly, in terms of Fig. 6, increase of time makes it less and less likely that a given oscillator can move from one group to another. The situation might be paraphrased as "the social strata become more firmly fixed as their general level rises."

To obtain a numerical measure of this decreasing mobility of level, a digression is in order to introduce the autocorrelation function, which is a widely used measure of predictability. The formal definition for the autocorrelation of a stationary statistical function for the time interval τ is

$$R(\tau) = \lim_{T\to\infty} \frac{1}{T} \int_0^T v(t)v(t+\tau)dt. \tag{28}$$

For steady-state noise in a passive (rectangular) narrow band circuit, (156) of Bennett, [1], may be converted to the form

$$R(\tau) = \frac{2U^2GQ}{\omega_0\tau} \sin \frac{\omega_0\tau}{2Q} \cos \omega_0\tau. \tag{29}$$

For the present purpose, it is convenient to consider the period of one full cycle and to introduce as the *autocorrelation per cycle*,

$$R_1 = (U^2GQ/\pi)(\sin \pi/Q). \tag{30}$$

This is recognized as the product of a (sine x)/x function and the mean-square power in the circuit. Interest centers in the extent to which the (sine x)/x function differs from unity. Eliminating Q with the use of (11) and (15), this expression can be converted to

$$R_1 = \frac{U^4G}{u^2} \sin \frac{u^2}{U^2} \cdot \tag{31}$$

Specifically, this is the predictability during the period of one cycle of a function having a Rayleigh distribution with mean-square magnitude U^2 and perturbed by noise impulses of magnitude u^2.

Returning to a consideration of the build-up process in an active circuit, it is seen that here, too, are found a Rayleigh amplitude distribution, a mean square amplitude which changes relatively little during any one cycle, and noise impulses measured by u^2. Therefore, it is concluded that (31) also represents the autocorrelation per cycle of an expanding wave, provided U^2 is interpreted as the mean-square amplitude of the total wave over the cycle in question, and u^2 represents the mean-square value of the impulses received during that cycle.

The result just derived does not distinguish between amplitude and phase perturbations, nor is it particularly easy to interpret. Fortunately, relevant statistics for both amplitude and phase distributions have been worked out and are tabulated. The essential fact is that, during any one cycle, a sinusoidal oscillation representable as a unique phasor is modified by the addition of another phasor which is randomly distributed. The situation is illustrated by Fig. 8, which shows in an exaggerated form the effect produced upon the oscillation phasor by noise impulses received during a single cycle. The extensive curves published by K. A. Norton, *et al.*, [9] [10], in another connection are directly applicable to the present problem.

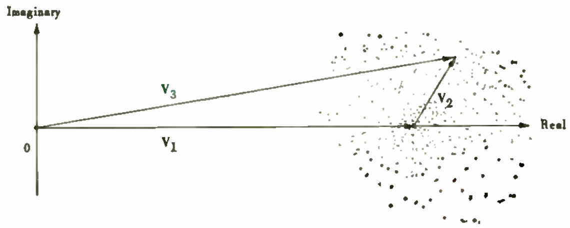

Fig. 8—Cosine wave plus random noise impulse.

F. Jitter in Starting Time

The starting time of an oscillator may be defined as the time after turn-on required for the voltage to build up to some preassigned reference value v_r which is below saturation but large compared to the level of noise. The starting time of a given oscillator for any single turn-on is a perfectly definite and exact quantity. However, this time varies in a statistical manner if the same oscillator is repeatedly tested, or if an ensemble of many oscillators is tested once.

To determine this distribution, it is noted that in the region of interest the level of a given oscillator increases exponentially with time. Therefore, the concern here is with amplitude distributions expressed in terms of a logarithmic (neper) scale rather than a linear scale.

The situation is illustrated by Fig. 9, which shows the growth of voltage toward the reference level. Like Fig. 6, it is plotted to semilog scales. However, several lines representing the cumulative probability distribution replace the single mean-square line of build-up. The superimposed curve at the left, representing the cumulative probability distribution, is replotted from Fig. 2. The shape of this curve is markedly altered by the change from a linear to a logarithmic abscissa. However, no mathematical subtleties are involved in this transformation of variable.

The construction of Fig. 9 makes clear the fact that the cumulative probability distribution of starting times (the upper superimposed curve) has the same shape as that for the logarithm of the voltages. Therefore, the cumulative Rayleigh distribution plotted against a suitable logarithmic abscissa represents the universal distribution of jitter in starting time. This curve is plotted upright and on a general basis as curve (B) of Fig. 10. The derivative of this curve, which represents the relative probability density distribution, is plotted as curve (A).

The reversal of the time and voltage scales of Fig. 10 arises from the fact that oscillators with relatively large amplitudes have shorter starting times than those with smaller amplitudes. Moreover, the effect of the logarithmic voltage scale is to inhibit short build-up times and to increase the number of long build-ups.

The numerical factor between the two abscissa scales of Fig. 10 derives from the exponential build-up equation. Setting[3]

$$\ln (v/v_r) = n\delta/2 = 1, \tag{32}$$

$$\tau = n/2F, \quad = n\pi/\omega_0 \tag{33}$$

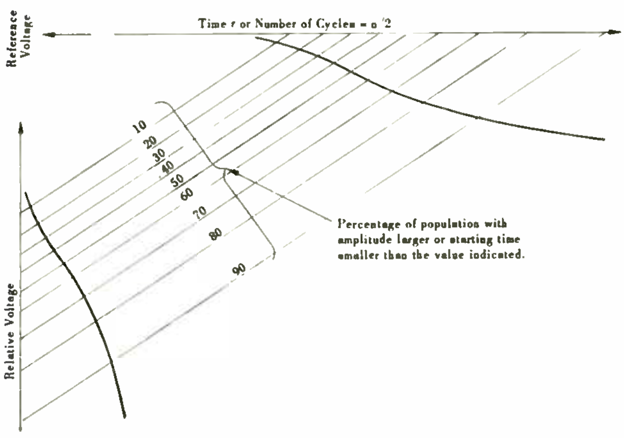

Fig. 9—Jitter in starting time caused by Rayleigh amplitude distribution.

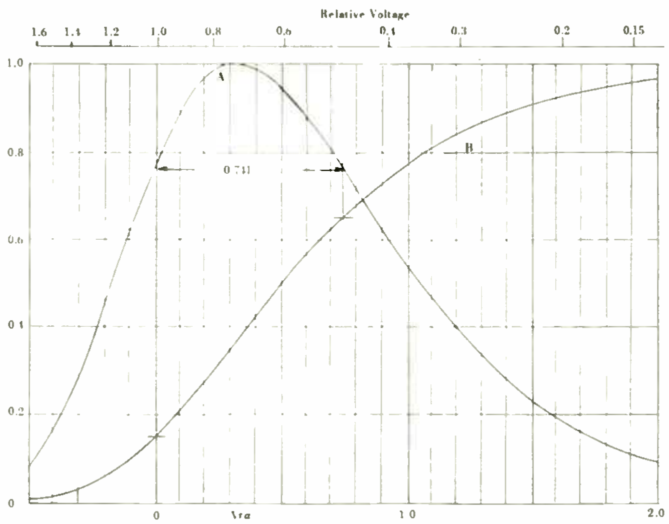

Fig. 10—Distribution of starting time jitter. (A) Relative probability density. (B) Cumulative probability.

[3] The symbol σ which would normally be used in (34) has already been used for the standard derivation and is therefore replaced by α.

and

$$\alpha = -F\delta, \tag{34}$$

one sees that a voltage ratio of one neper corresponds to a time interval

$$\Delta\tau = 1/\alpha. \tag{35}$$

Because the density distribution curve (A) of Fig. 10 is skewed, it is necessary to treat it with some caution. Exploring the steepest portion of curve (B), we see that 50 per cent of the total population (15 per cent to 65 per cent) is included in a voltage ratio of 2.09 corresponding to 0.741 neper. The associated time jitter, which is the principal result of this section is

$$\tau_j = 0.741/\alpha \text{ seconds.} \tag{36}$$

Ignoring the question of skew distribution, this value is twice the probable error of an observed starting time.

The important conclusion is that *the amount of jitter in the starting time of an oscillator varies inversely with the expansion coefficient α and is independent of all other influences.* In particular, reduction of the noise level of the oscillator affords no relief at all. Except for a slight numerical refinement, this result is in agreement with one previously derived, [6], by a much less rigorous argument.

In certain situations the magnitude of the starting transient produced by switching on an oscillator is comparable to the noise level. Under these conditions, the amount of jitter and the randomness of starting phase are reduced. It is clear that the foregoing methods, together with the results of Norton, *et al.*, [9], [10], are sufficient to calculate the behavior which will occur under such circumstances.

Finally, for the sake of emphasis, it is repeated that none of the foregoing results predict the behavior of a particular oscillator when turned on once; they correctly represent the statistical behavior of a given oscillator subjected to many repetitive tests or to the behavior of a great many identical oscillators tested simultaneously. The behavior of a given oscillator is thoroughly random and disorderly in the interval preceding and immediately following turn-on. However, as build-up occurs, the noise disturbances become relatively less significant, and the growth is more and more accurately exponential. The growth period comes to an end when saturation sets in and brings about the equilibrium condition commonly referred to as sustained oscillation which is treated in the following section. The essential new fact is that nonlinearity now couples the sine and cosine components so that they are no longer statistically independent. Fortunately, amplitude and phase components of an envelope function *are independent,* even in a nonlinear system. The results obtained in the following section are based upon this fact.

II. Sustained Oscillation

A. Amplitude Perturbation

In the absence of noise, an oscillator would generate a succession of nearly sinusoidal cycles identical in amplitude, period and shape. The presence of noise perturbs this situation slightly, making both period and amplitude vary randomly about the previous undisturbed value. To evaluate the magnitude of this disturbance, it should be noted that the amplitude of any oscillator changes slightly when a locking or synchronizing signal is injected. If the injected signal is small and has exactly the same frequency and phase as that of the spontaneous oscillation, then the amplitude change is directly proportioned to the input amplitude. This condition is representable by

$$sG = I/\Delta V \tag{37}$$

where ΔV is the increase in amplitude which results from injection of a synchronizing current I. Here G is the normal circuit conductance and s is a dimensionless parameter which depends upon the degree of limiting or saturation and is of the order of ten in typical oscillators. The product sG represents an equivalent conductance closely related to one of the compliance coefficients of Huntoon and Weiss, [8]. Additional insight into the situation is gained by noting that if it is increased or decreased slightly from its normal value, the amplitude of a free-running oscillator relaxes exponentially toward its normal value at a rate corresponding to the time constant C/sG where C is as usual the capacitance of the oscillatory circuit. The introduction of the parameter s represents one form of the process of equivalent linearization. Because this step is basic to what follows, it is well to examine its validity. Experience shows that within a narrow range of amplitude, the deviation *is* accurately proportional to the perturbing influence and that the deviations of interest *are* very small. Therefore, the procedure *is* valid. The value of s depends upon the passive components of the system, the characteristics of the tube or other nonlinear elements, and the magnitude of the applied voltages. Thus, while a unique value of s exists for each condition of oscillation, it may be hard to evaluate in practical situations.

The treatment is directly applicable to conventional self-limiting oscillators. Oscillators employing automatic output control or other sophisticated forms of limiting will relax toward the normal level of oscillation in a somewhat more complicated manner. Therefore, the amplitude perturbations calculated by the following method are only approximate for unconventional forms of oscillators.

Let us now assume that a noise-free system is in the condition of sustained oscillation described by

$$v = A \cos \omega t, \tag{38}$$

and that starting at time $t=0$, noise consisting of a series of impulses is introduced. Consistent with the methods developed in Section I on the Initiation of Oscillation, consider first only those impulses occurring at odd quarter-cycles and generating cosine-phase voltages.

In view of the foregoing discussion, it can be seen that the result of the noise is the addition to the original voltage wave of another which has the statistical properties associated with the given noise impulses operating on a passive equivalent circuit consisting of L, C, and sG. That is, the cosine-phase noise impulses act to amplitude-modulate the original wave.

Consistent with (11), the mean-square value of the envelope deviation reached in the steady state is

$$U^2 = u^2/s\delta, \tag{39}$$

where U^2 and δ have the values given in (16) and (15), respectively.

Elimination of undesired terms yields, as the mean-square voltage deviation of the oscillation envelope from the unperturbed value,

$$V^2 = kT/sC \tag{40}$$

where T is the effective temperature including all noise sources, and s is the limiting or saturation factor. It is seen that this expression differs from (12) only by the factor s.

The derivation of (40) shows that the situation is exactly the same as if noise currents produced by a conductance G at temperature T acted upon an antiresonant circuit consisting of L, C and a conductance sG. Therefore, the resulting spectrum is lower and broader by a factor s than that associated with the corresponding passive circuit. Taking this factor into account, together with the quantity 2 arising from the fact that only the cosine component is involved, an equation from Pierce, [12], can be modified to read

$$V_1^2(\omega) = \frac{2kTG}{s^2G^2 + 4C^2(\omega - \omega_0)^2} \text{ effective volts}^2 \text{ per cycle.} \tag{41}$$

This equation is regarded as one of the principal results of the present paper, and, as will be proven later, is quite general. The present simple derivation gives correct results because amplitude and phase disturbances are independent, even though the system is nonlinear.

The spectral distribution of the envelope function is often of interest. Because the upper and lower sidebands are coherent and add directly, voltage must be superimposed rather than power. The desired expression is

$$V_e^2 = \frac{8kTG}{s^2G^2 + 4\omega_e^2C^2} \text{ effective volts}^2 \text{ per cycle,} \tag{42}$$

where $\omega_e = (\omega - \omega_0)$ is the envelope angular frequency.

In anticipation of future use, consider the autocorrelation function corresponding to the envelope distribution of (42). The needed partial autocorrelation function is given by the definition.

$$R_1(\tau) = \frac{1}{2\pi}\int_0^\infty v^2(\omega)\cos(\omega\tau)d\omega. \tag{43}$$

Eq. (42), with special consideration of the origin, results in

$$R_1(\tau) = \frac{A^2}{2} + \frac{4}{\pi}\int_0^\infty \frac{kTG}{s^2G^2 + 4\omega_e{}^2C^2}\cos(\omega_e\tau)d\omega_e \tag{44}$$

$$= \frac{A^2}{2} + \frac{4kT}{\pi s^2 G}\int_0^\infty \frac{\cos(\omega_e\tau)d\omega_e}{1 + 4\omega_e{}^2C^2/s^2G^2}. \tag{45}$$

Using Peirce, [11], formula 490, letting

$$x = 2\omega_e C/sG, \tag{46}$$

and normalizing to $A^2/2$ yields

$$R_1(\tau) = 1 + \frac{2kT}{A^2 sC}e^{-(sG\tau/2C)}. \tag{47}$$

In all typical situations the first term, specifying the nominal power output, is very large compared with the second term which describes amplitude fluctuations.

B. Phase Perturbation

To evaluate the phase perturbations of the oscillator output we consider the sine-phase noise produced by impulses occurring in even-numbered quarter cycles. The situation is best visualized in terms of Fig. 8 showing a phasor, which, in a noise-free system, would have constant length and a constant angular velocity. Length perturbations are counteracted by the inherent limiting action, and produce the random amplitude modulation described in the previous section.

In contrast, no mechanism exists to counteract phase perturbations. The situation corresponds to a well-regulated clock; if once set forward it continues to read fast until it is reset. The sine-phase noise impulses consitute a series of random clock-settings, which produce much the same effect as a random deviation of rate.

Using (16) we obtain, as the standard deviation or rms value of the phase error accumulated in the period of n half-cycles, the value

$$\phi_n = \sqrt{n}u/A = \sqrt{\pi nkTG/\omega_0 C^2 A^2}\text{ rms radians.} \tag{48}$$

It is clear that this result is valid for small values of n because $u \ll A$ and the angle ϕ is closely equal to its tangent. For large values of n the accumulated phase error also becomes large, and it might be thought that the method would fail. Actually (48) is valid for all values of n. The situation is saved by the fact that noise impulses continually and naturally resolve themselves parallel and perpendicular to the instantaneous direction of the phasor. This corresponds to the small and gradual readjustment of the sampling interval mentioned in Section I-A.

To obtain a basis for converting (48) to other more useful forms we note that the total generated power is given by

$$P = A^2G/2. \tag{49}$$

Eqs. (14) and (33) may be used with (49) to convert (48) to the convenient form

$$\phi_n = \sqrt{\pi n\omega_0 kT/2PQ^2} = \sqrt{\omega_0{}^2 kT\tau/2PQ^2}\text{ rms radians.} \tag{50}$$

Consistent with (7), the standard deviation of the cumulative phase error increases with the square root of the number of cycles (or half-cycles) which elapse. Thus, while the probable *phase error* increases without limit, the probable *frequency error* approaches zero over a long period of time because it varies in the manner $\sqrt{n}/n = 1/\sqrt{n}$.

In an interval of n half-cycles, the total average phase will advance by $n\pi$ radians. Thus, the standard deviation or rms value of the fractional phase or frequency in an interval τ is

$$\phi_n/\phi = \sqrt{\omega_0 kT/2n\pi PQ^2} = \sqrt{kT/2\tau PQ^2}. \tag{51}$$

It is significant and perhaps surprising that ω_0 does not appear in this expression; (51) is important because it represents the ultimate limit of the accuracy with which the frequency of an oscillator may be observed in a given period τ.

A useful measure of frequency indeterminacy may be defined as the *period of coherence*. This is the time interval τ_c required for the standard deviation of the phase to increase to a full radian. Setting (50) equal to one radian, the period of coherence may be stated

$$\tau_c = 2PQ^2/kT\omega_0{}^2. \tag{52}$$

This period, which is quite long for typical oscillators, is the interval during which the rms phase deviation between an actual and an ideal oscillator would reach one radian, and represents approximately the longest period over which a useful measurement may be made by the zero-beat method.

The autocorrelation is established as the most meaningful measure of coherence. Referring only to phase perturbations,

$$R_2(\tau) = \frac{A^2}{2}\overline{\cos\{\omega_0 t + \phi(t)\}\cos\{\omega_0[t+\tau] + \phi(t+\tau)\}} \tag{53}$$

where the bar indicates averaging over the ensemble for the time interval τ. This expression is equivalent to

$$R_2(\tau) = \frac{A^2}{2}\,\overline{\cos\phi(t)\cos[\omega_0\tau + \phi(t+\tau)] - \sin\phi(t)\sin[\omega_0\tau + \phi(t+\tau)]}, \tag{54}$$

$$\frac{A^2}{2}\,\overline{\cos[\omega_0\tau + \phi(t+\tau) - \phi(t)]} \tag{55}$$

$$= \frac{A^2}{2}\cos\omega_0\tau\,\overline{\cos[\phi(t+\tau) - \phi(t)]} - \sin\omega_0\tau\,\overline{\sin[\phi(t+\tau) - \phi(t)]}. \tag{56}$$

It has already been shown that $[\phi(t+\tau)-\phi(t)]$ is a random variable with a standard deviation given by (50). Since the argument has a symmetrical distribution, the term $\overline{\sin[\phi(t+\tau)-\phi(t)]}$ is equal to zero, and the desired autocorrelation simplifies to

$$R_2(\tau) = \frac{A^2}{2}\cos(\omega_0\tau)\cdot\overline{\cos[\phi(t+\tau) - \phi(t)]}. \tag{57}$$

To evaluate the last term, a cosine weighting function is applied to the Gaussian phase distribution. The probability distribution of the phase deviation ϕ accumulated in a time interval τ may be written in the form

$$p = be^{-a^2\phi^2}, \tag{58}$$

where, from (50) and the definition of the normal distribution,

$$b = \sqrt{PQ^2/\pi\omega_0^2\tau kT} \tag{59}$$

and

$$a^2 = PQ^2/\omega_0^2\tau kT. \tag{60}$$

Thus,

$$\overline{\cos[\phi(t+\tau) - \phi(t)]} = b\int_{-\infty}^{\infty} e^{-a^2\phi^2}\cos\phi\, d\phi. \tag{61}$$

Use of B. O. Peirce, [11], formula 508 reduces this expression to

$$\overline{\cos[\phi(t+\tau) - \phi(t)]} = \frac{b\sqrt{\pi}}{a}e^{-(1/4a^2)}. \tag{62}$$

As a result, the autocorrelation with respect to phase fluctuations is

$$R_2(\tau) = \frac{A^2}{2}\cos(\omega_0\tau)e^{-\omega_0^2\tau kT/4PQ^2}. \tag{63}$$

To obtain the corresponding spectral distribution, the following identity is used:

$$v^2(\omega) = 4\int_0^{\infty} R(\tau)\cos(\omega\tau)d\tau. \tag{64}$$

Substitution of (63) in this expression yields

$$v_2^2(\omega) = 2A^2\int_0^{\infty}\cos(\omega_0\tau)\cos(\omega\tau)e^{-\omega_0^2\tau kT/4PQ^2}d\tau. \tag{65}$$

This formidable expression is evaluated by the use of (7), section 262 of Bierens de Haan, [3], which gives

$$\int_0^{\infty} e^{-px}\cos(qx)\cos(rx)dx = \frac{p^2 + q^2 + r^2}{[p^2 + (q-r)^2][p^2 + (q+r)^2]}, \tag{66}$$

where

$$r = \omega, \qquad q = \omega_0 \quad \text{and} \quad p = \omega_0^2 kT/4PQ^2.$$

In typical situations p is very small compared to q and r. Moreover, since the interest here is in the narrow-band situation, ω may be replaced by ω_0 except where the difference occurs. With these simplifications (66) reduces to

$$v_2^2(\omega) = \frac{pA^2}{[p^2 + (q-r)^2]}. \tag{67}$$

By substituting and using (14) and (49),

$$v_2^2(\omega) = \frac{2kTG}{(\omega_0^2 CkT/2PQ^2)^2 + 4C^2(\omega_0 - \omega)^2}$$

effective volts² per cycle. (68)

From this expression, by doubling the denominator term, the full noise bandwidth[4] is obtained:

$$\delta\omega = \omega_0^2 kT/2PQ^2 \text{ (radians per second)}, \tag{69}$$

which is the reciprocal of the period of coherence τ_c given by (52).

Finally, the fractional bandwidth of phase noise is

$$\delta\omega/\omega_0 = \omega_0 kT/2PQ^2. \tag{70}$$

These three equations constitute the principal result of the present section. Because the numerators and second denominator terms of (41) and (68) are identical, it follows that, at frequencies remote from the carrier, the disturbances caused by amplitude and phase fluctuations are equal.

[4] This result is exactly half as large as that given as (15.21) in [6], which was derived in a much simpler manner. The discrepancy is attributed to the fact that the earlier treatment included equally the effects of amplitude and phase fluctuations.

C. Complete Spectrum

The product of (47) and (63) gives the following equation as the autocorrelation function for the complete wave.

$$R(\tau) = \frac{A^2}{2}\cos(\omega_0\tau)e^{-(\omega_0^2\tau kT/4PQ^2)} + \frac{kT}{sC}e^{-(sG\tau/2C+\omega_0^2\tau kT/4PQ^2)}. \tag{71}$$

As noted previously, in all practical situations the coefficient of the first term is extremely large compared with the coefficient of the second term. Likewise, the first term of the exponent in the second term is large compared with its associated quantity. In the light of these statements, the operation indicated by (71) for determining the complete spectrum simply leads to the sum of (41) and (68). Therefore, it is concluded that the total spectrum of an oscillator represents the sum of two expressions already derived. These are similar in form but have very different proportions. The amplitude spectrum which is very weak is also extremely broad, whereas the phase perturbed spectrum contains nearly all the energy of the wave in a very narrow spectrum. Because the derivation of the phase-perturbed spectrum does not involve nonlinearity, the results are applicable to all kinds of oscillators, including those using automatic output control.

III. Harmful Effects of Noise

A. General Discussion

The superheterodyne method is used in nearly all kinds of receivers to simplify the problem of obtaining the needed amplification and selectivity.

In typical situations, the local oscillator supplies a voltage large compared with that to be received, and the desired difference frequency is produced in some sort of nonlinear element, usually a rectifier. Such nonlinear elements preserve phase relationships, so that any irregularity in the phase of the local oscillator appears at the output, indistinguishable from phase irregularities in the input signal. That is, phase noise of the local oscillator is fully preserved.

Receivers for frequency-modulated waves usually employ clipping in the intermediate-frequency amplifier. Such a clipper removes amplitude fluctuations of all kinds, whether caused by the local oscillator or otherwise. In view of these facts, it is concluded that *for receivers of frequency modulated signals, the amplitude fluctuations of the local oscillator have no significance and any noise contributed to the output is caused by phase perturbation and is calculable by use of (50).*

Just the opposite situation exists in receivers for amplitude modulated signals. To show this, assume that no signal is present and note that for small variations of the local oscillator level the rectified output varies in a linear manner with changes of the oscillator level. Stated in a different way, amplitude perturbations in the output of the local oscillator are indistinguishable from the desired signal. Phase perturbations have substantially no effect because the rectification process (in the absence of a signal) is insensitive to phase. Thus, the threshold sensitivity of a receiver for amplitude modulated signals is affected only by amplitude perturbations described by (42). In particular, it is noted that knowledge of the complete noise spectrum is neither necessary nor desirable for the calculation of local oscillator noise. The reason is that correlation exists between noise components in the upper and lower sidebands, both of which are ordinarily applied to the nonlinear element. Therefore, the net result differs from what would be anticipated on the basis of simple spectrum considerations.

B. Numerical Examples

To appraise the implications of the foregoing analysis, some numerical values should be obtained. As two rather extreme examples, consider a K-band reflex klystron of the variety commonly used for local oscillators and a low-frequency precision oscillator using a very excellent quartz crystal resonator. Numerical values typical of such oscillators are given in Table I. The very low value of power attributed to the low-frequency oscillator is identified with the fact that precision crystals are adversely affected by higher powers. The capacitance value is an equivalent value appropriate for the present purposes.

Substitution in various formulas yields the results presented in Table II.

Suppose that a radar receiver uses a simple superheterodyne arrangement, and that the first mixer may be taken as an ideal noise-free diode. Thus, assume that in the absence of a received signal the noise voltage delivered to the input of the IF amplifier is simply the envelope of the wave developed by the local oscillator. Substituting in (42) the values given in Table I for the klystron and choosing a nominal intermediate frequency of 30 mc,

$$v_s^2 \approx 3 \times 10^{-16} \text{ volts}^2 \text{ per cycle.} \tag{72}$$

This value changes relatively slowly in the region of 30 mc. Therefore, in a 5-mc band centered on 30,

$$\bar{v}^2 = 1.4 \times 10^{-9} \text{ volts}^2 \quad \text{or} \quad \bar{v} = 3.8 \times 10^{-5} \text{ volts.} \tag{73}$$

Because in typical cases the level of oscillation is about five times higher than that applied to the mixing rectifier, 6×10^{-6} volts may be taken as approximating the threshold of sensitivity of the receiver. Assuming that the impedance level of the signal source is 100 ohms, the limiting power sensitivity is 4×10^{-13} watts or -94

TABLE I

Oscillator Parameters

	Klystron	Crystal
T degrees K	2×10^4	4×10^2
C farads	2×10^{-12}	4×10^{-5}
G mhos	2×10^{-4}	2×10^{-4}
s	5	5
ω_0 radians per second	10^{11}	10^7
Q	10^3	2×10^6
P watts	5×10^{-2}	10^{-5}

TABLE II

Oscillator Noise Characteristics

	Klystron	Quartz
Bandwidth envelope noise, radians per second	5×10^8	25
Bandwidth phase noise, radians per second	2.76×10^{-2}	6.9×10^{-15}
Envelope deviation, rms volts	1.66×10^{-4}	5.25×10^{-9}
Standard deviation of frequency per second	1.66×10^{-12}	8.3×10^{-15}
Period of coherence, seconds	36.2	1.45×10^{14}
Fractional phase noise bandwidth	2.76×10^{-13}	6.9×10^{-22}

dbm, a plausible figure. The sensitivity of a practical receiver using such a local oscillator will be inferior to this figure because of signal loss and noise in the converter. Conversely, as is well known, nearly all the noise caused by the local oscillator may be eliminated by use of a balanced mixer. Therefore, other factors limit the sensitivity of typical practical radar receivers.

IV. Effect of an Auxiliary Filter

It is clear that any kind of imperfect spectrum can be improved by passage through a sufficiently narrow filter. This is true because the energy storage of the filter elements tends to remove existing perturbations in both amplitude and phase. Because the primary spectrum of practical oscillators is very narrow, one might think that no substantial improvement could be obtained with filters of conventional selectivity. This pessimism is unjustified because the most troublesome effects come from perturbations which occur in relatively short time intervals and therefore correspond to relatively broad spectra.

To explore this situation, refer to Fig. 11 which differs from Fig. 3 by the addition of the load circuit ($L'C'$, G'), which is tuned to ω_0 and presumed to have such a high impedance that the noise characteristics of the voltage V are unaffected. Interest centers in the spectral purity of I' as compared with that of V. Evidently, an important parameter is the selectivity of the tuned load circuit in relationship to that of the basic oscillator. It is readily seen that rapid variations of amplitude or phase are smoothed out by the tuned load circuit but that slow changes are little affected. Stated another way, the added filtering removes remote spectral components but transmits nearby ones.

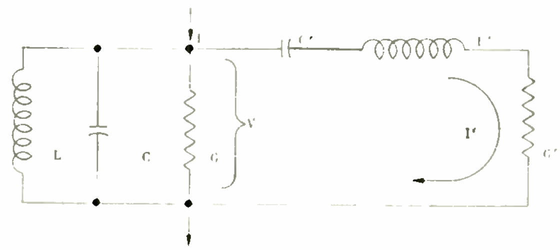

Fig. 11—Effect of tuned load circuit.

The foregoing analysis has shown that at frequencies far from the nominal frequency of oscillation, the spectrum of an oscillator is correctly represented by calculating the filtering effect of the resonator upon the smooth noise spectrum of the driving current I. This idea is readily extended to cover the present situation. The relative suppression of noise in I' with respect to that in V is obtained directly from the selectivity of the added elements.

The use of auxiliary filtering in low-frequency oscillators is rarely necessary. However, at microwave frequencies it is considerably more attractive. This is true for two principal reasons. Microwave oscillators such as magnetrons and reflex klystrons are inherently quite noisy, and mechanical construction problems often preclude the attainment of a large value of Q in the basic resonator. Under these circumstances, it is often possible to obtain substantial benefits from the addition of an auxiliary high-Q resonator. Ordinarily, the added cavity is tightly coupled to the other so that a single resonance results. In effect, large values of inductive and capacitive susceptance are added to the basic circuit without appreciably affecting the inherent impedance level. This arrangement increases the long-term stability of the system to the extent that the added susceptances are more stable than those of the original internal resonator. However, superior performance as a local oscillator is to be expected if the auxiliary cavity is loosely coupled. This is true because the attenuation of remote spectral components produced by the two resonators is compounded in nepers or db in this situation, whereas the tightly coupled arrangement produces a fixed additive figure which is only 6 db if the effect of the added resonator is to double the working Q. Thus, the choice of loose vs tight coupling depends upon whether interest centers in slow or fast perturbations of the output amplitude and phase.

Appendix I[5]

Let $I(t)$ represent the typical member of the noise current generator ensemble and $g(t)$ the impulse response for $I(t)$ applied at time $t=0$ is given by

$$V(t) = \int_0^t v(x)g(t-x)dx.$$

The ensemble average of the squared voltage at fixed time t is then

$$\text{av } V^2(t) = \int_0^t v(x)g(t-x)dx \int_0^t v(y)g(t-y)dy$$

$$= \int_0^t \int_0^t g(t-x)g(t-y) \text{ av } [v(x)v(y)]dxdy$$

$$= \int_0^t \int_0^t R_I(x-y)g(t-x)g(t-y)dxdy$$

where $R_I(\tau)$ is the autocovariance function of the I ensemble. If the current generator is white thermal noise,

$$R_I(\tau) = 2GkT\delta(\tau)$$

and

$$\text{av } V^2(t) = 2GkT \int_0^t g^2(t)dt.$$

In the circuit of Fig. 3,

$$g(t) = \frac{\exp(-\alpha t)}{\omega_1 C}(\omega_1 \cos \omega_1 t - \alpha \sin \omega_1 t),$$

$$\omega_1^2 = \frac{1}{LC} - \alpha^2, \qquad \alpha = \frac{G}{2C},$$

$$\text{av } V^2(t) = \frac{kT}{C}\left[1 - e^{-2\alpha t}\left(1 + \frac{\alpha^2}{\omega_1^2} - \frac{\alpha}{\omega_1}\sin 2\omega_1 t - \frac{\alpha^2}{\omega_1^2}\cos 2\omega_1 t\right)\right].$$

Setting $t = n\pi/\omega_1$, where n is an integer, gives

$$\text{av } V^2(n\pi/\omega_1) = \frac{kT}{C}(1 - e^{-2n\pi\alpha/\omega_1})$$

$$= \frac{kT}{C}(1 - \beta^n) \text{ where } \beta = e^{-\pi G/\omega_1 C}.$$

This checks (10), except that the text sets $\omega_1 = \omega_0$, which is valid when $\alpha^2 \ll 1/LC$ but represents an unnecessary restriction.

[5] The author gratefully acknowledges this rigorous analysis provided by an anonymous reviewer.

Acknowledgment

The author is indebted to Dr. W. W. Harman for reading the manuscript and contributing to the evaluation of the spectrum. Thanks are also due to Dr. L. C. Hale for suggesting the application of the noise impulse method to this problem. Acknowledgment is also made to W. R. Bennett and to S. O. Rice for their publication of very helpful papers on the subject.

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Background Noise in Nonlinear Oscillators*

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Summary—For a general black box model of a resistive one terminal-pair nonlinear oscillator, it is shown that the noise output from noise bands around the oscillatory frequency is composed of an additive noise of the shape of the oscillator resonant circuit and a very small FM broadening of the oscillator line. A noise generator across the oscillator terminals will produce the same amount of additive noise in the load as if the nonlinearity were absent, but the upper and lower sidebands are fully correlated, not independent as in the linear case. The noise bandwidth of the additive noise can be determined by the passive elements and the change in oscillator output power with changing load conductance at the operating load conductance. When the load conductance is adjusted for maximum oscillator output power, the noise bandwidth is the same as if the value of load conductance gave maximum power from a linear circuit.

I. Introduction

Problems connected with noise in nonlinear networks contain all the usual difficulties associated with nonlinear circuits, so that not nearly so much is known as for the corresponding problems involving linear circuits.

One of the new features introduced by nonlinearity is the possibility of frequency changing, so that one cannot expect, in general, that noise components of the output have arisen from input components of the same frequency. However, in this paper, effects obviously of this type, *e.g.*, low-frequency noise which modulates the oscillator output, are not considered. The problem to be treated is one which is as close to the typical linear problem as possible, namely the effects on self-excited oscillators of broad-band noise at or near the oscillator frequency, such as shot or thermal noise.

The case of lumped constant oscillators with van der Pol nonlinearity has been solved, with many interesting applications, in an intuitive fashion by Blaquiere [1, 2]. Rytov [3], extending Berstein's [4] work, has exploited van der Pol's technique of expressing the oscillator output as an angle and amplitude modulated sine wave [3, 4], combined with Boguliubov's technique of introducing "slow time" as the device for obtaining the perturbation expansion. Rytov's method is quite general, but he has not obtained results in terms of readily observable quantities.

Edson has used the model of a stable feedback amplifier as a guide to the oscillatory case. An alternative method has been proposed which separates the output noise into linear and nonlinear parts but has been unable to calculate the correlation between them [7]. van der Ziel has reported on an analysis which represents the noise voltage as a sine wave [6]. Earlier analyses are then applicable; however, a single sine wave is not an appropriate model because, in the noise case, the continually fluctuating amplitude and phase give effects which are, of necessity, neglected entirely in the sine wave picture. In fact, the noise bandwidth is twice as large as is predicted by the single sine wave model.

Results of reasonable generality and utility appear to be most readily and reliably obtained by the use of Rytov's method; here, however, "slow time" is not used. This paper extends the earlier work by including general forms of resistive nonlinearity in such a way that all the quantities which appear are measurable on a "black box" oscillator, at the desired operating point, and by obtaining the spectrum of the instantaneous output of the oscillator in the general case.

* Original manuscript received by the IRE, June 11, 1959; revised manuscript received, February 24, 1960.
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II. Derivation of Equations

In the neighborhood of the main oscillating frequency the oscillator can be represented as a linear resonant circuit shunted by a nonlinear element and a load conductance. In addition, the resonant circuit is shunted by a noise current generator. This representation is sufficiently detailed to obtain useful results on noise bandwidth; for a determination of the noise level of the output, the detailed nature of the internal circuitry of the oscillator must be taken into account. The representation of the actual noise generators in terms of a single one connected directly across the output appears too simple to be true in general; furthermore, even if it be true, a calculation of the strength of the equivalent generator in terms of the actual generators and circuitry is a distinctly nontrivial problem. Nevertheless, a single generator will suffice for several points of interest, and is certainly appropriate for a black box theory.

The equation for the current between the nodes is

$$C\frac{dV}{dt} + (G + G_L)V + \frac{1}{L}\int V dt + \phi(V) = i_n(t), \quad (1)$$

where $\phi(V)$ is the current through the nonlinear element. This can be expressed as

$$\frac{d^2V}{dt^2} + \frac{\omega_0}{Q}\frac{dV}{dt} + \omega_0^2 V + \frac{1}{C}\frac{d\phi}{dt} = \frac{1}{C}\frac{di_n}{dt}. \quad (2)$$

The essential idea of the method of analysis is that the solution for V should be an oscillatory function with a slowly varying amplitude and phase.

If we then substitute $V = R\cos(\omega_0 t - \theta)$ in (2), we get

$$
\begin{aligned}
&-2\omega_0 R' \sin(\omega_0 t - \theta) + 2\omega_0 R\theta' \cos(\omega_0 t - \theta) \\
&- \omega_0 \frac{\omega_0}{Q} R \sin(\omega_0 t - \theta) + \frac{1}{C}\phi'(R\cos(\omega_0 t - \theta)) \\
&+ \Big\{(2R'\theta' + R\theta'')\sin(\omega_0 t - \theta) \\
&+ (R'' - R\theta'^2)\cos(\omega_0 t - \theta) \\
&+ \frac{\omega_0}{Q}(R'\cos(\omega_0 t - \theta) + R\theta'\sin(\omega_0 t - \theta))\Big\} = \frac{1}{C} i_n(t), \quad (3)
\end{aligned}
$$

where the primes stand for differentiation with respect to time and the expression in braces is of order ω_0^{-1} compared to the remainder of the left-hand side.

This substitution is only approximately correct, since in general it must be expected that harmonic terms will occur. However, the harmonic parts are at least an order smaller than the main oscillation, so that back-coupling to the fundamental frequency does not appear in the equation giving the major oscillation, (7) and (8) below. In the next approximation, (13) and (14), harmonic components may appear as extra forcing terms which give a correction to the steady oscillatory terms but do not affect the noise in any way. Since the object of this paper is to discuss the noise properties of the system for which the harmonics are irrelevant, the harmonic terms have been assumed to be absent.

The presence of the rapidly oscillating terms can be eliminated by multiplying (3) by $\sin(\omega_0 t - \theta)$ and integrating over one period of $\omega_0 t$, with similar use of $\cos(\omega_0 t - \theta)$. During this time, R and θ and their derivatives will be nearly constant so that the method of stationary phase can be applied [8]. The resulting equations are

$$
\begin{aligned}
&R' + \frac{\omega_0}{2Q} R - \frac{1}{\omega_0 C}\langle\phi'(R\cos(\omega_0 t - \theta))\sin(\omega_0 t - \theta)\rangle \\
&\quad - \frac{1}{\omega_0}\left\{R'\theta' + \frac{1}{2}R\theta'' + \frac{\omega_0}{2Q}R\theta'\right\} = \frac{n_s}{\omega_0 C} \quad (4)
\end{aligned}
$$

$$
\begin{aligned}
&R\theta' + \frac{1}{\omega_0 C}\langle\phi'(R\cos(\omega_0 t - \theta))\cos(\omega_0 t - \theta)\rangle \\
&\quad + \frac{1}{\omega_0}\left\{\frac{R'' - R\theta'^2}{2} + \frac{\omega_0}{2Q}R'\right\} = \frac{n_c}{\omega_0 C} \quad (5)
\end{aligned}
$$

where, for convenience we use

$$n_s = \langle i_n'(t)\sin(\omega_0 t - \theta)\rangle$$

$$n_c = \langle i_n'(t)\cos(\omega_0 t - \theta)\rangle.$$

The next step is to represent the envelope and phase as a series in powers of ω_0^{-1}.

$$
\begin{aligned}
R(t) &= R_0(t) + \frac{1}{\omega_0}R_1(t) + \cdots \\
\theta(t) &= \theta_0(t) + \frac{1}{\omega_0}\theta_1(t) + \cdots. \quad (6)
\end{aligned}
$$

The nonlinear term is certainly of first order since the oscillator will not operate without it. The noise, on the other hand, should be taken as a second order quantity, since it is known experimentally to be small compared to the steady oscillation.

Then the first order equations are

$$
\begin{aligned}
&R_0' + \frac{\omega_0}{2Q}R_0 - \frac{1}{\omega_0 C} \\
&\quad \cdot\langle\phi'(R\cos(\omega_0 t - \theta_1))\sin(\omega_0 t - \theta_0)\rangle\big|_{R_0} = 0 \quad (7)
\end{aligned}
$$

$$
\begin{aligned}
&R_0\theta_0' + \frac{1}{\omega_0 C} \\
&\quad \cdot\langle\phi'(R\cos(\omega_0 t - \theta_1))\cos(\omega_0 t - \theta_0)\rangle\big|_{R_0} = 0. \quad (8)
\end{aligned}
$$

The amplitude of the steady oscillation should be constant, so that (7) with $R_0' = 0$, is an implicit equation for the amplitude of the steady oscillation.

Eq. (8) gives the first order correction to the frequency of oscillation, which in this paper will be required to be zero. The correction represents a change in frequency from the "cold" resonant circuit, *i.e.*, with the nonlinear element removed, to the "hot" oscillating frequency. Any thorough theory of noise in oscillators must include this shift, since it cannot be neglected in practical microwave oscillators. However, its inclusion complicates and deepens the analytical apparatus required sufficiently that inclusion of the frequency shift is best postponed to a later date. Accordingly, throughout the remainder of this paper, the second term of (8) will be set equal to zero.

As an example, if the nonlinearity is of the van der Pol type, *i.e.*, purely resistive and containing a cubic term,

$$\phi(V) = C(-\alpha V + \gamma V^3) \quad (9)$$

with α and γ both positive, then

$$
\begin{aligned}
&\langle\phi'(R\cos(\omega_0 t - \theta))\sin(\omega_0 t - \theta)\rangle \\
&\quad = \frac{C\alpha}{2}(\omega_0 - \theta')R\left(1 - \frac{3\gamma}{4\alpha}R^2\right), \quad (10)
\end{aligned}
$$

$$
\begin{aligned}
&\langle\phi'(R\cos(\omega_0 t - \theta))\cos(\omega_0 t - \theta)\rangle \\
&\quad = -\frac{C\alpha}{2}R'\left(1 - \frac{9\gamma}{4\alpha}R^2\right). \quad (11)
\end{aligned}
$$

For the van der Pol nonlinearity, the first order solutions are

$$
\begin{aligned}
\theta_0' &= 0, \\
R_0^2 &= \frac{4}{3\gamma}\left(\alpha - \frac{\omega_0}{Q}\right). \quad (12)
\end{aligned}
$$

III. The Second Order Equations

The second order equations are in general (for $\theta_0' = 0$) [9]

$$R_1' + \frac{\omega_0}{2Q} R_1 - \frac{1}{\omega_0 C} \frac{\partial}{\partial R} \cdot \langle \phi(R \cos(\omega_0 t - \theta)) \sin(\omega_0 t - \theta) \rangle \big|_{R_0} R_1 = \frac{n_s}{\omega_0 C}, \quad (13)$$

$$R_0 \theta_1' = \frac{n_c}{\omega_0 C}. \quad (14)$$

Simplification of (13) proceeds as follows. The output power in the load is $P_{out} = \frac{1}{2} R_0^2 G_L$, so that

$$\frac{\partial P_{out}}{\partial G_L} = \frac{1}{2} R_0^2 + R_0 G_L \frac{\partial R_0}{\partial G_L}. \quad (15)$$

Now, from (7), with $R_0' = 0$, $\omega_0/Q_L = G_L/C$, and $\omega_0/Q = (G_L + G)/C$,

$$\frac{1}{2C} R_0 + \frac{G_L + G}{2C} \frac{\partial R_0}{\partial G_L} - \frac{1}{\omega_0 C} \frac{\partial}{\partial R} \cdot \langle \phi'(R \cos(\omega_0 t - \theta)) \sin(\omega_0 t - \theta) \rangle \Big|_{R_0} \frac{\partial R_0}{\partial G_L} = 0, \quad (16)$$

$$\frac{\omega_0}{2Q} - \frac{1}{\omega_0 C} \frac{\partial}{\partial R} \langle \phi'(R \cos(\omega_0 t - \theta) \sin(\omega_0 t - \theta)) \rangle \big|_{R_0} = -\frac{R_0}{2C} \frac{1}{\dfrac{\partial R_0}{\partial G_L}} = \frac{\dfrac{\omega_0}{Q_L}}{1 - \dfrac{G_L}{P_{out}} \dfrac{\partial P_{out}}{\partial G_L}} \equiv \frac{\omega_0}{Q'}. \quad (17)$$

The equations are then

$$R_1' + \frac{\omega_0}{Q'} R_1 = \frac{n_s}{\omega_0 C}. \quad (18)$$

$$R_0 \theta_{in}' = \frac{n_c}{\omega_0 C}. \quad (19)$$

IV. The Oscillator Output Spectrum

The spectra of the right-hand sides are evaluated in Appendix I, and found to be W_0/C^2, where W_0 is the spectral density of the noise current.

The spectra of envelope and phase components are then found by Fourier transform techniques, whose applicability to random processes have been previously established [10]. The results are

$$W_{R_1}(f) = \frac{W_0}{C^2 \left[\omega^2 + \dfrac{\omega_0^2}{Q'^2} \right]}. \quad (20)$$

$$W_{\theta_1}(f) = \frac{W_0}{R_0^2 C^2} \frac{1}{\omega^2}. \quad (21)$$

Now the correlation function of the over-all wave, in terms of the correlation functions of envelope and phase (for gaussian noise), is [11]

$$R_V(t) = \tfrac{1}{2} \exp\{-R_\theta(0) + R_\theta(t)\} R_R(t) \cos \omega_0 t. \quad (22)$$

Since the envelope and phase arise from independent linear operations on independent gaussian processes, they are independent. The correlation function of the envelope is

$$R_R(t) = R_0^2(1 + n e^{-\omega_0 t/Q'}), \quad (23)$$

and of the phase

$$R_\theta(0) - R_\theta(t) = n \frac{\omega_0 t}{Q'}, \quad (24)$$

where $n = W_0 Q'/4\omega_0 C^2 R_0^2$.

Eq. (24) gives in (22) a shape of the same form as the envelope noise. However, the width of the phase noise is the envelope width reduced by the same quantity (to first order) as the ratio of coefficients of terms in (23), which we can interpret as the noise-to-carrier ratio.

As estimates of the order of these quantities, let us suppose the carrier-to-noise ratio to be one million, which is neither a high nor low value for oscillators. Then, since Q' is of the same size as a passive Q, for a 3-kmc oscillator the envelope noise should be a few megacycles wide and the phase noise a few cycles wide [12]. In superheterodyne systems with incoherent detection the phase noise can usually be neglected since the system usually is insensitive to phase. This is the type of system for which the theory of noise in linear networks was primarily constructed.

However, in coherent detection systems or FM systems, the phase noise must be considered.

If the phase noise can be neglected, the oscillator case bears a close resemblance to the amplifier one; in fact, the spectrum of the oscillator output is

$$W(f)\big|_{\text{AM only}} = R_0^2 \left[\frac{1}{2} \delta(f - f_0) + \frac{n \dfrac{\omega_0}{Q'}}{(\omega - \omega_0)^2 + \dfrac{\omega_0^2}{Q'^2}} \right]. \quad (25a)$$

The complete spectrum is

$$W(f) = R_0^2 \left[\frac{n \dfrac{\omega_0}{Q'}}{n^2 \dfrac{\omega_0^2}{Q'^2} + (\omega - \omega_0)^2} + \frac{n(1+n) \dfrac{\omega_0}{Q'}}{(1+n)^2 \dfrac{\omega_0^2}{Q'^2} + (\omega - \omega_0)^2} \right]$$

$$\doteq R_0^2\left[\frac{n\dfrac{\omega_0}{Q'}}{n^2\dfrac{\omega_0^2}{Q'^2}+(\omega-\omega_0)^2}+\frac{n\dfrac{\omega_0}{Q'}}{\dfrac{\omega_0^2}{Q'_2}+(\omega-\omega_0)^2}\right]. \quad (25b)$$

Since the noise is so much the weaker, its spreading effect can be neglected in the second term. On the other hand, the spreading of what was the carrier line by the FM noise cannot be neglected, since its contribution is of the same size as the AM term. Although the bandwidth of the first term, as measured by the half-power points, is extremely small, the initial amplitude is so large (as befits an approximate δ-function) that the tails of its spectrum are of the same size as the second term.

V. Discussion of the Output Spectrum

The output spectrum, as observed by an RF spectrum analyzer, will appear to have a single peak rising to a maximum at the center frequency [13]. Even without any frequency pushing to cause an additional FM output, the background noise spreads the carrier line enough to obscure the shape of the AM noise. At the half-power point of the AM noise spectrum, the FM noise spectrum is twice as large; and closer to the center frequency the FM noise is even larger. On the tail of the spectrum (more than an octave past the half-power point of the AM noise), the contributions of AM noise and FM noise are equal.

The effect of the noise output will differ depending upon the type of system the oscillator finds use in. The closest analog to amplifier behavior ocurs in an AM system. If the oscillator output is fed to an AM receiver, which does not respond to FM, the receiver output will have a dc output corresponding to the carrier line plus the AM noise shifted down to base band. Thus the additive behavior of signal plus noise in amplifiers is preserved.

The level of AM noise in the oscillator is the same as the noise level in the corresponding amplifier, given the same noise generator in the two cases. However, the strength of the noise generator will ordinarily depend on the nonlinearity and, therefore, will not be the same as in the linear case, nor can it be neatly expressed in general. Also, the character of the noise is different.

In the amplifier case, each small frequency band contributes to the output spectrum independently, so that the upper and lower sidebands are independent of each other. The instantaneous value of the noise sidebands has a gaussian probability density, which implies that the phase of the noise then has a uniform distribution, and the envelope of signal plus noise, a generalized Rayleigh distribution [12]. In the oscillator case, the upper and lower sidebands are the result of amplitude and frequency modulation of the steady oscillation. The noise sidebands have each a gaussian probability density, but are now statistically dependent in such a way that the phase density of the carrier and noise is either a δ-function (neglecting the small FM broadening) or a gaussian distribution, and the envelope has again a gaussian distribution (with nonzero mean value).

This result is made the more plausible if one thinks of the oscillator with sidebands as a parametric amplifier in which the main oscillation plays the role of a pump signal. For instance, for a van der Pol nonlinearity, the apparent conductance in the circuit, *i.e.*, the coefficient of V in the circuit equation, varies as $|G|(1+2\cos 2\omega_0 t)$. This change of the damping term at twice the frequency of oscillation appears to be characteristic of near-sinusoidal oscillators. This variation in conductance will look like a pump signal in a linear varying parameter circuit to the sidebands. If we are to get steady state output at a "signal frequency," there must also be an output at the "idling frequency." The signal and idler frequency are such that their sum is the pump frequency, that is, the signal and idler are upper and lower sidebands of the main oscillating frequency.

The noise generator can be considered as having equal power in phase and in quadrature with a sinusoidal signal. Eqs. (4) and (5) show that one phase acts to produce AM and the other acts to produce FM. Since only the AM adds to the output power, half the strength of the current generator is wasted. On the other hand, the effective half gives two sidebands so the end result is the production of just as much noise as in the linear case. The FM contribution to the power, far from the main oscillation, is equal to that of the AM noise, but the FM part represents a redistribution of main oscillation power, not an added source of noise.

In fact, it should be possible to derive the output spectrum by this method. However, the difficulties involved in properly including general forms of nonlinearity and properly showing that the only significant interactions are those between noise components equidistant above and below the main oscillation are formidable. The method of passing to equations for envelope and phase appears to be both simplest and safest.

The output of the oscillator is not completely described by its spectrum, as is the case with an amplifier, precisely because the amplifier output is gaussian, and the oscillator output is not. As measured by an AM receiver, the oscillator would appear to have a carrier line, and additive noise sidebands of width ω_0/Q_L'. As measured by an FM receiver tuned to the main oscillation, the noise output would be white. These various appearances simply mean that here one cannot rely on the spectrum alone, and that a *complete* parallelism between the amplifier and oscillator cases is not to be obtained. The fact that the AM oscillator output is so similar to the corresponding amplifier output certainly leads one to hope that some results from one domain may be carried into the other. In particular, one may hope for the future that the bounds on noise performance recently established for amplifiers can be extended to oscillators [15].

At present an equality between the noise bandwidths at maximum output power in the two cases can be found. In the oscillator case, the half-power half-bandwidth of the AM noise term is

$$\frac{f_0}{Q'} = \frac{f_0}{Q_L\left[1 - \dfrac{G_L}{P_{\text{out}}}\dfrac{\partial P_{\text{out}}}{\partial G_L}\right]}. \tag{26}$$

When the load conductance is adjusted for maximum power, the derivative is zero, and the noise bandwidth is $f_0/Q_{L\text{opt}}$.

In a linear circuit, with the same linear elements, the bandwidth would be

$$\frac{f_0}{2Q} = \frac{G_L + G}{4\pi C}. \tag{27}$$

and, for maximum power $G=G_L$, so that the bandwidth is $f_0/Q_{L\text{opt}}$ again.

VI. Concluding Remarks

By combining the techniques of van der Pol for nonlinear oscillations and Langevin for noise in linear circuits, it has been possible to show, for a rather general representation of a self-excited oscillator which does not shift in frequency from "cold" to "hot," that the AM noise output from noise sources nearly at the resonant frequency has the shape of the oscillator resonant circuit and that the nonlinearity does not alter the magnitude of coupling of a noise generator across the output to the load. Also, the AM noise bandwidth of the oscillator can be determined from measurements made with the nonlinear element disconnected, and from measurements of power output as a function of load conductance around the operating value of conductance, *i.e.*, without direct measurement of the noise. The fact that the necessary measurements can be made around the operating point of the oscillator is a very useful point in practice, because, particularly for microwave oscillators in early stages of development, tubes are both expensive and mysterious. One hesitates to run them under unusual operating conditions for fear that the tube will burn out or become permanently noisy, *e.g.*, by eroding the cathode surface because of using excessive current density. The necessary information should be gained from measurements which use only normal conditions, if possible; and the result of the derivation above makes it so.

It has been the aim of this paper to develop the "black box" properties of noise in nonlinear oscillators as far as possible so that the results will be applicable in fair generality, in particular to microwave oscillators, as well as lumped constant ones [16]. Neglect of the amplitude-frequency dependence of the oscillator limits the scope severely; however, the similarities and differences between noise in oscillators and amplifiers is more evident without it.

A black box point of view is insufficiently detailed for the purpose of obtaining low noise performance. For this, one must be able to know the internal configuration of the oscillator with the physical noise sources in place. The problem is then to decide how much noise at the output will be produced by a noise generator across some internal pair of terminals.

Another form of noise output, which is often quite troublesome in oscillators, comes from noise sources at other frequencies which modulate the oscillator output [20, 21]. This type of noise has not been treated here at all; it is what may be called an essentially nonlinear kind of noise since in linear circuits where no frequency changing occurs, such effects are not possible [22].

The noise sources with which this paper is concerned are precisely those which appear in linear networks, and thus the results of this analysis lend themselves to direct comparison with linear results. For instance, in a one-stage vacuum tube amplifier with a tuned load, the shot noise of the tube will cause a certain amount of noise in the output. If now the output is coupled back to the input to make the tube an oscillator, without altering the frequency of the resonant load, the shot noise will produce the same noise level at resonance in the output (assuming that the tube draws the same average current as an oscillator, otherwise the noise power should increase as the ratio of average currents) and the amount of noise power can then be found knowing the oscillator noise bandwidth. The upper and lower sidebands of the oscillator output are fully correlated, however. The effect of thermal noise in the input circuit resistance cannot be determined directly from the analysis of this paper because this noise source is not directly across the load.

Appendix

One can reason to the evaluation of the averages containing the noise term from physical considerations. The object of multiplying by the cosine or sine of the oscillating frequency and averaging over a period is to transfer the origin of frequency from zero to the oscillating frequency and then find the inphase and quadrature components of the noise. Now for white noise, one expects the inphase and quadrature components to be equal, and equal to the original value.

Also, the power spectrum of the derivative is the (circular) frequency squared times the original power spectrum. Thus the power spectrum of i_n' referred to f_0 as the zero frequency, should be $(\omega-\omega_0)^2 W_0$, and, since ω is always very small compared to ω_0 (high-Q approximation), this is nearly equal to $\omega_0^2 W_0$ over the pass band of the resonant circuit. The inphase and quadrature components should equal this, so that the spectrum of the noise term [see (4), or (29) below] should be W_0/C^2.

A careful evaluation of the noise term must take account of several more or less delicate considerations. θ and i_n are output and input, respectively, and therefore not statistically independent. The derivative of a white noise process is very badly behaved, so that the operations in which it is involved must be justified.

These difficulties can be avoided by using the fact that the first order approximation to θ, which in no way depends on the noise term, is constant, and the higher orders of θ are both small and slowly varying and can therefore be neglected. The part of the phase which is correlated with the input noise can thus be removed before any statistical considerations enter. Furthermore, the noise terms can be carried through as though they belonged to a differentiable random process; and then, in the final answer, a flat spectrum, which belongs to a discontinuous process but is convenient, can be substituted. The point is that one could use a spectrum belonging to a differentiable process, in particular, for instance, the quantum thermal noise spectrum, whose process is infinitely differentiable. However, at the frequencies with which we are concerned, the quantum formula reduces to a flat spectrum, so that we may use a white noise spectrum in the final formula.

The evaluation of the noise term cannot be made on the slowly varying assumption because the derivative of thermal noise is emphatically not slowly varying. However, θ_1 is constant and higher order θ's are both slowly varying and small. Thus

$$\begin{aligned}\frac{n_e}{\omega_0 C} &= \frac{1}{\omega_0 C}\langle i_n' \cos(\omega_0 t - \theta)\rangle \\ &= \frac{1}{\omega_0 C}\langle i_n'[\cos(\omega_0 t - \theta_0)\cos(\theta - \theta_0) \\ &\quad + \sin(\omega_0 t - \theta_0)\sin(\theta - \theta_0)]\rangle \\ &\doteq \frac{1}{\omega_0 C}\langle i_n' \cos(\omega_0 t - \theta_0)\rangle. \qquad (28)\end{aligned}$$

The next term will be third order and therefore can be neglected. Now the correlation function of the noise term can be found.

$$\begin{aligned}R(t) = \frac{1}{\omega_0^2 C^2}\langle\langle\langle i_n'(t_0+\tau)\cos[\omega_0(t_0+\tau) - \theta_0] \\ i_n'(t_0+t+\sigma)\cos[\omega_0(t_0+t+\sigma) - \theta_0]\rangle_\tau\rangle_\sigma\rangle_{t_0}. \qquad (29)\end{aligned}$$

The noise is independent of the trigonometric part, so that the t_0 average can be done separately on noise and cosine term.

$$R(t) = \frac{1}{\omega_0^2 C^2}\langle\langle R_n'(t+\sigma-\tau)\cos\omega_0(t+\sigma-\tau)\rangle_\sigma\rangle_\tau. \qquad (30)$$

The term containing the sine average will give this same value of correlation function. The power spectrum is given by

$$W(f) = 2\int_{-\infty}^{\infty} R(t)e^{-j\omega t}dt \qquad (31)$$

$$\begin{aligned}W(f) = \frac{1}{2\omega_0^2 C^2}\langle\langle\{e^{-i\omega(\sigma-\tau)}W_n'(f+f_0) \\ + e^{i\omega(\sigma-\tau)}W_n'(f-f_0)\}\rangle_\sigma\rangle_\tau, \qquad (32)\end{aligned}$$

whereupon the two averages split, and

$$\langle e^{-j\omega\sigma}\rangle = \frac{\omega_0}{2\pi}\int_{-\pi/\omega_0}^{\pi/\omega_0} e^{-i\omega\sigma}d\sigma = \frac{\sin \pi\dfrac{\omega}{\omega_0}}{\pi\dfrac{\omega}{\omega_0}}. \qquad (33)$$

The spectrum of the derivative of a random process is ω^2 times the spectrum of the process itself; and, if the noise has a flat spectrum of level W_0, the spectrum can be taken outside the braces, so that

$$W(f) = \frac{1}{2\omega_0^2 C^2}\frac{\sin^2 \pi\dfrac{\omega}{\omega_0}}{\left(\pi\dfrac{\omega}{\omega_0}\right)^2}\{(\omega+\omega_0)^2 + (\omega-\omega_0)^2\}W_0. \qquad (34)$$

$$W(f) = \frac{1}{C^2}\left|\frac{\sin\pi\dfrac{\omega}{\omega_0}}{\pi\dfrac{\omega}{\omega_0}}\right|^2 W_0\left(1+\frac{\omega^2}{\omega_0^2}\right). \qquad (35)$$

Now f is measured from the resonant frequency, and therefore is no larger than a few fractional bandwidths, f_0/Q. Thus ω/ω_0 is very small, so that

$$W(f) \doteq \frac{W_0}{C^2}, \qquad (36)$$

where the neglected terms are of higher order.

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Monochromaticity and Noise in a Regenerative Electrical Oscillator*

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***Summary*—The regenerative oscillator is postulated to be equivalent to an RLC circuit in which a negative resistance $-R_H$, placed parallel with R, has a value which is a slowly varying function of the mean-square voltage across it. Expressions are derived for the departure from monochromaticity of the regenerative oscillator so represented, as well as for the intensity and bandwidth of the thermal noise generated in the oscillator by the resistance *R*.**

Introduction

THE PROBLEM of noise in oscillators has received relatively little attention from communication engineers. Indeed, the most pertinent work in this area was published in 1954 and 1955, on the occasion of a theoretical treatment of the maser.[1,2] In these publications the authors, in effect, liken the maser to a regenerative oscillator in which a negative resistance is carefully adjusted so that it nearly cancels the damping action of a positive resistance in an oscillating circuit, so that the Johnson noise generated by that positive resistance is amplified to the point at which the power delivered to the positive resistance by the negative one equals a pre-assigned value. Thus, the output of the oscillator so treated theoretically is seen to be similar to a narrow optical line, and to contain no coherent component.

Since a regenerative oscillator is characterized by its coherent output, it is evident that the existence of this coherent output should be taken into account in any theoretical treatment of this oscillator purporting to discuss quantitatively the departure of the oscillator output from an ideal line frequency. Continuing this line of reasoning, it is also evident that there is a physical interest in determining the effect of this departure from perfect monochromaticity on the measurement of the oscillator frequency. Since the most accurate frequency measurements are made by observing the time elapsed during a given number of zero crossings of the oscillator output voltage, it is again evident that any amplitude modulation of this output voltage due to noise, while causing this output to have a noise energy spectrum of a given spectral width, will not, at least in the first order, cause any error in the observation of these zero crossings. Thus, it is finally evident that any inquiry into the departure from monochromaticity of an oscillator cannot be fully satisfied merely by acquiring knowledge about the spectral distribution of the oscillator output, but should bear instead on the causes of irregularities in these zero crossings, which are the noise components of this output in quadrature with the coherent output.

* Received by the IRE, September 25, 1959; revised manuscript received, February 2, 1960.
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The "Irregularity" of a Regenerative Oscillator Caused by Johnson Noise

Fig. 1 illustrates the prototype of a regenerative oscillator.[3] When the oscillator voltage V is low, the positive resistance R is more than balanced by the negative resistance $-R_B$, which is then smaller than R. The circuit has poles in the right half of the complex frequency plane and any oscillation present will grow in amplitude. Meanwhile, the squared voltage output is passed by an RC filter with a time constant T_1, and actuates a servo which increases R_B in accordance with the relation

$$\frac{1}{R_B} = \frac{1}{R}\left(1 - a\frac{V^2 - V_0^2}{V_0^2}\right). \tag{1}$$

The time delay T_1 is presupposed to be much larger than the oscillating period, so that the servo remains unaware of the individual waves of the oscillator output. The time delay T_1 is also presupposed to be somewhat shorter than an upper bound, which is defined later.

It is clear that in the absence of Johnson noise, the oscillating voltage will build up until V equals V_0, at which point R_B equals R and the oscillations are just maintained at a constant level.

When Johnson noise is generated by R, the end of the rotating voltage vector representing the oscillator output will have erratic excursions with respect to the ideal noise-free rotating vector which represents a pure line frequency. The excursions parallel to the vector itself will cause no change in the zero crossings, but the excursions normal to the vector will displace these zero crossings so as to cause irregularities in the frequency measurements.

In order to study the effect of noise components normal to the coherent voltage vector, the circuit of Fig. 1 is replaced by the twin low-pass circuits of Fig. 2, in which the two voltages V_1 and V_2 represent the two components of the oscillating voltage of the circuit of Fig. 1 which are, respectively, parallel and perpendicular to the voltage vector with the components $\cos 2\pi\nu_B t$ and $\sin 2\pi\nu_B t$.

The replacement of the band-pass circuit of Fig. 1 by twin low-pass circuits of Fig. 2 is based on the so-called narrow-band approximation, which is permitted whenever the admittance of the reactive components of the first circuit may be written with sufficient accuracy,

$$2\pi i\nu C_B + \frac{1}{2\pi i\nu L_B} \cong 4\pi i(\nu - \nu_B)C_B, \tag{2}$$

where

$$(2\pi\nu_B)^2 L_B C_B = 1.$$

[3] A. Blaquiere has attacked this problem by assuming a non-linear regenerative element and has reached conclusions which have many points of agreement with the conclusions of this study. His two most important publications are: "Effect of background noise on frequency of tube oscillators. Ultimate accuracy of electronics clocks," *Ann. Radioelec.*, vol. 8, pp. 36–80, January, 1953; and "Power spectrum of non-linear oscillator perturbed by noise," *Ann. Radioelec.*, vol. 8, pp. 153–179, April, 1953.

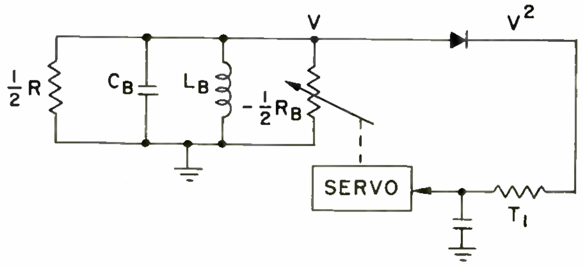

Fig. 1—Oscillating circuit with regenerative servo loop.

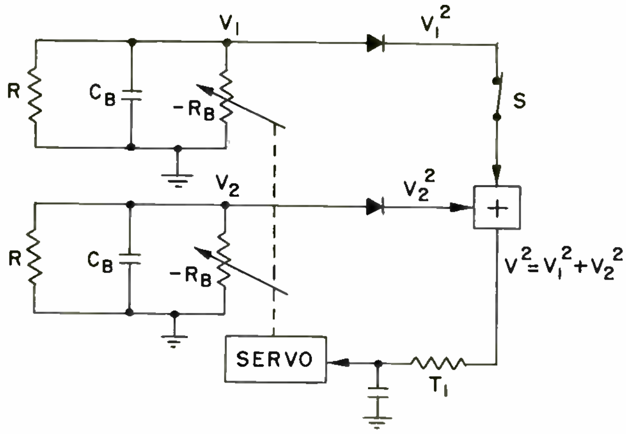

Fig. 2—Twin low-pass equivalents of the regenerative oscillator of Fig. 1.

As in Fig. 1, the squared voltage $V^2 = V_1^2 + V_2^2$, which corresponds to the square of the voltage vector in the circuit of Fig. 1, serves to control the action of the servo which actuates simultaneously the two negative resistances $-R_B$.

Assume first that, instead of being bounded, T_1 is indefinitely large. R_B will eventually adjust itself to the value at which the noise voltages V_1 and V_2 will cause no further change in R_B. The Johnson noise generated by R in both circuits will be equal to

$$dV_1^2 = dV_2^2 = \frac{4kTRd\nu}{\left(1 - \frac{R}{R_B}\right)^2 + (2\pi\nu RC_B)^2}$$

$$= \frac{4kTRd\nu}{\left(a\frac{V^2 - V_0^2}{V_0^2}\right)^2 + (2\pi\nu RC_B)^2}, \tag{3}$$

and it will be assumed, with no loss of generality, that $-R_B$ is noiseless.

Integration of (3) gives

$$V^2 = \int_{\nu=0}^{\nu=\infty} dV_1^2 + dV_2^2 = \frac{4\pi kTR}{a\frac{V^2 - V_0^2}{V_0^2}\cdot 2\pi RC_B} \tag{4}$$

where

$$a\frac{V^2 - V_0^2}{V_0^2} = \frac{2kT}{V^2 C_B} \cong \frac{2kT}{V_0^2 C_B}. \tag{5}$$

Equating the two parts of the denominator of the right-hand side of (3) when $\delta\nu$ is substituted for ν gives the low-pass bandwidth of the noise, which is one-half the band-pass bandwidth of the noise in the circuit of Fig. 1, where T_1 has also been made indefinitely large:

$$\delta\nu = a\frac{\dfrac{V^2 - V_0^2}{V_0^2}}{2\pi RC_B} = \frac{2kT}{2\pi RC_B{}^2V^2} \cong \frac{kT}{\pi}\left(\frac{1}{RC_B}\right)^2\frac{R}{V_0{}^2}$$
$$= 4\pi kT\frac{\Delta\nu_B{}^2}{P_B}, \tag{6}$$

where

$$2\pi RC_B\Delta\nu_B \cong 2\pi R_BC_B\Delta\nu_B = 1, \tag{7}$$

and where $P_B = V_0{}^2/R$ designates the power delivered by $-R_B$ to the system and dissipated in R. The bandwidth given by (6) is seen to be identical with that given by (34) of Gordon, *et al.*[2]

The rigorous treatment of the circuits of Figs. 1 and 2 when T_1 is finite appears to offer considerable mathematical difficulty. Fortunately, the assumption that T_1 is bounded, as will be defined shortly, permits a considerably simplified treatment of Fig. 2. This treatment is based on the device of first opening the switch S, and subsequently correcting the effect of this artificiality.

Opening S causes the lower low-pass circuit to be alone in the servo loop, so that it builds up to a dc voltage $V_2 \cong V_0$, while the upper circuit contains noise only. The lower circuit also contains noise, but since this noise is in phase with the dc voltage corresponding to coherent oscillations, the zero crossings of $\cos 2\pi\nu_Bt$ are not affected by the noise in the lower circuit.

The noise in the upper low-pass circuit constitutes a source of irregularity in the zero crossings of the ac voltage, but since this noise has no dc component when S is open, the deviations of these zero crossings from those of $\cos 2\pi\nu_Bt$ will average zero. This is, of course, a result of S being left open indefinitely.

If, instead, S is open at the time $t=0$, at which time the noise voltage V_1 will be assumed to be zero, and then closed at the time $t=T_1$, *i.e.*, after a time of the order of that required for $V_1{}^2+V_2{}^2$ to pass the low-pass filter of the servo loop and reach the servo, the combination of the $V_2 \cong V_0$ voltage in the lower circuit and of the noise voltage which has built up during the time T_1 in the upper circuit represents a voltage with zero crossings shifted by some phase angle $\Delta\phi_1$ from those of $\cos 2\pi\nu_Bt$. Assume now that the corresponding voltage vector is decomposed in a new system of coordinates which forms the angle $\Delta\phi_1$ with respect to the first. Two low-pass circuits can again be associated with the band-pass circuit in accordance with the decomposition of the voltage vector in the new system of coordinates. The low-pass circuit which carries the full value of this voltage vector may then be, as formerly, the low-pass circuit within the servo loop, while the other low-pass circuit, which carries zero voltage at the instant of the new decomposition, corresponds to the upper low-pass circuit of Fig. 2. Now again, the switch is opened and a voltage generated by R builds up within this low-pass circuit. But since the decomposition of the voltage vector at time $t=T_1$ is equivalent to resetting the low-pass circuit at $V_1=0$, the new V_1 noise voltage will build up independent of the last, and the successive V_1 voltages which build up every time S is opened will cause the end of the coherent voltage vector to execute a random walk, which can be immediately associated with any error made when measuring the oscillator frequency by observing zero crossings.

An expression for the $\Delta\phi$ increments is obtained as follows. Consider first the circuit formed by R and C_B with $-R_B$ disconnected. When R is open-circuited, it generates a mean-square voltage $4kTR$ per unit bandwidth and integration of the squared voltage generated across C_B gives the well-known result,

$$\overline{V_N{}^2} = \frac{kT}{C_B}.$$

If the noise voltage is noted as the time $t=0$ and is allowed to decay for the time t, the noise voltage observed at the end of the t interval will be composed of the voltage decayed from the origin of time plus a voltage built up on the capacitor by R. But statistically we should have

$$\overline{V_N{}^2} = \overline{\left(V_N \exp\left(-\frac{t}{RC_B}\right) + \Delta V_N\right)^2}.$$

Since the two voltages of the right-hand side are independent we have

$$\overline{V_N{}^2} = \overline{V_N{}^2}\exp\left(-\frac{2t}{RC_B}\right) + \overline{\Delta V_N{}^2}.$$

Hence, for the noise voltage ΔV_N built up in the time t,

$$\overline{\Delta V_N{}^2} = \overline{V_N{}^2}\left(1 - \exp\left(-\frac{2t}{RC_B}\right)\right) \cong \frac{2t}{RC_B}\overline{V_N{}^2} \tag{8}$$

where the approximation will be considered tentatively to be valid only when $2t/RC_B$ is small.

When R_B is connected in parallel with R, the mean-square noise voltage per bandwidth $d\nu$ generated across R, taking (1) and (5) into account, is

$$dV^2 = \frac{4kTd\nu}{\dfrac{1}{R} - \dfrac{1}{R_B}} = 2V_0{}^2RC_Bd\nu.$$

The time constant of the R, $-R_B$, C_B circuit, again taking (1) and (5) into account, is

$$\frac{RR_BC_B}{R_B - R} = \frac{RC_B{}^2V_0{}^2}{2kT}. \tag{9}$$

and the noise voltage across C_B at time t after discharge will be

$$\Delta V^2 = \frac{kT}{C_B\left(1 - \frac{R}{R_B}\right)}\left[1 - \exp\left(-\frac{2t(R_B - R)}{C_B R R_B}\right)\right]$$

$$= \frac{1}{2} V_0^2\left[1 - \exp\left(-\frac{4tkT}{V_0^2 C_B^2 R}\right)\right]$$

$$\cong \frac{2kT}{C_B}\frac{t}{C_B R} = 4\pi\Delta\nu_B t\frac{kt}{C_B}. \tag{10}$$

The approximation for the right-hand side of (1) requires that the exponent in the second expression given for ΔV^2 in (10) be somewhat smaller than unity,

$$t < \frac{V_0^2 C_B^2 R}{4kT} = \frac{1}{4\pi\delta\nu}. \tag{11}$$

Eqs. (10) and (11) indicate that as long as the servo loop time constant T_1 is appreciably smaller than $\delta\nu^{-1}$, the successive phase increments of the ac voltage vector which are acquired randomly at intervals T_1 will build up, in time t, the random number of cycles

$$\Delta n = \frac{\Delta V}{2\pi V_0} = \frac{1}{2\pi V_0}\sqrt{\frac{2tkT}{C_B^2 R}}. \tag{12}$$

The frequency error caused by the random rotations of the ac voltage vector will be given by

$$\Delta\nu = \frac{\Delta n}{t} = \frac{1}{2\pi}\sqrt{\frac{2kT}{V_0^2 C_B^2 R t}} = \Delta\nu_B\sqrt{\frac{2kT}{P_B t}}$$

$$= \Delta\nu_B\sqrt{\frac{2kT}{E_B}}, \tag{13}$$

where $E_B = P_B t$ designates the energy contributed by $-R_B$ to the oscillating circuit during the time of measurement t.

Eq. (13) represents, therefore, the departure from monochromaticity of a Johnson noise perturbed oscillator in which the regenerative element, $-\frac{1}{2}R_B$, is assumed noise-free.

While $\Delta\nu$ is the quantity of pertinent interest in an oscillator, the questions of the actual amount and spectral distribution of the oscillator noise remains of some interest, and an attempt at obtaining quantitative expressions for these will be described next.

Spectral Distribution of Noise in a Regenerative Oscillator

The noise in a regenerative oscillator of the type illustrated by Fig. 1 contains two basically different components: the noise components normal to the coherent ac voltage vector, and those parallel to it.

Let ϕ be the instantaneous phase of the ac voltage vector with respect to $\cos 2\pi\nu_B t$. The two instantaneous components of the ac voltage vector are $V_0 \cos\phi$ and $V_0 \sin\phi$, plus the noise components. The noise components parallel to this vector will be assumed to be zero at this instant. The noise perpendicular to the ac voltage vector is obtained by making the Fourier analyses of the random walk components of this vector of components $V_0\cos\phi$ and $V_0\sin\phi$, and a very rough measure of the bandwidth of this noise, $\Delta\nu n$, can be given by $1/2t_0$, where t_0 designates the time interval at the extreme ends of which there is substantial loss of correlation between the two values of $\cos\phi$ or $\sin\phi$. This may be regarded as the time interval during which the ac voltage vector will probably make a quarter turn, and which is obtained by making $\Delta n = \frac{1}{4}$ in (12). This gives for $1/2t_0$,

$$\Delta\nu_n = \frac{1}{2t_0} = 16\Delta\nu_B^2\frac{kT}{P_B}, \tag{14}$$

a figure in close agreement with that obtained in (6) for $\delta\nu$. The squared noise voltage is given simply by

$$\Delta V_n^2 = V_0^2, \tag{15}$$

since this noise is due to the random rotation of a voltage vector of length V_0.

An expression for the noise parallel to the ac vector is much more complicated to obtain, and the calculations made to obtain it will be sketched below. These calculations will be based on the model of Fig. 2 with S open, *i.e.*, the fluctuations of V_2 alone will be considered now since the fluctuations of V_1 are connected with the noise components perpendicular to the coherent voltage vector.

The basic procedure will consist of replacing the low-pass filter of the servo loop by a time delay network whereby the voltage V_2 is sampled at intervals T_1, and the value found at any sampling epoch controls the value of $-R_B$ during the following T_1 interval.

Consider first the noiseless case, and assume that, at the first sampling epoch for which $t=0$, V_2 has the value

$$V_2 = V_0 + \Delta V_0.$$

This value determines the value of $1/R_B$ for the next T_1 interval

$$\frac{1}{R_B} = \frac{1}{R}\left[1 - a\frac{V_2^2 - V_0^2}{V_0^2}\right] \cong \frac{1}{R}\left[1 - 2a\frac{\Delta V_0}{V_0}\right]. \tag{16}$$

Eq. (16) gives for the combined admittance of $1/R$ and $-1/R_B$ the value

$$G = \frac{1}{R} - \frac{1}{R_B} = \frac{2a\Delta V_0}{RV_0}, \tag{17}$$

and we obtain for the time constant T_0 of the lower low-pass circuit

$$T_0 = \frac{G_B}{G} = \frac{V_0}{2a\Delta V}RC_B. \tag{18}$$

After the time interval T_1, which is assumed to be smaller than T_0, V_2 decays from the value $V_0+\Delta V$ to

$$V_2' \cong (V_0 + \Delta V)\left(1 - \frac{T_1}{T_0}\right)$$

$$= (V_0 + \Delta V)\left(1 - \frac{2a\Delta T}{V_0}\frac{T_1}{RC_B}\right)$$

$$\cong V_0 + \Delta V\left(1 - 2a\frac{T_1}{RC_B}\right). \tag{19}$$

It will be noted that everything occurs as if any departure of V_2 from V_0 decays with the time constant

$$T_a = \frac{RC_B}{2a}. \tag{20}$$

Consider now the noisy case. We start again with the voltage V_2 given by (15) and allow the departure of V_2 from V_0 to decay, but meanwhile permitting a noise voltage generated by R to build up for a time T_1 in accordance with the expression given for ΔV^2 in (10):

$$\Delta V_p^2 = \frac{kT}{C_B\left(1 - \frac{R}{R_B}\right)}\left[1 - \exp\left(-\frac{2T_1(R_B - R)}{C_BRR_B}\right)\right]$$

$$\cong \frac{2T_1kT}{C_B^2R}. \tag{21}$$

When added randomly to the decayed voltage

$$\Delta V_p\left(1 - 2a\frac{T_1}{RC_B}\right),$$

the noise voltage given by (21) should reproduce itself, on an average. The random addition of these two voltages is made by adding the squares, and we obtain the relation

$$\Delta V_p^2\left(1 - 2a\frac{T_1}{RC_B}\right)^2 + \frac{2T_1kT}{C_B^2R} = \Delta V_P^2, \tag{22}$$

giving

$$\Delta V_p = \sqrt{\frac{kT}{2aC_B}}. \tag{23}$$

Combining (17) and (23) gives the admittance of the low-pass circuit

$$G = \frac{1}{R}\sqrt{\frac{2akT}{C_BV_0^2}}. \tag{24}$$

The bandwidth of the low-pass circuit is given by

$$\Delta\nu_P = \frac{G}{2\pi C_B} = \Delta\nu_B\sqrt{\frac{2akT}{C_BV_0^2}} = (\Delta\nu_B)^{3/2}\sqrt{\frac{4\pi akT}{P_B}}, \tag{25}$$

and the squared noise voltage is given by

$$\Delta V_p^2 = \frac{kT}{2aC_B}. \tag{26}$$

Conclusion

When the regeneration of an oscillating circuit is controlled by the servo loop of Fig. 1 and by the servo equation of (1), the following are true:

1) When the time constant T_1 of the servo loop is nearly infinite, the voltage output is incoherent, and the half bandwidth of this essentially noisy output is given by

$$\delta\nu = 4\pi kT\frac{\Delta V_B^2}{P_B},$$

where $\Delta\nu_B = 1/2\pi RC_B$ designates the half bandwidth which the oscillating circuit would have if the regenerative resistance $-1/R_B$ were taken out, and where P_B designates the power delivered to the circuit by this regenerative resistance.

2) When the time constant T_1 is appreciably smaller than $\delta\nu^{-1}$, and appreciably greater than the oscillation period, the voltage output is coherent, and it becomes necessary to distinguish between departure from monochromaticity or irregularity of oscillations and bandwidth of the noise in the output.

The departure from monochromaticity is given by the frequency error

$$\Delta\nu = \Delta\nu_B\sqrt{\frac{2kT}{E_B}},$$

where E_B designates the energy delivered to the circuit by the regenerative resistance during the time of measurement.

The bandwidth of the noise normal to the coherent output vector is given by

$$\Delta\nu_n = 16\Delta\nu_B^2\frac{kT}{P_B},$$

and the voltage square of this noise is

$$\Delta V_n^2 = V_0^2.$$

The bandwidth of the noise parallel with the coherent output vector is

$$\Delta\nu_p = (\Delta\nu_B)^{3/2}\sqrt{\frac{4\pi akT}{P_B}}.$$

where a is the servo constant of (1). The voltage square of this noise is

$$\Delta V_p^2 = \frac{kT}{2aC_B}.$$

Correspondence

A Note on Tunnel Emission*

I would like to call attention to several errors in my recent communication.[1] In (1), the quantity between J_0 and the exponential should be squared, and there should be a 2 in the denominator of the expression for E_0. Also in the last equation for the figure of merit M, the E_0 in the numerator should be deleted and the E in the denominator should be changed to E_0.

It should be noted that these equations are very similar to those for ordinary field emission from the surface of a metal into a vacuum. However, the mechanism involved is somewhat different in principle. In the field emission case, electrons are confined by a region of classically forbidden energy, and are permitted to penetrate this region by virtue of their quantum mechanical properties. In contrast, the "forbidden" region considered in tunnel emission (or Zener breakdown) is itself a consequence of the wave nature of the electrons and the periodicity of the lattice. Because the amplitude of the wave function is attenuated similarly in both cases, the voltage-current expression for tunnel emission is essentially identical to the well-known Fowler-Nordheim expression for ordinary field emission when appropriate correction is made for the effective mass of the electron.

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* Received by the IRE, May 23, 1960.

[1] C. A. Mead, "The tunnel emission amplifier," Proc. IRE, vol. 48, p. 359; May, 1960.

Noise Performance of Tunnel-Diode Amplifiers*

The tunnel diode, first reported by Esaki,[1] is being considered for possible use as a low-noise amplifier[2,3] as well as for other applications.[4,5] For comparison of this amplifier with others, it is important to know the fundamental limit on noise performance.

A general way of obtaining the fundamental limit of amplifier noise performance has been given by Haus and Adler,[6] and this was simplified somewhat for negative-resistance amplifiers by this writer.[7] On this basis, we will calculate F, the lowest noise figure an amplifying system of high gain can have if it employs the tunnel diode and no other "better" amplifiers.

The tunnel diode can amplify because its volt-ampere characteristic shows a negative slope at some points. A typical curve is shown in Fig. 1, showing the dc operating point, and the product of the dc current I_0 and the magnitude of the negative resistance R.

Sommers[4] has proposed the small-signal high-frequency equivalent circuit of Fig. 2. The negative resistance is $-R$, so that R is a positive quantity; C is the barrier capacitance; and R_s is the series resistance caused by lead resistance, bulk resistance of the semiconductor, and contact resistance. Fig. 3 shows the equivalent circuit with two noise generators: one to account for shot noise associated with the tunnelling process, represented by the current generator with mean-squared current

$$\overline{|i_n|^2} = 2qI_0\Delta f; \tag{1}$$

and the other to represent thermal noise of the series resistance, with mean-squared voltage

$$\overline{|e_n|^2} = 4kT_dR_s\Delta f \tag{2}$$

where T_d is the diode temperature.

The best noise figure F is given by the formula[6,7]

$$F = 1 - \frac{\text{exchangeable noise power of the tunnel diode}}{kT_0\Delta f}. \tag{3}$$

The exchangeable noise power is calculated from Fig. 3 to be

$$-\frac{\overline{|e_n|^2} + \overline{|i_n|^2}\dfrac{R^2}{1+(\omega RC)^2}}{4\left[\dfrac{R}{1+(\omega RC)^2} - R_s\right]} \tag{4}$$

where ω is the operating frequency; so the fundamental limiting noise figure is

$$F = 1 + \frac{I_0R}{2kT_0/q}\,\frac{R}{R - R_s[1+(\omega RC)^2]} + \frac{T_d}{T_0}\,\frac{R_s}{\dfrac{R}{1+(\omega RC)^2} - R_s}. \tag{5}$$

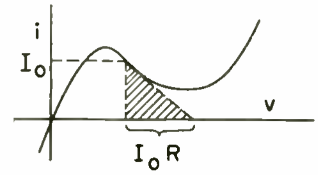

Fig. 1.

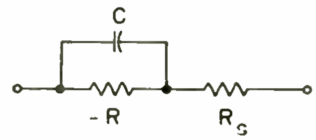

Fig. 2.

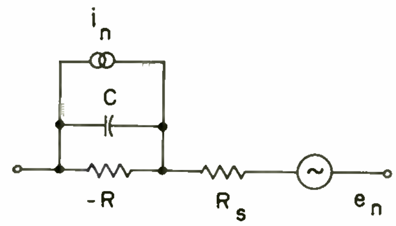

Fig. 3.

An interpretation of this formula is aided by consideration of some special cases.

Zero Series Resistance Limit

In the limit as $R_s \to 0$, F becomes

$$1 + \frac{I_0R}{2kT_0/q} \tag{6}$$

which can be determined from Fig. 1, if the high-frequency value of R is the same as the dc value. The product I_0R is shown in Fig. 1, and F is merely one plus this value normalized to $2kT_0/q = 50$ mv. This model was used by Chang[2] to interpret experimental noise figures, but his numbers are somewhat higher than the limit given here because he 1) was not always operating with high exchangeable gain, 2) had extra loss in the circuit, and 3) included noise from the load.

Low-Frequency Limit

In the low-frequency limit the noise figure does not approach unity, but rather

$$1 + \frac{I_0R}{2kT_0/q}\,\frac{R}{R - R_s} + \frac{T_d}{T_0}\,\frac{R_s}{R - R_s}. \tag{7}$$

Since low-noise amplifiers can be made at low frequencies, either with varactors or vacuum tubes, the tunnel diode will prob-

* Received by the IRE, January 15, 1960. This work was supported in part by the U. S. Navy, under contract NObsr 77621, with Microwave Associates, Inc., Burlington, Mass.

[1] L. Esaki, "New phenomenon in narrow germanium p-n junctions," *Phys. Rev.*, vol. 109, pp. 603–604; January, 1958.

[2] K. K. N. Chang, "Low-noise tunnel-diode amplifier," Proc. IRE, vol. 47, pp. 1268–1269; July, 1959.

[3] H. S. Sommers, Jr., K. K. N. Chang, H. Nelson, R. Steinhoff, and P. Schnitzler, "Tunnel diodes for low noise amplifications," 1959 WESCON Convention Record, pt. 3, pp. 3–8.

[4] H. S. Sommers, Jr., "Tunnel diodes as high-frequency devices," Proc. IRE, vol. 47, pp. 1201–1206; July, 1959.

[5] I. A. Lesk, N. Holonyak, Jr., U. S. Davidsohn, and M. W. Aarons, "Germanium and silicon tunnel diodes—design, operation, and application," 1959 WESCON Convention Record, pt. 3, pp. 9–31.

[6] H. A. Haus and R. B. Adler, "Circuit Theory of Linear Noisy Networks," Technology Press, Mass. Inst. Tech., Cambridge, Mass., and John Wiley and Sons, New York, N. Y.; 1959.

[7] P. Penfield, Jr., "Noise in negative-resistance amplifiers," IRE Trans. on Circuit Theory, vol. CT-7, pp. 166–170; June, 1960.

ably not be competitive as a low-frequency amplifier, except on its other merits. We are led, then, to look at the behavior at high frequencies, *i.e.*, in the microwave region.

High-Frequency Limit

The parasitic series resistance R_s dictates the high-frequency limit. By analogy to the varactor, we define the "cutoff frequency"

$$\omega_c = 2\pi f_c = \frac{1}{R_s C}. \tag{8}$$

The tunnel diode can amplify only if it presents a negative-real impedance, so for gain

$$R_s < \frac{R}{1 + (\omega RC)^2}. \tag{9}$$

Since

$$\frac{R}{1 + (\omega RC)^2} \leq \frac{1}{2\omega C}, \tag{10}$$

there is gain only at frequencies less than one-half the cutoff frequency. At high frequencies, a simple lower bound for the noise figure is

$$F \geq 1 + \frac{I_0 R}{2kT_0/q} \frac{\omega_c}{\omega_c - 2\omega} + \frac{T_d}{T_0} \frac{2\omega}{\omega_c - 2\omega}. \tag{11}$$

Thus, F gets large for frequencies an appreciable fraction of the cutoff frequency.

Conclusions

The fundamental noise-performance limit of tunnel diode amplifiers has been derived under the assumption that the only significant noise sources are: 1) shot noise of the dc tunnelling current, and 2) thermal noise of the series resistance. It is tempting to compare the noise-performance limit with that of a varactor parametric amplifier,[8] since each uses a semiconductor diode and since the series resistance of each is important in setting the noise limits.

Although such a comparison is not of rigorous lasting value, since there is no fundamental way to compare the two cutoff frequencies, we can still compare them roughly, with some assurance that technological advances which prove useful in raising one cutoff frequency will probably also be useful in raising the other. Both cutoff frequencies are determined by the product of a series resistance and a capacitance; for the tunnel diode the resistance can probably be lower, but the capacitance is much higher than the minimum varactor capacitance. The design of the tunnel diode must be a compromise between achieving usable negative resistance and obtaining high cutoff frequency, whereas the design of a varactor can be directed solely to high cutoff frequency. In view of this, it seems reasonable to compare a given tunnel diode with a varactor whose cutoff frequency is three or more times that of the tunnel diode. Under this comparison the varactor is capable of a lower noise figure[8] than the last term in (11) alone.

Of course, low noise is not the only desirable quality an amplifier can have. No high-frequency pump is required for the tunnel diode amplifier—a point in its favor. And, there are many applications for tunnel diodes other than as linear amplifiers. It is possible that tunnel-diode oscillators will be useful as pumps for varactor amplifiers.

The author is indebted to Dr. A. Uhlir, Jr., for discussions of some of the points presented here.

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[8] H. A. Haus and P. Penfield, Jr., "Noise Performance of Parametric Amplifiers," Energy Conversion Group, Elec. Engrg. Dept., Mass. Inst. Tech., Cambridge, Mass., unpublished internal memorandum No. 19, August 11, 1959; submitted to the IRE.

High-Frequency Radar Echoes from the Sun*

In the presence of a magnetic field, a radar wave incident on the sun is split into the ordinary and extraordinary magneto-ionic components. These have different reflection levels. That for the ordinary wave[1] is at $X=1$, which is the same as the field-free reflecting level. The extraordinary component is reflected at the level $X=1-|Y|$, which is ***higher*** than the field-free level. The extraordinary wave has a shorter path, and may suffer less absorption, than the ordinary wave.

Kerr,[2] Bass and Braude,[3] and Hatanaka[4] have considered the field-free case, and have concluded that there will be no echo at high frequencies (hundreds of megacycles and higher) because the wave is all absorbed before it reaches the level $X=1$. Hatanaka,[4] for example, estimated that at 400 mc the optical depth (τ) at the reflection level would be about 10^3. In Hatanaka's example, the reflection level is in the chromosphere, where the temperature is about 10^{4}°K. Because of the low temperature, most of the absorption takes place close to the level $X=1$. If the magnetic field is strong enough, however, the reflecting level for the extraordinary wave will be lifted into the corona where the temperature is 10^6°K. The absorption is proportional to $T^{-3/2}$, so the optical depth at the extraordinary reflecting level is reduced by a factor of about 10^3, giving $\tau_r \sim 1$.

We thus have the possibility that high-frequency echoes will be obtained from regions where strong magnetic fields exist in the corona. The reflection will occur where $|Y| \approx 1$, since X will be small. A magnetic field of about 140 gauss is required for reflection of a 400 mc wave. The direction of the field is immaterial.

Coronal magnetic fields with strengths on the order of 100 gauss have been suggested on several different grounds. Westfold[5] used the field to explain shock waves, Warwick,[6] to explain active prominences, and Akabane[7] and Gelfreich *et al.*,[8] to explain radio eclipse observations. A number of people also have suggested that the polarized noise storms at meter wavelengths are to be explained by fields of about 100 gauss in large clouds very high in the corona (up to one solar radius.)

We can make a rough estimate for the strength of the echo from a high magnetized cloud, the seat of a noise storm. Assume that the cloud is 6 minutes of arc, or 3×10^5 km, in diameter;[9] then the echo is about 3 db greater than that from Venus when it is closest, assuming that the two bodies have similar reflectivities. The background noise will, however, be much higher for the solar cloud, because the solar thermal radiation will contribute 10^5–10^6°K to the antenna temperature. A radar would have to have a sensitivity 20 to 30 db greater than that of the Millstone radar,[10] to detect the cloud marginally.

The space radar suggested by Gordon[11] would, at 400 mc, have a sensitivity about 40 db above Millstone, so echoes from the cloud could easily be seen by it, provided the magnetic field were strong enough. The beamwidth of Gordon's radar would be about 10 minutes of arc, so that if a coronal object were on the limb, and high enough (5×10^5 km or more), it could be seen without the main lobe's being on the hot disk. This would increase the sensitivity by about 20 db, so that quite small, high, limb objects could be seen if they had sufficient field.

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* Received by the IRE, April 11, 1960. This work was partially supported by the Air Force Cambridge Research Center.

[1] J. A. Ratcliffe, "The Magneto-Ionic Theory and its Application to the Ionosphere," Cambridge University Press, Cambridge, Eng., ch. 6; 1959.

[2] F. J. Kerr, "On the possibility of obtaining radar echoes from the sun and planets," Proc. IRE, vol. 40, pp. 660–666; June, 1952.

[3] F. G. Bass and S. Ia. Braude, "On the question of reflecting radar signals from the sun," *Ukr. J. Phys.*, vol. 2, pp. 149–163; 1957.

[4] T. Hatanaka, "Reflection of Radio Waves from the Sun," School of Elec. Engrg., Cornell University, Ithaca, N. Y., Res. Rept. EE450; October, 1959.

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[9] J. A. Högbom, "The instantaneous position and diameter of short duration bursts of solar radio emission," *ibid.*, pp. 251–252.

[10] R. Price, P. E. Green, Jr., T. J. Goblick, Jr., R. H. Kingston, L. G. Kraft, Jr., G. H. Pettengill, R. Silver, and W. B. Smith, "Radar echoes from Venus," *Sci.*, vol. 129, pp. 751–753; March, 1959.

[11] W. E. Gordon, "Incoherent scatter of radio waves by free electrons with applications to space exploration by radar," Proc. IRE, vol. 46, pp. 1824–1829; November, 1958.

WWV and WWVH Standard Frequency and Time Transmissions*

The frequencies of the National Bureau of Standards radio stations WWV and WWVH are kept in agreement with respect to each other and have been maintained as constant as possible with respect to an improved United States Frequency Standard (USFS) since December 1, 1957.

The nominal broadcast frequencies should, for the purpose of highly accurate scientific measurements, or of establishing high uniformity among frequencies, or of removing unavoidable variations in the broadcast frequencies, be corrected to the value of the USFS, as indicated in the table below.

WWV FREQUENCY
WITH RESPECT TO U. S. FREQUENCY STANDARD

1960 May 1600 UT	Parts in 10^{10}*
1	−146
2	−146
3	−146
4	−146
5	−145
6	−145
7	−145
8	−144
9	−144
10	−144
11	−144
12	−144
13	−144
14	−144
15	−145
16	−145
17	−145
18	−145
19	−145
20	−145
21	−145
22	−145
23	−145
24	−144
25	−144
26	−144
27	−144
28	−144
29	−144
30	−144
31	−144

* A minus sign indicates that the broadcast frequency was low.

The characteristics of the USFS, and its relation to time scales such as ET and UT2, have been described in a previous issue,[1] to which the reader is referred for a complete discussion.

The WWV and WWVH time signals are also kept in agreement with each other. Also, they are locked to the nominal frequency of the transmissions and consequently may depart continuously from UT2. Corrections are determined and published by the U. S. Naval Observatory. The broadcast signals are maintained in close agreement with UT2 by properly offsetting the broadcast frequency from the USFS at the beginning of each year when necessary. This new system was commenced on January 1, 1960. The last time adjustment was a retardation adjustment of 0.02 s on December 16, 1959.

NATIONAL BUREAU OF STANDARDS
Boulder, Colorado

* Received by the IRE, June 27, 1960.

[1] "United States National Standards of Time and Frequency," PROC. IRE, vol. 48, pp. 105–106; January, 1960.

The Significance of Transients and Steady-State Behavior in Nonlinear Systems*

I believe Dr. Wolf[1] is in error in his use of the terms "steady state" and "transient" as applied to linear systems.

It should be noted that since $g_1(x)=0$ for all $x<0$, then Wolf's "transient" term

$$\int_t^\infty k(\tau)g_1(t-\tau)d\ \ = 0, \qquad (1)$$

since $g_1(t-\tau)=0$ for $\tau>t$. This last follows from the limits of the integral. Hence we have

$$g_2(t) = \int_0^t k(\tau)g_1(t-\tau)d\tau$$
$$= \int_0^\infty k(\tau)g_1(t-\tau)d\tau, \qquad (2)$$

and this is usually called the transient response of the system K.

The terms "steady-state" response and "transient distortion," which I believe Wolf has in mind, hold only when $g_1(t)$ is a suddenly applied sine wave. Paraphrasing Carson,[2] let

$$g_1(t) = \epsilon^{j\omega t} \qquad 0 < t. \qquad (3)$$

Then

$$g_2(t) = \int_0^t k(\tau)\epsilon^{j\omega(t-\tau)}d\tau$$
$$= \int_0^\infty k(\tau)\epsilon^{j\omega(t-\tau)}d\tau$$
$$- \int_t^\infty k(\tau)\epsilon^{j\omega(t-\tau)}d\tau$$
$$= \epsilon^{j\omega t}K(j\omega) - \int_t^\infty k(\tau)\epsilon^{j\omega(t-\tau)}d\tau. \qquad (4)$$

The first term on the right is called by Carson the "final steady state" and the second the "transient distortion, which ultimately dies away for sufficiently large values of time."

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Author's Comment[3]

The author would like to thank Mr. Doba for his comments on the note[4] concerning transients in nonlinear systems.

Let us carefully examine (1) of Doba's comments and the conclusion which followed, namely (2). In (1) Doba shows that under the condition for $t<\tau$ the kernel, namely $g(t-\tau)$, vanishes. Hence, he rightfully concludes that

$$g_{2T}(t) = \int_t^\infty k(\tau)g_1(t-\tau)d\tau = 0 \qquad (1)$$

for all $t<\tau$.

From this Doba derives (2) and implies in general that the total instantaneous response

$$\int_0^t k(\tau)g_1(t-\tau)d\tau$$

is identically equal to the result obtained by allowing t in the upper limit to approach infinity. If we consider the integrals of the right member of (2) to be Riemann integrals, then what Doba is really saying is that the area under the curve bounded by $k(\tau)g_1(t-\tau)$ and the τ axis is the same no matter whether we integrate from 0 to t or from 0 to ∞. This is obviously false in general for the class of functions considered.[5–7] The only class of functions $g_1(t)$ satisfying (2) for all t is the trivial class belonging to the null set. What Doba neglected to consider was the case for $t>\tau$. In this case his (1) is in general not zero, that is,

$$g_{2T}(t) = \int_t^\infty k(\tau)g_1(t-\tau)d\tau \neq 0 \qquad (2)$$

for all $t>\tau$.

Therefore, I cannot agree in general with Doba's conclusions concerning (4) of my original note given above as (1).

With regard to Doba's remarks concerning the use of my terminology, I should like to point out that these terms have been previously established in the literature.[5–10] In the mathematical literature the corresponding terms are "particular integral" and "complementary solution."[11,12]

In Carson's[13] work referred to in Doba's note, the definitions are a special case of the definitions given in my original note, namely,

Transient Response:

$$g_{2T}(t) = \int_t^\infty k(\tau)g_1(t-\tau)d\tau. \qquad (3)$$

Steady-State Response:

$$g_{2S}(t) = \int_0^\infty k(\tau)g_1(t-\tau)d\tau. \qquad (4)$$

Total Instantaneous Response:

$$g_2(t) = \int_0^t k(\tau)g_1(t-\tau)d\tau. \qquad (5)$$

* Received by the IRE, January 5, 1960.

[1] A. A. Wolf, "The significance of transients and steady-state behavior in nonlinear systems," PROC. IRE, vol. 47, pp. 1785–1786; October 1959.

[2] J. R. Carson, "Electric Circuit Theory and the Operational Calculus," McGraw-Hill Book Co., Inc., New York, N. Y.; 1926.

[3] Received by the IRE, January 27, 1960.

[4] A. A. Wolf, "The significance of transients and steady-state behavior in nonlinear systems," PROC. IRE, vol. 47, pp. 1785–1786; October, 1959.

[5] G. Doetsch, "Theorie und Anwendung der Laplace Transformation," Dover Publications, Inc., New York, N. Y.; 1943.

[6] D. V. Widder, "The Laplace Transform," Princeton University Press, Princeton, N. J.; 1946.

[7] I. I. Hirschman and D. V. Widder, "The Convolution Transform," Princeton University Press, Princeton, N. J.; 1955.

[8] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y.; 1942.

[9] J. S. Bendat, "Principles and Applications of Random Noise Theory," John Wiley and Sons, Inc., New York, N. Y.; 1958.

[10] Y. H. Ku, "Analysis and Control of Nonlinear Systems," Ronald Press, New York, N. Y.; 1958.

[11] E. L. Ince, "Ordinary Differential Equations," Dover Publications, Inc., New York, N. Y.; (reprint) 1926. (First printing by Longmans, Green and Co., New York, N. Y.)

[12] F. J. Murray and K. S. Miller, "Existence Theorems," New York University Press, New York, N. Y.; 1954.

[13] J. R. Carson, "Electric Circuit Theory and the Operational Calculus," McGraw-Hill Book Co., Inc., New York, N. Y.; 1926.

The condition that $g_1(t)$ be periodic is not a necessary condition for (3), (4) and (5) to exist.[6,7,14] A weaker condition (more general) is that $g_1(t)$ be of exponential type, of exponential order one. Other conditions are given in references.[6-8,11,15]

Example

To check these concepts with our intuitive notions of transient, steady-state, and total response[16] as given above let us consider a simple example with a nonperiodic forcing function. Let

$$K(s) = \frac{\alpha}{s+\alpha} \tag{6}$$

be the system transmission and let

$$g_1(t) = 1. \quad \text{for all } t > 0. \tag{7}$$

Then utilizing (5) it is easy to show that

$$g_2(t) = 1 - e^{-\alpha t} \tag{8}$$

noting that

$$k(t) = \alpha e^{-\alpha t}. \tag{9}$$

The first term of the right member, namely, 1 is the steady-state term. The second term, namely, $e^{-\alpha t}$, is the transient term. The difference in these two is the total instantaneous response. These are well known ideas.[6-11] We see that these terms can also be calculated with the aid of (3) and (4), that is,

$$g_{2T}(t) = e^{-\alpha t} \tag{10}$$

$$g_{2S}(t) = 1 \tag{11}$$

which is consistent with our intuitive notions.

In order to perform calculations using (3) and (4) one must note that the integrals are *improper* because of the infinite upper limits.[15,17,18] Care must be exercised in evaluating such integrals, especially when the integrands are of the convolution type containing a variable parameter taking on all possible values, that is,

$$0 \leq t < \infty. \tag{12}$$

The calculation of $g_{2S}(t)$ is effected by its definition given in (13).

$$g_{2S}(t) = \lim_{t\to\infty} g_2(t). \tag{13}$$

The calculation of $g_{2T}(t)$ is most easily accomplished by subtracting the result of (13) from (5). Other methods of calculating (3) are described in the literature cited.[5-7,11,15,18] These precautions are only necessary when the kernel containing the parameter t "blows up" at infinity. Otherwise the calculation is generally straightforward.

The subject of improper convolution integrals and sums with varying parameters taking on, among other things, the value at infinity, are considered somewhat elsewhere.[5-7,11,19-23]

ACKNOWLEDGMENT

I should like to acknowledge that the idea of splitting the integral of (5) so as to display its transient and steady-state terms was first suggested to me by Dr. W. Bogoshian of the Moore School at the University of Pennsylvania. I later discussed this matter with the late Dr. H. Pender, who originally invited Dr. J. Carson to the Moore School in the spring of 1925 to deliver his famous lectures on Heaviside's method.

Thanks are also due Drs. J. G. Brainerd, and Y. H. Ku of the Moore School and Dr. B. Naumov of the Soviet Union.

I am very grateful for the constructive criticisms rendered by J. Dietz, M. Flomenhoft and T. Truitt especially regarding the improper integral calculations.

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Direction-Finding Experience and the Performance of Transmitting Navigational Aids*

When considering the factors affecting the performance of transmitting navigational aids which provide bearing indications directly, it is useful to remember the properties of the direction-finding systems which are their receiving analogs. This is true at the design stage and also after installation. Theoretical analyses of the site-error suppression properties of various types of direction finders and site-error measurements in the field are examples where direction-finding experience is of value when considering transmitting systems.

For theoretical studies of the effect of an interfering ray on the accuracy of the Doppler Omnirange described by Anderson and Flint,[1] the corresponding direction finder is the cyclical phase-measuring instrument[2] more recently designated as the commutated aerial direction finder;[3] the conventional narrow aperture VOR has its counterpart in the Adcock direction finder. That the transmitting systems use horizontally polarized waves and the direction finders referred to use vertical polarization is irrelevant when the error due to an interfering signal of specified strength relative to the direct signal is calculated. An account of the errors of these and other direction finders under this condition was given some years ago.[4] It is not surprising to find that the course-scalloping effects of a single interfering ray shown in (26) and Fig. 6 of the Anderson and Flint paper are identical with the site-errors already derived for the analogous direction finders.

Anderson and Flint (Fig. 7) use the ratio of the maximum course-scalloping occurring on the two omnirange systems, displayed as a function of azimuth, as a means of comparing the properties of the systems in relation to siting errors. In the direction-finding field it has been found useful to develop a figure of merit (site-error susceptibility) for each class of instrument, which takes account of the random distribution of obstacles on the site. Instrument classes are then compared by means of their site-error susceptibilities; the same method is applicable to transmitting navigational aids and the calculations already made for direction finders apply immediately.

The site-error susceptibility is derived by letting the direction of the direct ray sweep through 360° in azimuth at a constant angle of elevation, and then computing the error introduced into the system by a horizontally incident interfering ray of fixed azimuth. The amplitude of the interfering signal is set at 0.1 relative to the direct and its phase is adjusted as the azimuth of the direct ray varies, so that the bearing error produced is always the maximum value ϵ_m. Now ϵ_m is a function of the azimuth of the direct ray, and the root-mean-square value, ϵ_σ, of ϵ_m obtained by integration over the full 360° of azimuth is a convenient measure of the site-error (or scalloping) susceptibility of the system at the chosen angle of elevation. Using the calculations for the corresponding direction finders,[4] at zero angle of elevation ϵ_σ is 4.05° for the normal omnirange system and 0.11° for the Doppler system having the dimensions described by Anderson and Flint.

Since the purpose of this note is to urge that direction-finding experience, both theoretical and practical, is not lost sight of when considering transmitting navigational aids providing bearings, it may be appropriate to conclude by giving references to other direction finding papers[5-7] which appear relevant.

* Received by the IRE, November 27, 1959.

[1] S. R. Anderson and R. B. Flint, "The CAA Doppler Omnirange," PROC. IRE, vol. 47, pp. 808–821; May, 1959.

[2] C. W. Earp and R. M. Godfrey, "Radio direction finding by the cyclical differential measurement of phase," *J. IEE*, vol. 94, pt. IIIA, pp. 705–721; April, 1947.

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[7] W. C. Bain, "Possible errors of a particular wide aperture direction finder," *Proc. IEE* vol. 103, pt. C, pp. 313–324; March, 1956.

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Printed Aluminum Capacitors*

According to Guentherschulze and Betz it is practically impossible to apply metallic counter electrodes to oxidized Al sheets without losing the insulating behavior of these oxide films.[1] By anodic forming of evaporated aluminum films, however, and subsequent application of evaporated aluminum counter electrodes, we have been able to obtain solid electrolytic capacitors with very good characteristics regarding capacity per unit area, dissipation factor, leakage current, breakdown strength, number of rejects, and temperature dependency of the capacity and the dissipation factor.

Aluminum spectroscopically pure is evaporated onto a substrate, *e.g.*, glass or pyrex slides, by the usual evaporation technique (pressure 10^{-5} mm Hg) and subsequently oxidized anodically (10 to 20 minutes at room temperature). After cleaning the oxidized films with distilled water, the counter electrodes (0.10 cm²) are applied by evaporation of aluminum (Fig. 1).

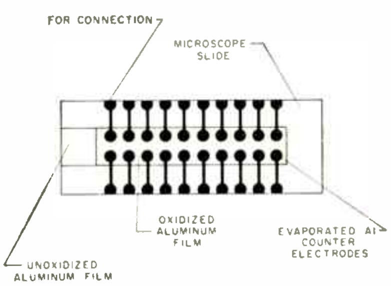

Fig. 1—Printed aluminum capacitor.

The capacity of such an Al capacitor, formed, *e.g.*, to 20 volts, is 0.31 μf/cm² with a leakage current of 0.005 μa/cm² at $\frac{2}{3}$ of the forming voltage and consequently a very small dissipation factor is obtained. In Table I data are given for the printed Al capacitors with different forming voltages. With lower forming voltages a larger scatter in the capacity values is obtained. The dissipation factor increases with decreasing forming voltage. Breakdown occurs usually at full forming voltage.

According to these results, a working voltage of about 90 per cent of the forming voltage can be tentatively suggested. This high breakdown strength compensates for the comparatively low dielectric constant of the aluminum oxide, so that for the same working voltage a smaller forming voltage (*i.e.*, a thinner oxide film) is permissible than for Ta. For printed Ta capacitors a working voltage of 50 per cent of the forming voltage is suggested.[2] Consequently, with Al capacitors the same capacity/unit area can be obtained as with sputtered Ta capacitors. It may be necessary to operate these Al capacitors at voltages somewhat lower than 90 per cent of the forming voltage in those circuits in which extreme reliability is required, but at the 90 per cent level, the reliability is adequate for most purposes.

In Fig. 2 the temperature dependency of the capacity is given in the temperature range from -200 to $+200$°C for a 45-volt formed capacitor. The dissipation factor, measured at a frequency of 1 kc (Fig. 3), shows also a slight increase with temperature in this temperature range. The leakage current at room temperature is smaller than 0.0001 μa/cm² at 30 volts.

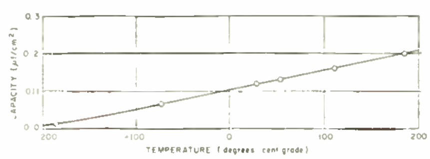

Fig. 2—Temperature dependence of the capacity of a printed aluminum capacitor, forming voltage 45 volts.

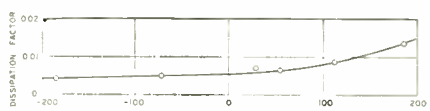

Fig. 3—Temperature dependence of the dissipation factor of a printed aluminum capacitor, forming voltage 45 volts.

This type of capacitor seems to be especially suitable for microminiaturized circuits. It is obvious that these Al capacitors are simple to make, for Al can be easily evaporated and no sputtering is necessary; they differ in this respect from tantalum condensers. Also the times required for forming are appreciably shorter than those reported by Berry and Sloan[2] for Ta (4–5 hours).

Experiments on these Al capacitors are being continued, especially in respect to lifetests; they will be reported later in more detail.

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TABLE I

Forming Voltage (V)	Capacity μf/cm²	Dissipation Factor	Leakage Current at 2/3 of Forming Voltage (μa/cm²)
20	0.310±8 per cent	0.050±0.030	0.005
30	0.158±6 per cent	0.043±0.013	0.001
45	0.112±3 per cent	0.017±0.006	<0.0001

* Received by the IRE, December 23, 1959.

[1] A. Guentherschulze and H. Betz, "Elektrolyt-Kondensatoren," Technischer Verlag Herbert Cram, Berlin, Ger., 2nd ed., p. 101; 1952.

[2] R. W. Berry and D. J. Sloan, "Tantalum printed capacitors," Proc. IRE, vol. 47, pp. 1070–1075; June, 1959.

Capacitance and Charge Coefficients for Parametric Diode Devices*

In this note, expressions are given for the capacitance and charge coefficients of parametric diode devices in terms of known and, in one important case, tabulated functions.

For a back-biased diode, the capacitance can be expressed as

$$C = C_{0b}(1 - v/A)^{-\nu}, \tag{1}$$

with $1 > \nu > 0$, v the voltage, $C_{0b} = C$ at $v = 0$, and A a constant. With V_b the back-bias voltage, V_0 the magnitude of the pump voltate, $\theta = \omega t$ (ω is the pump radian frequency), and $v = V_b + V_0 \cos\theta$,[1] (1) becomes

$$C = C_d(1 - \alpha\cos\theta)^{-\nu}, \tag{2}$$

with

$$C_d = C_{0b}(1 - V_b/A)^{-\nu}, \tag{3}$$

and

$$\alpha = V_0/(A - V_b). \tag{4}$$

Usually $V_b < 0$ and $V_0 < A - V_b$, to avoid forward conduction; hence $1 > \alpha > 0$. Expansion of (2) yields

$$C = C_0 + 2C_1\cos\theta + 2C_2\cos 2\theta + \cdots \tag{5}$$

with

$$C_n = \frac{C_d}{2\pi}\int_0^{2\pi}(1 - \alpha\cos\theta)^{-\nu}\cos n\theta\, d\theta, \quad n \geq 0. \tag{6}$$

Similarly for the charge with $Q = \int C dv$, from (1)

$$Q = \frac{-AC_{0b}}{1-\nu}(1 - v/A)^{1-\nu}, \tag{7}$$

where the integration constant has been dropped. Similar substitutions as above yield

$$\begin{aligned} Q &= Q_d(1 - \alpha\cos\theta)^{1-\nu} \\ &= Q_0 + 2Q_1\cos\theta + 2Q_2\cos 2\theta + \cdots \end{aligned} \tag{8}$$

with

$$Q_d = -\frac{C_{0b}A}{1-\nu}(1 - V_b/A)^{1-\nu} \tag{9}$$

* Received by the IRE, November 30, 1959; revised manuscript received, December 30, 1959.

[1] This assumption of a single frequency component across the capacitor is strictly valid only for a constant voltage (low internal impedance) source. Also, it neglects the significant harmonic voltage which would appear across the capacitor in the case of a high-efficiency harmonic generator.

and

$$Q_n = \frac{Q_d}{2\pi}\int_0^{2\pi}(1-\alpha\cos\theta)^{1-\nu}\cos n\theta\, d\theta,\ n\geq 0. \quad (10)$$

Comparison of (2) and (8) as well as (6) and (10) shows that if

$$C_n/C_d = f(\alpha, n, \nu) \quad (11a)$$

then

$$Q_n/Q_d = f(\alpha, n, \nu - 1). \quad (11b)$$

GENERAL VALUES OF ν

With

$$k^2 \equiv \frac{2\alpha}{1+\alpha}, \quad (12)$$

some changes in the integration variables allow (6) to be written

$$C_n = \frac{2C_d(-1)^n}{(1+\alpha)^\nu\pi}\int_0^{2\pi}\frac{\cos 2n\beta}{(1-k^2\sin^2\beta)^\nu}\, d\beta. \quad (13)$$

Note that $1>k>0$ for $1>\alpha>0$. Expansion of the denominator in (13) yields known integrals,[2] and there results, after some manipulation,

$$C_n = \frac{C_d}{(1+\alpha)^\nu}\sum_{r=n}^{\infty}P(r,n)T(\nu,r)k^{2r},\ n\geq 0, \quad (14)$$

with

$$P(r,n) = \frac{[\Gamma(r+1)]^2}{\Gamma(r+1-n)\Gamma(r+1+n)}, \quad (15)$$

and

$$T(\nu,r) = \frac{\Gamma(\nu+r)}{\Gamma(\nu)}\,\frac{\Gamma(r+1/2)}{\Gamma(1/2)}\,\frac{1}{[\Gamma(r+1)]^2}, \quad (16)$$

where $\Gamma(z)$ is the gamma function. Note that the summation in (14) starts with the $r=n$ term since all prior terms are zero. Also, $C_n>0$ for $n\geq 0$.

From (11) it is evident that

$$Q_n = Q_d(1+\alpha)^{1-\nu}\sum_{r=n}^{\infty}P(r,n)T(\nu-1,r)k^{2r},\quad n\geq 0. \quad (17)$$

Here with Q_d negative, $Q_0<0$, and $Q_n>0$ for $n\geq 1$.

Direct integration of (6) in terms of known functions is possible; with

$$\alpha = \frac{2t}{1+t^2}, \quad (18)$$

there results

$$C_n = (1+t^2)^\nu t^n\frac{\Gamma(n+\nu)}{\Gamma(\nu)\Gamma(n+1)}\cdot{}_2F_1(\nu, n+\nu; n+1; t^2)C_d,\quad n\geq 0, \quad (19)$$

where ${}_2F_1(a, b; c; z)$ is the hypergeometric function.[3] Note $1>t>0$ for $1>\alpha>0$. Since[3]

$${}_2F_1(\nu, n+\nu; n+1; x) = \frac{\Gamma(n+1)\Gamma(\nu)\Gamma(1-\nu)}{(-1)^n\Gamma(\nu+n)\Gamma(1-\nu+n)}\,\frac{1}{(1-x)^{\nu-1}}\,\frac{d^n}{dx^n}\cdot\left[(1-x)^{\nu+n+1}{}_2F_1(\nu,\nu;1;x)\right]. \quad (20)$$

C_n can be obtained from C_0 with $x=t^2$ in (20). Also, the use of

$$(1+t)^{2\nu}{}_2F_1(\nu,\nu;1;t^2) = {}_2F_1(\nu, 1/2; 1; k^2)\ ^4 \quad (21)$$

in (19) and (20) establishes the equivalence of C_n given by (19) and (14). The expression for C_n in terms of the hypergeometric function has, aside from its intrinsic interest, the advantage that various transformations, recursion formulas, and identities are available. For example, with k^2 near 1, (14) is slowly convergent, but the equivalence of (14) and (19) enables a highly convergent series in $1-k^2$ to be obtained easily.[4] Also from (19) and (20) and other hypergeometric function relationships, it can be shown that

$$C_{n+2} = -\frac{(n+\nu)}{(n-\nu+2)}C_n + \frac{(1+t^2)}{t}\,\frac{(n+1)}{(n-\nu+2)}C_{n+1}. \quad (22)$$

The hypergeometric function involved in C_0 can be expressed as a fractional-order Legendre function; thus[5]

$${}_2F_1\left(\nu,\nu;1;\tanh^2\frac{u}{2}\right) = \left(\cosh^2\frac{u}{2}\right)^\nu \mathcal{P}_{-\nu}(\cosh u), \quad (23)$$

where $t=\tanh u/2$, but no useful tables of these functions seem to be available.

From (11) and (19) it is clear that

$$Q_n = (1+t^2)^{\nu-1}t^n\frac{\Gamma(n+\nu-1)}{\Gamma(\nu-1)\Gamma(n+1)}\cdot{}_2F_1(\nu-1, n+\nu-1; n+1; t^2)Q_d,\quad n\geq 0, \quad (24)$$

and that the relationships noted in (20)–(23) relating to C_n apply equally well to Q_n with the substitution indicated in (11).

THE CASE OF $\nu=\frac{1}{2}$

In this important special case, evaluation results in terms of the tabulated complete elliptic integrals from (21)[6] or from (13).[7] Thus,

$$C_0 = \frac{2C_d}{(1+\alpha)^{1/2}\pi}K, \quad (25)$$

$$C_1 = C_0 - \frac{4C_d}{(1+\alpha)^{1/2}\pi k^2}[E-(1-k^2)K], \quad (26)$$

etc. K and E are the complete elliptic integrals of the first and second kind, respectively, with modulus k.

Similarly[8]

$$Q_0 = \frac{2Q_d}{\pi}(1+\alpha)^{1/2}E, \quad (27)$$

$$Q_1 = -Q_0 - \frac{4Q_d}{3\pi k^2}(1+\alpha)^{1/2}\cdot[(1-2k^2)E + K(k^2-1)], \quad (28)$$

etc.

The transmission phase shift in a parametric amplifier can be dependent on C_0. The influence of the amplitude of the pump voltage, V_0, on this parameter as derived from (25) is given by

$$\frac{\partial C_0}{\partial V_0} = -\frac{C_0}{2V_0}\left[1 - \frac{1}{1-\alpha}\,\frac{E}{K}\right]. \quad (29)$$

After completion of the above analysis and preparation of this note, it was brought to our attention that related work had already appeared in the document literature.[9] However, appearance of the independent work reported here still seems worthwhile. It not only calls more widespread and deserved attention to the prior work but also points out the unified and generalized results which accrue from the fact, apparently not previously noted, that the capacitance and charge coefficients can be expressed in terms of the hypergeometric function.

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[2] W. Magnus and F. Oberhettinger, "Formulas and Theorems for the Special Functions of Mathematical Physics," Chelsea Publishing Co., New York, N. Y., p. 5; 1949.
[3] *Ibid.*, p. 11.
[4] *Ibid.*, p. 9.
[5] *Ibid.*, p. 56.
[6] *Ibid.*, p. 10.
[7] P. F. Byrd and M. F. Friedman, "Handbook of Elliptic Integrals for Engineers and Physicists," Springer-Verlag, Berlin, Germany, pp. 192–193; 1954.
[8] *Ibid.*, p. 194. The integrals involved here can be obtained by differentiating with respect to k.
[9] D. I. Breitzer, R. Gardner, J. C. Greene, P. P. Lombardo, B. Salzberg, E. W. Sard, and K. Siegel, "Quarterly Progress Reports, Application of Semiconductor Diodes to Low-Noise Amplifiers, Harmonic Generators, and Fast-Acting TR Switches," Airborne Instruments Lab., Huntington, Long Island, N. Y., Rept. 4589-M-1, Appendix IV, September, 1958; Rept. 4589-I-4, Appendix E, June 1959; Contract AF 30(602)-1854.

Space-Charge Capacitors for Parametric Amplifiers*

In parametric amplifiers, which depend upon the voltage-sensitive nonlinearity of a reactive element for their action, it is desirable that such nonlinearity be as large as possible over the voltage range of interest. In addition, low-noise and high-frequency response require minimum in-phase (resistive) losses in the reactive element.[1] In view of the great current interest in parametric amplifiers, a short summary of some of the work on space-charge capacitors pertinent to their use as varactors is presented below.

Some time ago, the author investigated in detail the capacitance behavior of certain semiconductor space-charge regions near blocking electrodes[2–6] and pointed out[6] that such systems should be useful for parametric

* Received by the IRE, January 29, 1960.
[1] W. E. Danielson, "Low noise in solid state parametric amplifiers at microwave frequencies," *J. Appl. Phys.*, vol. 30, pp. 8–15; January, 1959.
[2] J. R. Macdonald and M. K. Brachman, "Exact solution of the Debye-Hückel equations for a polarized electrode," *J. Chem. Phys.*, vol. 22, pp. 1314–1316; August, 1954.
[3] J. R. Macdonald, "Static space-charge effects in the diffuse double layer," *J. Chem. Phys.*, vol. 22, pp. 1317–1322; August, 1954.
[4] J. R. Macdonald, "Theory of the differential capacitance of the double layer in unadsorbed electrolytes," *J. Chem. Phys.*, vol. 22, pp. 1857–1866; November, 1954.
[5] J. R. Macdonald, "Static space charge and capacitance for a single blocking electrode," *J. Chem. Phys.*, vol. 29, pp. 1346–1358; December, 1958.
[6] J. R. Macdonald, "Static space charge and capacitance for two blocking electrodes," *J. Chem. Phys.*, vol. 30, pp. 806–816; March, 1959.

amplifiers since they may yield exponential dependence of capacitance on applied voltage as well as low resistance losses. In the earlier work,[2–4] the situation was considered where mobile charge of both signs is blocked (no transfer of charge carriers between the charge-containing material and the metallic electrodes) at one or two electrodes, and an essentially exponentially increasing dependence of differential capacitance on applied bias voltage was obtained. Here, charges of both signs are mobile, and the capacitance increases for either sign of the applied voltage because for either sign an accumulation layer of excess charge (space-charge) forms in the region of the blocking electrode(s). This situation applies to the case of intrinsic semiconduction in a solid or to a completely dissociated univalent electrolyte.

Later, the case was considered where charge of only one sign is mobile; the charge carriers are blocked at one[5] or two[6] electrodes; and mobile charge can recombine to any degree with fixed charge of opposite sign arising from the dissociation or ionization of neutral centers distributed uniformly throughout the material. This situation corresponds to photoconduction in insulators (*e.g.*, F-centered alkali-halide crystals) or to extrinsic semiconduction in materials sufficiently doped or with a large enough bandgap that the effects of mobile minority charge carriers may be neglected. In the case of complete dissociation, where the recombination parameter R is zero, the situation is essentially that treated by Schottky[7] and Spenke[8] in their theory of rectification. This theory was applied to such systems as copper-oxide and selenium rectifiers which show partial (noninfinite barrier height) blocking at a metallic electrode.

The capacitance behavior of a system with only negative charges mobile, but blocked at one electrode,[5] is shown in Fig. 1 for different degrees of recombination R. Here C_d is the differential capacitance, and C_0 is its value in the limit of zero potential difference across the space-charge region. For a positive potential difference across the space-charge region, essentially exponentially increasing capacitance is obtained no matter what the value of R since a negative space-charge accumulation layer is set up at the blocking electrode. On the other hand, a negative potential difference yields a charge depletion layer for $R=0$ whose capacitance eventually decreases as the inverse square root of the bias voltage. For large R, however, recombination essentially mobilizes the immobile charge, and again an exponentially increasing capacitance can occur for a limited voltage range.

Because it is not easy and is sometimes impossible to produce an electrode-solid interface which approximates well to ideal blocking behavior over an appreciable applied voltage range, the effect of an artificial blocking layer was also considered in detail in the earlier work.[5,6] Here, the charge-containing material is abutted by an essentially charge-free region between it and a metallic electrode. Such a region can be produced by an oxide layer on the material or by the interposition of a thin insulating layer of such material as mica or mylar. If this charge-free region is sufficiently thin, its capacitance, which is essentially in series with any space-charge capacitance, may be much larger initially than the latter, and a considerable range of exponential over-all capacitance increase may still be possible. Fig. 2 shows the calculated behavior for different values of the ratio C_c/C_{ser}, where C_{ser} is the capacitance of the charge-free layer and C_c is the combined capacitance of C_d and C_{ser}. Note especially that the value of C_c/C_{ser} determines the potential range over which the effect of the space-charge capacitance is important. Because the potential across the charge-free region and that across the space-charge region are inversely proportional to their respective capacitances, and the space-charge capacitance is a nonlinear function of the potential across it, a transcendental equation must be solved to obtain the final dependence of C_c on over-all applied potential.[9] The solution of such an equation can be avoided, however, if the potential across the space-charge layer is taken as the main variable and the over-all applied potential derived from it. This situation has not been well

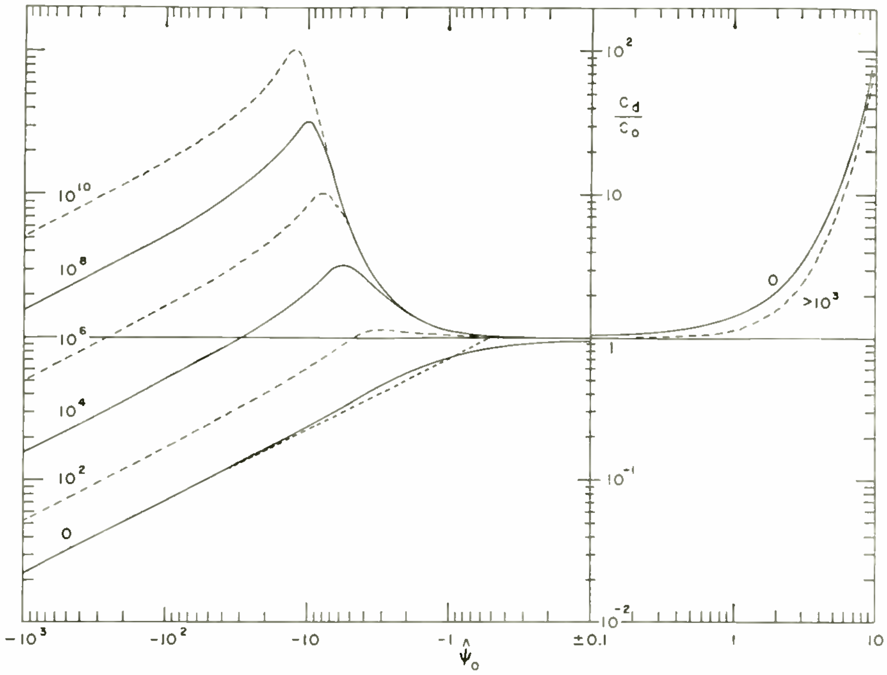

Fig. 1—Relative space-charge differential capacitance versus normalized potential ($\hat{\psi}_0 = e\psi_0/kT$) across the space-charge region for various values of the recombination parameter R.

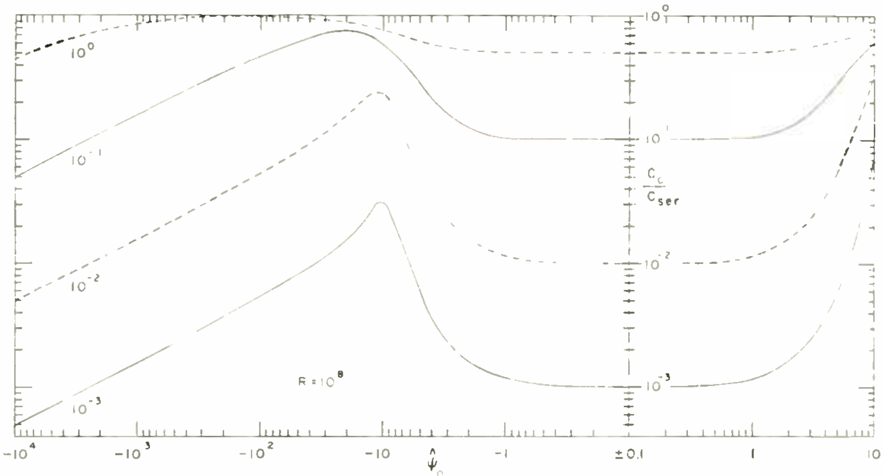

Fig. 2—Dependence on normalized potential of relative space-charge differential capacitance of a system comprising a voltage-independent capacitance C_{ser} in series with a voltage-dependent space-charge capacitance C_d for various values of the ratio C_0/C_{ser} and $R = 10^8$.

[7] W. Schottky, "Vereinfachte und erweiterte Theorie der Randschichtgleichrichter," *Zeits. für Phys.*, vol. 118, pp. 539–592; September-October, 1941.

[8] E. Spenke, "Zur Randschichttheorie der Trockengleichrichter," *Zeits. für Phys.*, vol. 126, pp. 67–83; January-February, 1949.

[9] On using the condition that the charge in the initial layer is equal to that in the space-charge region (*i.e.*, no discontinuity in the normal component of dielectric displacement at the interface between the regions), one sees that the static (not differential) capacitance of the space-charge region must be used in this equation. Note that the potential variable in Figs. 1 and 2 includes in it the contribution of any "built-in" or barrier potential.

appreciated in the past, but was discussed as early as 1954 in the case of charges of both signs mobile.[4]

A capacitor made up of a plate of n-type silicon with an oxide layer covered by a metallic contact on one side and either the same situation or an ohmic contact on the other side may possibly be a good realization of the state of affairs discussed above. Such a capacitor has been under investigation in a number of laboratories for some time. Because an oxide or other intermediate layer may be made essentially charge-free with consequent very low conduction, the resistive losses of the over-all device may, in principle, be considerably lower than those of a p-n diode, especially in regions where exponential capacitance variation may be obtained. Terman[10] has, in fact, found a change in over-all capacitance as high as 10:1 for a two-volt bias change. This change seems to be largely an exhaustion decrease rather than an accumulation increase in capacitance, but Dewald[11] has apparently observed such an increase for zinc oxide in contact with an electrolyte, a blocking-electrode situation. Wallmark[12] has suggested the use of an oxide layer device to yield improved gate control of a field effect transistor. Recently, Pfann and Garrett have also discussed the properties of such systems in a qualitative way and have pointed out some of the frequency-response characteristics of these space-charge devices.[13]

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[10] L. Terman, "Silicon Oxide Capacitors," Solid-State Electronics Lab., Stanford University, Stanford, Calif., Consolidated Quarterly Status Rept. No. 4, pp. 20–27; July 1 to September 30, 1959. See also J. L. Moll, "Variable capacitance with large capacity change," 1959 IRE WESCON Convention Record, pt. 3, pp. 32–36.

[11] J. F. Dewald, "The charge and potential distributions at the zinc oxide electrode," *Bell Sys. Tech. J.*; to be published.

[12] J. T. Wallmark, U. S. Patent No. 2900531, August 18, 1959.

[13] W. G. Pfann and C. G. B. Garrett, "Semiconductor varactors using surface space-charge layers," Proc. IRE, vol. 47, pp. 2011–2012; November, 1959.

An Electrostatically Focused Electron Beam Parametric Amplifier*

Electron beam parametric amplifiers employing magnetic focusing have been described.[1,2] In one of these devices,[1] gain is achieved by pumping the fast space charge wave. In the other device,[2] gain is achieved by pumping the fast cyclotron wave. Thus, in the latter device, the magnetic field plays a vital role in addition to that of focusing the beam. The electron beam parametric amplifier which is proposed here employs electrostatic focusing of an electron sheet beam and achieves gain by pumping the fast wave of the natural resonant electron frequency associated with the electrostatic focusing fields.

Consider a sheet beam which passes between a series of pairs of identical planar plates at alternate dc voltages V_1 and V_2 (see Fig. 1). The voltage difference between adjacent pairs of plates forms an electrostatic field which acts to focus the beam by creating a time average restoring force, accelerating the electrons towards the plane midway between the plates.[3] The electrons in such a focusing system have a natural transverse resonant frequency f_e first described by Adler.[3] A more exact expression for f_e is[4]

$$f_e = 0.187 \times 10^6 \left(\frac{\epsilon}{a}\right) \sqrt{V_0} \cdot \left[\sum_{n=1,3,5,7\ldots}^{\infty} \frac{1}{\cosh^2 (n\pi d/2a)} \right]^{1/2}, \quad (1)$$

where

V_0 = the space average beam voltage

$$= \frac{V_1 + V_2}{2},$$

$$\epsilon = \frac{V_2 - V_1}{2V_0},$$

a = the periodicity of the focusing plates (Fig. 1), and

d = the separation between opposing plate pairs (Fig. 1).

In both (1) and (2), all terms are expressed in mks units.

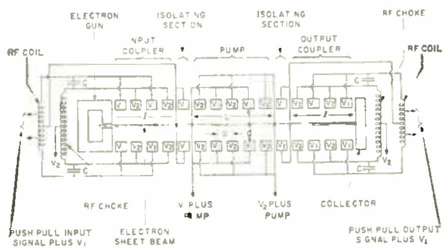

Fig. 1—A simplified drawing of an electrostatically focused parametric amplifier.

The means by which parametric amplification is achieved in this electrostatically focused system can be most simply described by dividing the plates into three groups with respect to RF: the input coupler, the pump, and the output coupler (Fig. 1). The plates of the input coupler are connected so that they form the capacitance of a tank circuit, wherein all of the plates on each side of the beam are at the same RF potential. The tank circuit should be designed to resonate at approximately the frequency of the signal to be amplified, f_s. The output coupler is identical to the input coupler. The leads for the input and output signals are tapped off the RF coils (Fig. 1) to effect a transformer action which reduces the mismatch between the resistive loading of the coupler[5] and the characteristic impedance of the transmission line. Balanced input and output arrangements are shown; however, an unbalanced system could also be used.

The theory of Cuccia applies to the input and output couplers. Thus, if $f_s \approx f_e$, and if the Cuccia equation[5] is satisfied, all of the energy introduced across the plates of the input coupler will be transferred to the sheet beam traveling through the coupler plates in the form of a transverse oscillatory motion of the beam. Simultaneously, the inherent noise will be stripped off the beam. The output coupler performs the inverse function of the input coupler; *i.e.*, it acts to convert the transverse motion of the sheet beam entering the coupler into an RF signal in the load connected to the output coupler.

Gain is achieved by increasing the amplitude of the transverse oscillation of the electron sheet beam as it travels from the input coupler to the output coupler. This increase is accomplished in the pumping section by superimposing a frequency of approximately twice the electron resonant frequency on both voltages V_1 and V_2 in a push-pull manner, as shown in Fig. 1. It may also be possible to achieve satisfactory pumping by modulating only one of these two voltages and keeping the other one constant. The modulation of the focusing voltages varies the restoring force acting on the electron beam. The increase of amplitude of the transverse oscillation of the electron beam is achieved in a manner analogous to the building up of the amplitude of oscillation of a pendulum by varying either the restoring force or the length of the pendulum at exactly twice its natural resonant frequency.[6] In direct comparison with this pendulum model, and assuming the pumping signal is properly synchronized with respect to f_e, the gain will vary as $e^{2\alpha L}$, where αL is given by

$$\alpha L = \frac{\pi \Delta f_e}{u_a} L \quad (2)$$

and where

L = the length of the pump,

Δf_e = the peak variation of f_e due to pumping,

u_a = the average beam velocity.

If $f_s \approx f_e$ but $f_s \neq f_e$, an idler frequency equal to the difference between the pump and signal frequencies will also be present on the beam as it leaves the pumping section. Under this condition,[2] the gain of the signal frequency will vary as $\cosh^2 \alpha L$, and the gain of the idler frequency will vary as $\sinh^2 \alpha L$. The value of Δf_e corresponding to the peak values of the applied pumping voltages may be determined from (1). There is a limit to the amplitude of the pumping signal that may be applied to the electrostatically focused parametric amplifier. The applied pumping voltages must be limited to that range over

* Received by the IRE, January 28, 1960.

[1] A. Ashkin, "Parametric amplification of space charge waves," *J. Appl. Phys.*, vol. 29, pp. 1646–1651; December, 1958.

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

[3] R. Adler, O. M. Kromhout, and P. A. Clavier, "Resonant behavior of electron beams in periodically focused tubes for transverse signal fields," Proc. IRE, vol. 43, pp. 339–341; March, 1955.

[4] W. E. Waters, "Periodic focusing of thin electron sheet beams," *DOFL Tech. Rev.*, vol. 2, pp. 1–25; July, 1959.

[5] C. L. Cuccia, "The electron coupler—a developmental tube for amplitude modulation and power control at ultra-high frequencies," *RCA Rev.*, vol. 10, p. 278 (18); June, 1949.

[6] G. Wade and R. Adler, "A new method for pumping a fast space-charge wave," Proc. IRE, vol. 47, pp. 79–80; January, 1959.

which a high percentage of beam transmission is maintained through the pumping section.

The isolating sections shown in Fig. 1 are included for the purpose of preventing coupling between the input and output couplers. This coupling could occur if asymmetries were present in the couplers.

Estimates can be made of some practical dimensions and voltages of an electrostatically focused parametric amplifier designed to have a gain of 20 db at 500 mc: $a=0.050$ inch, $d=0.012$ inch, $w=0.5$ inch, $l=0.5$ inch, $L=4.5$ inches, $V_1=70$ volts, and $V_2=130$ volts. However, these values do not purport to represent an optimum design. All of these symbols, except w, are explained in Fig. 1. The w represents the width of the focusing plates, the dimension perpendicular to the plane of Fig. 1. The length of the pumping section, L, was determined from (2) based on the assumption that the value of Δf_c is 10 per cent of f_c. This value of Δf_c appears to be conservative from the standpoint of the range of f_c over which constant beam transmission may be obtained in dc tests. It is expected that the beam current will be somewhat less than that given by the theoretical equilibrium perveance (minimum beam rippling) for this structure. Assuming beam dimensions of 0.400 inch by 0.003 inch, the current corresponding to the equilibrium perveance would be about 2.25 ma. The total length of the structure, including an electron gun which generates a well-collimated sheet beam, a collector, and 112 pairs of focusing plates, would be about 6.6 inches. It should be noted that a technique for constructing tubes employing planar periodic electrostatic focusing of a sheet beam has been developed, in which both excellent mechanical alignment and ruggedness may be achieved.[7]

Two means by which the frequency of operation of the electrostatically focused parametric amplifier may be shifted will now be considered: 1) if the dc and pumping voltages are kept constant and all of the tube and electron beam dimensions are scaled, the frequency will vary inversely as the dimensional scaling factor; 2) if the tube dimensions are kept constant and both dc and pumping voltages are scaled, the frequency will vary directly as the square root of the voltage scaling factor. In both of these scaling processes, the tube gain and the beam perveance through the focusing structure will remain unchanged.

If the same beam current is maintained through each of these scaling processes, the Cuccia equation[5] will continue to be satisfied. If the same beam current is not maintained, for example because of limitations of the electron gun in the dimensional scaling, then one or more parameters must be adjusted to satisfy the Cuccia equation.

The practical limit to which the dimensions a and d of the structure discussed previously for operation at 500 mc could be scaled down would probably be a factor of two. By also increasing the voltages by a factor of four, it should be possible to achieve 2000-mc operation. The limit to which the operating frequency could be raised by increasing the beam voltage would probably be determined by the amount of pumping power that could be tolerated plus the voltage difference that could be withstood between adjacent focusing plates.

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[7] B. J. Udelson and M. R. Bradley, "A Method of Constructing Tubes Employing Planar Periodic Electrostatic Focusing of Sheet Beams," presented at the Conference on Electron Devices, Washington, D. C.; October 30, 1959.

A Non-Return to Zero (*NRZ*) Mode of Operation for a Magnetostrictive Delay Line*

In conventional operation of a magnetostrictive delay line, a digital "one" is obtained by generating in the transmitter coil a current pulse whose width is equal to the time of propagation, τ, through identical transducers.[1] A digital "zero" is indicated by the absence of a current pulse. The minimum digit spacing for maximum resolution is 2τ.

This note describes a non-return to zero (*NRZ*) mode of operation for a magnetostrictive delay line in which a digital "one" is generated by a change in current level in the transmitter coil from zero to maximum or vice versa. A digital "zero" is indicated by no change in current level. The minimum *NRZ* spacing for maximum resolution is τ. Therefore, the maximum digital clock rate of any given magnetostrictive delay line can be doubled by this *NRZ* technique.

The theory of *NRZ* operation may be illustrated by transmitting along a magnetostrictive delay line a digital message, 1101, of the form and spacing shown in Fig. 1(a). Each digital period is equal to τ, the time of propagation through identical transducers. A digital "one" is indicated by a pulse at the beginning of a digital period. A digital "zero" is indicated by no pulse. This waveform is fed to a counting flip-flop at whose output it appears as in Fig. 1(b). The flip-flop output is the shape of the current waveform in the transmitter coil of the delay line.

The idealized voltage output from the receiver coil will appear as in Fig. 1(c), after a time delay of $n\tau$. This waveform is a linear addition of the voltage doublets generated by the rising and falling edges of the current waveform in the transmitter coil.[1] The rising wavefront at t_0, Fig. 1(d), produces the doublet at Fig. 1(e). The falling wavefront at t_1, Fig. 1(f), generates the doublet at Fig. 1(g). The rising wavefront at t_3, Fig. 1(h), produces the doublet at Fig. 1(i).

The receiver voltage waveform is amplified to clipping levels and is processed through the transistor circuitry shown in Fig. 2 to decode the original digital message.

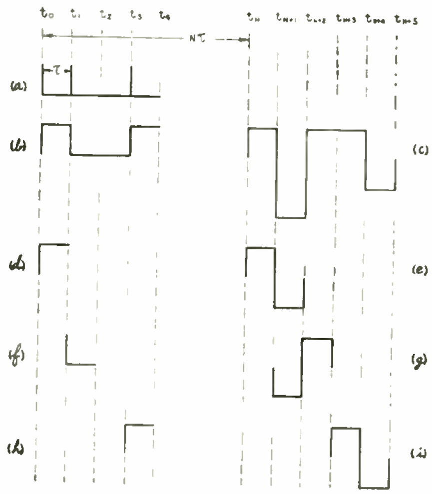

Fig. 1—Ideal transducer waveforms for *NRZ* operation.

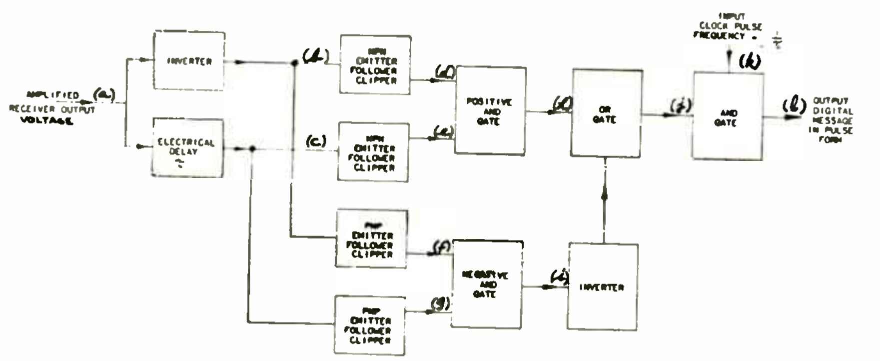

Fig. 2—Block diagram of receiver circuitry for *NRZ* operation.

The idealized voltage waveforms at the various lettered points (a, b, c, etc.) in Fig. 2 are shown in Fig. 3 with corresponding titles.

The processed receiver waveforms shown in Fig. 3(b) and 3(c) are complementary waveforms which have a relative time displacement of τ. The positive portion of each waveform is removed by clipping to obtain Fig. 3(d) and 3(e), which are AND gated to produce the resultant waveform at Fig. 3(h). Likewise the negative portions of the processed receiver waveforms are clipped to

* Received by the IRE, December 16, 1959.

[1] L. Rosenberg and A. Rothbart, "Electrical design of the transducer networks of a magnetostrictive delay line," 1958 IRE NATIONAL CONVENTION RECORD, pt. 2, pp. 92–101.

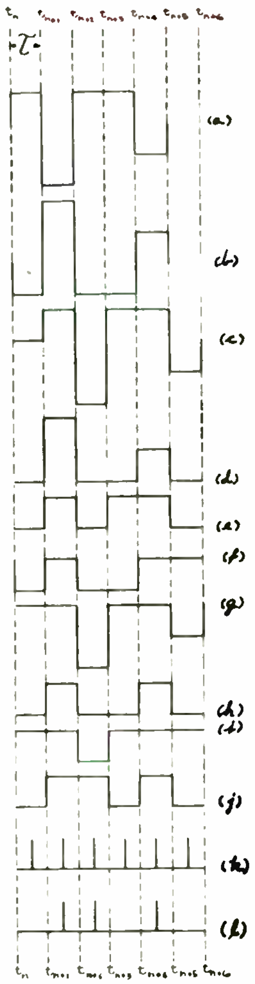

Fig. 3—Waveforms for NRZ circuitry.

obtain Fig. 3(f) and 3(g) which are AND gated to produce the waveform at Fig. 3(i). This output waveform is inverted and combined with Fig. 3(h) to provide the composite waveform at Fig. 3(j) which is strobed by the clock pulses, Fig. 3(k), to obtain the original input message 1101.

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Calculation of the Rise and Fall Times of an Alloy Junction Transistor Switch*

The rise and fall times of a junction transistor switch were first calculated by J. L. Moll.[1] These calculations were based on a linear equivalent circuit model of the grossly nonlinear transistor. The purpose of this note is to show a method of calculating the rise and fall times of a current-driven saturating alloy junction transistor switch from a base charge analysis.[2]

In order to turn the transistor ON (or OFF) a certain amount of charge must be supplied to the base lead of the transistor. This charge must be supplied to 1) change the active charge distribution in the base so that the collector current may change, 2) charge the collector depletion layer capacitance and 3) "replenish" that charge lost through bulk recombination of minority carriers in the active base region.

Hence, for the common-emitter connection for the total charge supplied to the base lead of the transistor during the transition time, one may write

$$\int_0^t I_B dt' = \int_0^t \frac{dq}{dt'} dt' + \int_0^t \frac{q}{\tau} dt' + \int_0^t c_{TC} \frac{dv_{cb}}{dt'} dt'. \qquad (1)$$

where

I_B denotes the constant base current,
t' denotes the time variable,
q is the total active charge in the base,
τ denotes the minority carrier lifetime in the base region, and
v_{cb} is the collector to base voltage.

In (1) it has been assumed that no charge is stored in regions external to the active base region. The charge that must be supplied to the emitter-base depletion layer capacitance (c_{TE}) has not been neglected but is accounted for in the term for the active charge.

By differentiating (1) with respect to t, one obtains

$$I_B = \frac{dq}{dt} + \frac{q}{\tau} + c_{TC} \frac{dv_{cb}}{dt}. \qquad (2)$$

In order to solve (2) for the rise (or fall) time, q and v_{cb} must be related to i_c, the collector current. For the analysis here, the approximation will be used that i_c is directly proportional to the total active charge in the base:

$$i_c = \bar{\omega}_T q. \qquad (3)$$

where $\bar{\omega}_T$ is defined as the average current-gain bandwidth product over the change of i_c and v_{cb}.

The variation of c_{TC} with v_{cb} for alloy junction transistors is taken as

$$c_{TC} = \frac{k}{\sqrt{\phi - v_{cb}}}. \qquad (4)$$

where ϕ is the equilibrium barrier potential and k is a constant.

Substituting (3) and (4) into (2), letting $v_{cb} = v_c$, and writing the resulting equation in integral form yields

$$\bar{\omega}_T I_B \int_0^{t_r} dt = \int_0^{I_{cs}} \frac{di_c}{1 - \frac{i_c}{\tau \bar{\omega}_T I_B}} + \frac{kR_L \bar{\omega}_T}{\sqrt{\phi + V_{CC}}} \cdot \int_0^{I_{cs}} \frac{di_c}{\left(1 - \frac{i_c}{\tau \bar{\omega}_T I_B}\right)\left(1 - \frac{i_c R_L}{\phi + V_{CC}}\right)^{1/2}}. \qquad (5)$$

where t_r is the rise time and is defined as the time interval after initiation of I_B until i_c reaches its saturated value, $I_{CS} = V_{CC}/R_L$. Integrating each term in (5) and setting $\tau\bar{\omega}_T = \beta_0$ (the low-frequency common-emitter current gain), we obtain the rise time as

$$\frac{t_r}{\tau_R} = \frac{\delta}{\sqrt{\frac{\beta_0}{\beta_c} - 1}} \tan^{-1}\left[\frac{1}{\sqrt{\frac{\beta_0}{\beta_c} - 1}}\right] + \ln\left[\frac{1}{1 - \frac{\beta_c}{\beta_0}}\right], \qquad (6)$$

where

$$\tau_R = \beta_0/\bar{\omega}_T,$$

$$\delta = \frac{2k\bar{\omega}_T R_L}{\sqrt{\phi + V_{CC}}},$$

and

$$\beta_c = \frac{I_{CS}}{I_B} \text{ (turn-on circuit beta).}$$

This equation is shown in Fig. 1. The fall time is obtained from (6) by replacing

$$\frac{1}{1 - \frac{\beta_c}{\beta_0}} \quad \text{with} \quad 1 + \frac{\beta_{c0}}{\beta_0}.$$

where β_{c0} is the TURN-OFF circuit beta, I_{CS}/I_B(OFF).

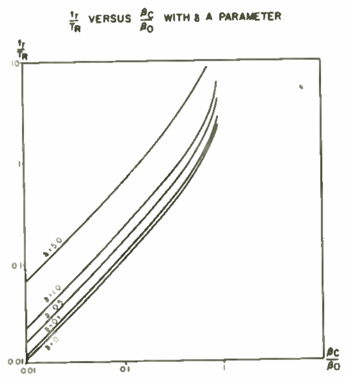

Fig. 1.

The rise time calculated on the basis of a linear circuit model is

$$\frac{t_r}{\tau_R} = (1 + \delta) \ln \frac{1}{1 - \frac{\beta_c}{\beta_0}}. \qquad (7)$$

Eq. 7 indicates that the collector current response is exponential, while (6) shows that this is not true.

* Received by the IRE, January 7, 1960.

[1] J. L. Moll, "Large signal transient response of junction transistors," Proc. IRE, vol. 42, pp. 1773–1784; December, 1954.

[2] R. Beaufoy and J. J. Sparkes, "The junction transistor as a charge controlled device," ATE J., vol. 13, April, 1957.

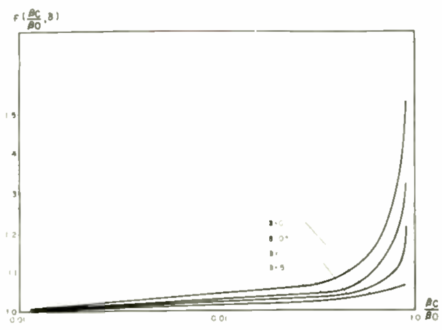

Fig. 2.

The results obtained from the more fundamental approach employed herein agree very closely with the modified Ebers and Moll equation. For $\beta_c/\beta_0 \leq 0.5$ and $\delta \leq 5$, (6) and (7) agree closely as shown by $F(\beta_c/\beta_0, \delta)$ in Fig. 2. $F(\beta_c/\beta_0, \delta)$ is the ratio of (6) to (7). As β_c/β_0 approach zero, (6) and (7) both approach $\beta_c/\beta_0(1+\delta)$.

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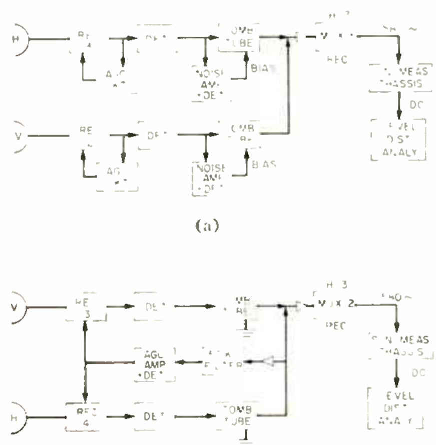
Experimental Comparison of Equal-Gain and Maximal-Ratio Diversity Combiners*

When two or more relatively independent paths are used for diversity reception of a fading radio wave, the problem arises of combining the signals from these channels in some manner that will yield an output that is optimum or nearly so. The earliest technique was what is usually called selection diversity, which does not involve combining the signals, but only selects the best of the channels and completely suppresses the other. It has been shown, however, that other schemes will yield better results, and for several years there has been considerable interest in maximal-ratio diversity combiners.[1-6] In fact, several "engineer-years" have been spent in the design and development of equipment that would yield an output SNR that was indeed maximal. The problems arise in the derivation of a weighting factor to be applied to each signal as required by the maximal-ratio system. This factor must be obtained by a continuous measurement of the SNR of the output of each receiver [see Fig. 1(a)]. The boxes labeled "noise amplifier and detector" are used to generate dc voltages which are applied as bias to the combiner tubes. These bias voltages in conjunction with the characteristics of the combiner tubes form the weighting factors. This system assumes that the AGC circuit holds the signal constant so that the SNR for the receiver is a function of the measurement of the noise. The operational difficulties are those involved in measuring the noise and making the gain of the several units track together over a wide range of input SNR.



(a)

(b)

Fig. 1—Basic maximal-ratio and equal-gain combining systems.

Recently, it has been shown that a type of combining known as equal-gain combining would closely approach the performance of maximal-ratio systems.[3,7] The equal-gain combining system does not require the measurement of the SNR of the receivers, as no weighting factors are necessary, but requires only that the gains of the receivers be equal at all times. This is done by using a common AGC circuit for all receivers. Referring to Fig. 1(b), note the relative simplicity of the equipment configuration. The ground symbols on the combiner tubes indicate that no variable bias is used, for they operate as mixing tubes only.

To determine experimentally the relative effectiveness of the two systems in practice, measurements were made recently using the Lincoln Laboratory tropospheric scatter circuit from Millstone Hill, Mass., to Sauratown Mountain, N. C.[8] The system normally employs fourth-order space diversity and uses maximal-ratio combining of the SSB signals. Fig. 2 shows the over-all system layout. The fading distribution on any one path is normally very close to Rayleigh. For the comparison test, two receivers were modified for equal-gain combining while the remaining pair operated normally. Fig. 1 shows the equipment configuration including the measuring apparatus. This system uses a transmitted pilot for AGC and demodulation purposes. Since this pilot disappears after detection, it was necessary to transmit another signal for equal-gain AGC. A 1-kc tone put into channel 7 of the multiplex was used. This translates to a 37-kc signal at baseband. A four-stage tuned amplifier and a detector were used to extract the pilot and convert it to dc to be fed to the IF AGC busses. A pair of receivers were used that had closely matching gain vs AGC voltage curves. This eliminated the need for separate dc amplifiers whose gains could be adjusted to compensate for different receiver gain characteristics. The maximal-ratio receivers were operated in conventional manner with their combiners adjusted for as near optimum as is possible in normal practice.

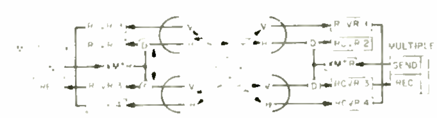


Fig. 2—UHF-SSB long-range communication system.

To determine the demodulated SNR for each pair of receivers, a 580-cps tone was transmitted on multiplex channel 3. This translates to 16,580 cps at baseband, which is roughly halfway between the two pilots used in this test. The combined outputs of each receiver pair were sent to separate multiplex demodulators. The 580-cps signals were then examined by two SNR measuring devices, the dc outputs of which fed two level distribution analyzers calibrated from 0 to 27 db SNR.

A total of 66 half-hour distributions were taken. Since the differences expected were small, the sets of measuring equipment were interchanged between runs in an effort to balance out calibration errors. The distributions were averaged per-cent-time-wise in pairs, then plotted and the pointwise db values tabulated and averaged. Fig. 3 shows the final average distributions adjusted to zero db signal-to-noise at the maximal-ratio median. The maximal-ratio curve shows a very small improvement in realized signal-to-noise. Fig. 4 shows the difference between the curves in larger scale. To give an idea of the spread of the data, the ranges within which 50 and 90 per cent of the differences fell are indicated by the vertical bars. Note that the difference is quite small; less than one db over most of the range. The dotted line shows the theoretical difference.[3,7] Note the close agreement between the theoretical and experimental curves. This is quite promising, considering that essentially no design was done for the equal-gain combiner used in this test. Since the combining was postdetection, all that was necessary was to provide an AGC circuit. This was improvised in the field in a few hours.

* Received by the IRE, January 12, 1960. The work reported in this paper was performed at Lincoln Laboratory, a center for research operated by Massachusetts Institute of Technology with the joint support of the U. S. Army, Navy, and Air Force.

[1] L. R. Kahn, "Ratio squarer," PROC. IRE, vol. 42, p. 1704; November, 1954.

[2] D. G. Brennan, "On the maximum signal-to-noise ratio realizable from several noise signals," PROC. IRE, vol. 43, p. 1530; October, 1955.

[3] F. J. Altman and W. Sichak, "A simplified diversity communication system for beyond-the-horizon links," IRE TRANS. ON COMMUNICATIONS SYSTEMS, vol. CS-4, pp. 50-55; March, 1956.

[4] H. Staras, "The statistics of combiner diversity," PROC. IRE, vol. 44, pp. 1057-1058; August, 1956.

[5] J. N. Pierce, "Diversity Improvement in Frequency-Shift Keying for Rayleigh Fading Conditions," Air Force Cambridge Research Center, Tech. Rept. 56-117; September, 1956.

[6] J. N. Pierce and S. Stein, "Multiple diversity with nonindependent fading," PROC. IRE, to be published.

[7] D. G. Brennan, "Linear diversity combining techniques," PROC. IRE, vol. 47, pp. 1075-1102; June, 1959.

[8] B. E. Nichols, "The AN/FRC 47-(XD-1) Mock-up System Performance Results," Lincoln Lab., Lexington, Mass., TR-203, pp. 3-5; April 23, 1959.

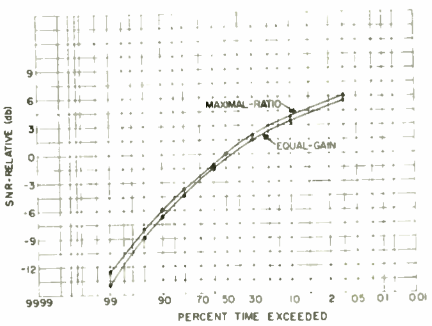


Fig. 3—Average SNR distributions for the two systems.

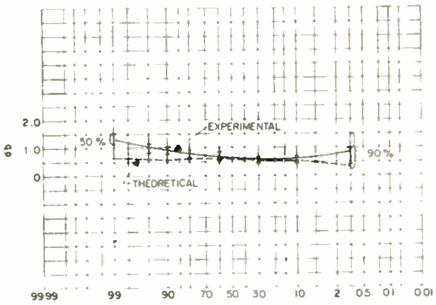


Fig. 4—Difference of distributions in Fig. 3 compared with theoretical.

A predetection equal-gain combining-scheme has been used for at least one operating FM tropo scatter link,[9] but to the author's knowledge has never been implemented for a long-haul SSB circuit. If this was done using some fraction of the design time that has been spent on maximal-ratio combiners, a predetection equal-gain system might be developed for SSB that could outperform the postdetection maximal-ratio type in practice. As shown by Adams and Mindes, the savings in system complexity are quite impressive.[10] Even with the necessary phase control, the stability and ease of adjustment are such that equal-gain system shows great promise and should be considered for inclusion in present and future communication systems.

Acknowledgment

The author wishes to acknowledge the assistance of T. McClean of the Mitre Corporation, Bedford, Mass., in obtaining the experimental data and Miss Cynthia Dexter of Lincoln Laboratory, Lexington, Mass., in the reduction of the data and preparation of the graphs. Special appreciation is due Dr. D. G. Brennan of Lincoln Laboratory for helpful criticism at all stages of this work.

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The Solutions for Nonuniform Transmission Line Problems*

There have been a great number of papers[1,2] on nonuniform transmission line problems. The practical application of the nonuniform transmission line has been predominantly in tapered matching sections. Other fields of electronics such as waveguides and traveling-wave tubes also utilize nonuniform transmission lines. But the exact solution has not been obtained for a general nonuniform transmission line. This correspondence presents exact solutions for three special cases, starting from the assumed forms of solution of the problem.

The differential equations describing propagating voltage $V(x)$ and current $I(x)$ for a general nonuniform transmission line are:

$$V' = -Z(x)I \tag{1}$$

$$I' = -Y(x)V \tag{2}$$

where $Z(x)$ and $Y(x)$ are series impedance and shunt admittance per unit length along x, respectively, and they are arbitrary functions of x. Primes denote derivatives with respect to x throughout this correspondence.

The differentiation of (1) and (2) results in

$$V'' - \frac{Z'}{Z}V' - YZV = 0 \tag{3}$$

$$I'' - \frac{Y'}{Y}I' - YZI = 0. \tag{4}$$

Because (3) and (4) are linear, a few techniques of integral transforms appear powerful; yet, in general, integral transform methods are useful for constant coefficient linear differential equations.

The previous research by many workers used the reflection coefficient defined by:

$$r(x) = \frac{V - IK(x)}{V + IK(x)}$$

where

$$K(x) = \sqrt{\frac{Z(x)}{Y(x)}}. \tag{5}$$

Combination of (1), (2), and (5) yields a generalized Riccati's differential equation[3]

$$r' + P_1(x)r + Q_1(x)r^2 = Q_1(x) \tag{6}$$

where

$$P_1(x) = -2\sqrt{Y(x)Z(x)}$$

$$Q_1(x) = -\frac{K'(x)}{2K(x)}.$$

Eq. (6) can be reduced to a linear form by two available transforms,[4] but the resultant linearized differential equation is of variable coefficient second order. Clearly, (5) makes no contribution for the solution of (1) and (2). In order to make any contribution, many workers in the past used approximations of various kinds, such as neglecting the square of the reflection coefficient. This correspondence follows a rather unconventional method; *i.e.*, the solution of the problem is assumed and the interrelation among parameters is sought.

The interrelation between arbitrary functions $P(x)$ and $Q(x)$ such that

$$n^2 = nP(x)\cot(nx + b) + Q(x) \tag{7}$$

and

$$y(x) = A\sin(nx + b), \tag{8}$$

where A, n, b are constants, is compatible with a linear second-order homogeneous differential equation:[5]

$$y'' + P(x)y' + Q(x)y = 0. \tag{9}$$

Eq. (7) itself presents a special nonuniform transmission line so that $Y(x)$ is defined by

$$n^2Z(x) = -nZ'(x)\cot(nx + b) - Y(x)Z^2(x) \tag{10}$$

where $Z(x)$ is an arbitrary function of x. The extension of (8) is

$$y(x) = Ae^{-u(x)}\sin(nx + w(x) + b) \tag{11}$$

where A and b are constants and $u(x)$ and $w(x)$ are to be determined in terms of line parameters $Z(x)$ and $Y(x)$.

When (11) is substituted into (9), the result is

$$Q - Pu' + u'^2 - u'' - (n + w')^2 + ((P - 2u') \cdot (n + w') + w'')\cot(nx + w + b) = 0. \tag{12}$$

Only two special cases of interest are discussed separately below.

Case 1: u is Constant (Constant Amplitude)

Eq. (12) becomes

$$Q - (n + w')^2 + (P(n + w') + w'') \cdot \cot(nx + w + b) = 0. \tag{13}$$

If $Q = (n + w')^2$,

$$w(x) = \pm\int\sqrt{Q}\,dx - nx + C_1 \tag{14}$$

where C_1 is the constant of integration.

This yields

$$\frac{Z'}{Z} = \frac{Y'}{Y}, \tag{15}$$

and for arbitrary constant A and C,

$$V(x) = A\sin\left(\int\sqrt{-Y(x)Z(x)}\,dx + C\right). \tag{16}$$

* Received by the IRE, December 28, 1959; revised manuscript received, January 8, 1960.

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[5] H. T. H. Piaggio, "An Elementary Treatise on Differential Equations and Their Applications," G. Bell and Sons, Ltd., London, Eng., p. 199; 1920.

Case 2: w is Constant (Constant Phase)

Eq. (12) is reduced to

$$Q - Pu' + u'^2 - u'' - n^2 + n(P - 2u') \cot (nx + w + b) = 0. \quad (17)$$

If $u' = P/2$, n must be obtained from

$$4Q(x) - 2P'(x) - P^2(x) = 4n^2 \quad (18)$$

where n is constant. If the left side of (18) is negative, hyperbolic sine functions appear in place of trigonometric sine functions. The connecting equation between $Y(x)$ and $Z(x)$ is

$$4YZ^3 = CZ^2 + 2Z''Z - 3Z'^2 \quad (19)$$

where C is any nonzero constant, and $Z(x)$ is arbitrary.

The above analysis is an extention of (8). Other forms of the solution can be assumed; yet, this method lacks generality in the sense that $Y(x)$ must be related to $Z(x)$, while the original equations, (1) and (2), claim two independent arbitrary functions $Z(x)$ and $Y(x)$.

A few nonlinear differential equations other than Riccati's equation are linearized and solved exactly. For instance,

$$\left(\frac{r}{y'}\right)' = -\frac{p}{y} \quad (20)$$

where $r(x)$ and $p(x)$ are arbitrary functions of x.

The exact solution of (20) is

$$y(x) = De^{\int q(x)dx} \quad (21)$$

where

$$q(x) = \frac{r}{\int (r - p)dx + C} \quad (22)$$

and two arbitrary constants C and D arise from integrations.

Eq. (20) seems of interest, as it has two independent arbitrary functions $r(x)$ and $p(x)$ for parameters of nonuniform transmission lines. The remaining unsolved problem is to relate (1) and (2) to (20). The possible undetected solution may lie in using a "quasi"-reflection coefficient [a form similar to (5) but more complicated] to obtain a nonlinear equation other than Riccati's equation, such as (20), and to solve this nonlinear equation by an ingenious transformation.

The prevailing idea is that the power series expansion method will solve any problem. This should be employed with due consideration. Eqs. (3) and (4) may have a solution for which the famous method of Frobenius fails if the solutions are of the form e^{-x}.[6] When all analytical approaches fail to obtain the exact solution, approximate methods by numerical analysis, using analog and digital computers, are left available.

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[6] *Ibid.*, p. 110.

An IF Power Comparator with Large Dynamic Range*

The results of recent studies of FM stronger-signal capture[1-3] suggested the application of the same principles to the development of a power comparator that compares two signals at adjacent frequencies and determines which is the stronger. FM stronger-signal capture may be said to occur when an FM discriminator follows faithfully the stronger of two input signals and ignores the weaker of the two. Baghdady[2] has shown that it is difficult to provide a discriminator that will give effective capture when the two input signals are of almost equal amplitude. The linear range of the discriminator and the bandwidths of its tuned circuits must be exceedingly large. In order to obtain capture of a stronger signal that has only a small advantage in amplitude, it is necessary to increase its amplitude advantage before it arrives at the discriminator. Studies by Granlund[1] indicated that this could be accomplished by successive operations of limiting and filtering.

The problem of designing a power comparator differs from that of FM receiver design in that the approximate frequencies of both signals are known in advance. A summary that shows the application of FM stronger-signal capture theory to the design of a wide-range power comparator appears elsewhere.[4] It is shown that the amplitude ratio (taken as the ratio of the smaller signal to the larger signal) is reduced by a limiter followed by a filter passing only the frequency components at the two original frequencies. Fig. 1 shows the output amplitude ratio of n cascaded ideal limiter-filter stages as a function of the input amplitude ratio. The curves are based on Granlund's[1] tables.

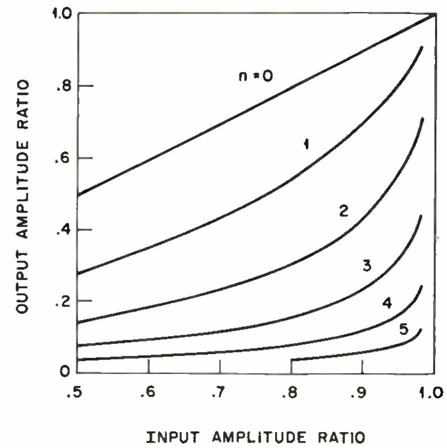

Fig. 1—Output amplitude ratio as a function of input amplitude ratio for n cascaded ideal limiter-filter stages.

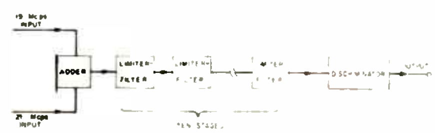

Fig. 2—Block diagram of power comparator.

Fig. 2 is a block diagram of the power comparator that has been built and tested. Input signal frequencies of 19 and 21 mcps are used. Input signals that occur at a different frequency, or frequencies, may be converted to 19 and 21 mcps. The input signals are added and the resulting signal is fed into ten cascaded stages, each consisting of a limiter and a filter. The filters are slightly over-coupled double-tuned circuits having equal peaks at 19 and 21 mcps. The last five limiter-filter stages limit internal noise. Input signals stronger than internal noise cause earlier stages to limit. Stages that are not limiting act only as signal amplifiers. It is essential that they amplify signals at 19 and 21 mcps equally to avoid altering the relative amplitudes of the input signals. The output of the last filter stage goes to a discriminator centered at 20 mcps. The discriminator meets Baghdady's criteria for following variations of instantaneous frequency over a 3-mcps range.[3] The polarity of the discriminator output indicates which of the two input signals is the stronger.

The power comparator was tested with 19 and 21 mcps CW signals produced by two signal generators with calibrated attenuators. The signal generator outputs were added and fed into the comparator. The discriminator output was displayed on an oscilloscope. A dc voltage midway between the voltages corresponding to 19 and 21 mcps was chosen as a reference level. Tests were made to evaluate the system in terms of the following parameters.

Switching Range

The discriminator output consists of a dc component corresponding to the average value of the instantaneous frequency of the discriminator input, and an ac waveform that results from the variations of the instantaneous frequency of the discriminator input. The switching range is defined as the change in the ratio of input amplitudes required to cause the discriminator output waveform to switch from a condition where it is entirely on one side of the reference level to a condition where it is entirely on the other side of the reference level. For input signals with amplitudes differing by half the switching range, the discriminator output waveform is entirely on the side of the reference level corresponding to the stronger signal.

The switching range was measured as a function of input signal level, measured in

* Received by the IRE, October 29, 1959; revised, January 15, 1960. The work reported in this paper was performed by Lincoln Laboratory, a center for research operated by Massachusetts Institute of Technology with the joint support of the U. S. Army, Navy, and Air Force.

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decibels below 11 mw. The accuracy of the measurements is estimated to be ± 0.1 db for values below 1 db. The switching range remains near 0.1 db for inputs larger than -60 db, and increases for smaller inputs. This increase at low levels is caused by the increasing influence of internal noise on the discriminator output waveform at low signal levels. The switching range is less than 2 db for all input signals larger than -90 db. This indicates that, for all input signals above -90 db, a difference in amplitude of 1 db is sufficient to maintain a two-state device, connected to the discriminator output, in one of its two states.

Tracking Error

The dc component of the discriminator output is equal to the reference voltage for a particular ratio of input signal levels. In an ideal power comparator this ratio would be the same at all input signal levels. The tracking error is the variation of this ratio with input signal level. Tracking error results from differences in the gains of limiter-filter stages at 19 and 21 mcps. It is dependent upon the alignment of the limiter-filter stages.

The tracking error is plotted as a function of input level in Fig. 3. The accuracy of the measurements is estimated to be $\pm$ 0.2 db. Curve *a* shows the tracking error when all tuning coils in the double-tuned circuits are peaked at 21.3 mcps. The tracking error is less than ± 1.0 db for the range of input signals between -17 and -97 db. The linearity of the tracking error with signal level for levels below -30 db results from each stage having slightly higher gain at 21 mcps than at 19 mcps. The departure from this linear variation at input signal levels above -20 db is caused by the first limiter stage which, at higher input signal levels, is limiting more heavily than the other stages. The dynamic range may be extended to include larger signals if precautions are taken to avoid overloading the input stage.

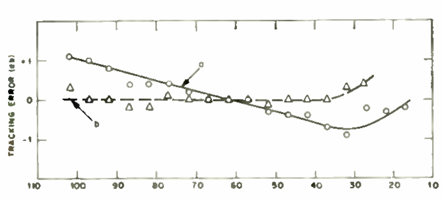

Fig. 3—Tracking error as a function of input signal level for two alignment methods. Curve *a*—all coils peaked at 21.3 mcps; Curve *b*—each stage aligned for equal response at 19 and 21 mcps with the aid of a swept-frequency signal generator.

The tracking error may be reduced as much as desired by accurate alignment. Curve *b* in Fig. 3 shows the tracking error that occurs when each filter stage is aligned for equal response at 19 and 21 mcps with the aid of a swept-frequency signal generator. The error is nearly zero except at high input levels where the first stage overloads.

Capture Range

Capture range is defined as the change in the ratio of the amplitudes of the input signals required to cause the dc component of the discriminator output to move from 10 to 90 per cent of the way from the voltage corresponding to one input signal to the voltage corresponding to the other. Input signals differing by half the capture range produce a dc component at the 10 or at the 90 per cent voltage. The capture range may be used to compare the actual system performance with the theoretical performance predicted for ideal limiters and filters.

Measurements of capture range were taken with the first five limiter-filter stages bypassed. When more than five limiter-filter stages are used, internal noise seriously deteriorates the capture range for small input signals. Fig. 4 shows the experimentally observed capture range as a function of input signal level. The numbered points on the curve indicate where the various stages saturate. At point 5, for example, the fifth stage is limiting strongly. At point 1 all five stages are limiting. At point 0 the input overloads.

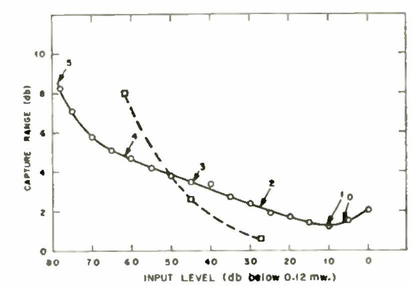

Fig. 4—Capture range as a function of input signal level; solid curve represents experimental data, dashed curve represents theoretical prediction.

According to the theory,[2] it should be sufficient to present to the experimental discriminator an amplitude ratio of less than 0.2 in order that the discriminator output be within 10 per cent of the value corresponding to the frequency of the stronger signal. From Fig. 1 it is found that, with two ideal limiter-filter stages limiting, an input amplitude ratio of 0.63 should be sufficient. With three stages limiting the input amplitude ratio may be as large as 0.86. With four stages limiting it may be 0.97. The corresponding capture ranges are 9, 2.6, and 0.6 db. These values are indicated in Fig. 4 as points on a dashed curve.

The discrepancy between the theoretical and experimental curves is considerable. It may be ascribed to the non-ideal nature of the experimental limiters and filters. Improving the limiters and filters should give closer agreement with the theory. A more attractive expedient, insofar as producing a useful power comparator is concerned, is to cascade more limiter-filter stages. Very satisfactory capture performance is indeed obtained by providing five stages that limit at all times.

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The Possibility of Obtaining Independent Samples from Stationary Gaussian Signals*

When autocorrelation analyses of random signals are made, it is often considered advantageous to use digital rather than analog techniques in order to obtain increased accuracy and signal handling facility. In the mathematical analysis of such systems it is assumed that the successive samples taken from the signal being analyzed are statistically independent. In practice, the correlation between samples usually decreases as the time interval separating them increases, since the autocorrelation function must be integrable. Thus, the degree of dependence between samples can be reduced to any desired level by increasing the separation between samples. However, the conditions that the signals must meet to obtain strict mathematical independence of the samples are thought to be of interest. It is shown below that any random signal from which statistically independent samples can be drawn has a power spectrum satisfying a certain integral equation. Examples are given to show that the existence of statistically independent samples in a random signal is neither required nor insured by its having a power spectrum which is identically zero outside a finite frequency band.

The finite-time autocorrelation function of a stationary Gaussian signal $x(t)$ can be given by

$$R_T(\tau) = \frac{1}{T}\int_0^T x(t)x(t+\tau)dx \qquad (1)$$

where τ is the delay time and T is the correlation time. When a sampling correlator is used, the correlation function is calculated as

$$R_N(\tau) = \frac{1}{N}\sum_{k=1}^{N} x(k\Delta t)x(k\Delta t+\tau) \qquad (2)$$

where Δt is the sampling interval and N is the number of samples used. For equal correlation intervals, so that $T = N\Delta t$, this is an approximation to $R_T(\tau)$ and in some cases may be equal to it. The sample products $x(k\Delta t)x(k\Delta t+\tau)$ are statistically independent only if the autocorrelation function of $x(t)$, $R(\tau)$, has periodic zeros at the sampling intervals, *i.e.*,

$$R(k\Delta t) = 0, \qquad k = 1, 2, 3, \cdots. \qquad (3)$$

For a Gaussian process, the condition that its autocorrelation function, $R(\tau)$, has periodic zeros can be converted into a condition on its power spectrum $G(\omega)$, as follows. By use of the Wiener-Khinchin relationship

$$R(\tau) = \int_0^\infty G(\omega)\cos\omega\tau d\omega \qquad (4)$$

and (3) becomes

$$\int_0^\infty G(\omega)\cos k\Delta t\omega d\omega = 0, \quad k = 1, 2, \cdots. \qquad (5)$$

* Received by the IRE, January 18, 1960. The research in this paper was sponsored by Rome Air Dev. Ctr., Air Res. and Dev. Command, Griffiss Air Force Base, Rome, N. Y., under Contract AF 30 (602)-1655).

This infinite set of conditions on $G(\omega)$ can be reduced to a single condition by defining a new function

$$F(\omega) = \sum_{k=1}^{\infty} a_k \cos k\Delta t\omega \tag{6}$$

where the a_k are any finite constants, real or complex. Then (5) can be written as

$$\int_0^{\infty} G(\omega)F(\omega)d\omega = 0. \tag{7}$$

The $G(\omega)$ demanded by the restriction of periodic zeros on $R(\tau)$ is thus defined as the solution of an integral equation. The nature of the functions $G(\omega)$ that satisfy (7) has not been determined. However, the following examples will show that the restrictions on $G(\omega)$ are not likely to be simple.

The first example will be the bandlimiting condition usually stated: $G(\omega)$ limited to the interval 0 to W cps, and *constant over this interval*. This latter condition is essential as will be seen below. Then if

$$\begin{aligned} G_1(\omega) &= \frac{1}{2W} \qquad 0 \leq |\omega| \leq 2W \\ &= 0 \qquad 2W < |\omega| \end{aligned} \tag{8}$$

the corresponding autocorrelation function is given by

$$R_1(\tau) = \frac{\sin W\tau}{W\tau}. \tag{9}$$

Then, choosing $\Delta t = \pi/W$,

$$R_1(k\Delta t) = R_1\left(k\frac{\pi}{W}\right) = 0,\ k = 1, 2, \cdots. \tag{10}$$

Thus statistically independent samples can be chosen.

On the other hand a random process with a bandlimited power spectrum of the form

$$\begin{aligned} G_2(\omega) &= a \cos a\omega \qquad 0 \leq |\omega| \leq \frac{\pi}{2a} \\ &= 0 \qquad \frac{\pi}{2a} \leq |\omega| \end{aligned} \tag{11}$$

has an autocorrelation function given by

$$R_2(\tau) = \frac{1}{1 - \left(\frac{\tau}{a}\right)^2} \cos \frac{\pi\tau}{2a}. \tag{12}$$

Since it is not possible to find a Δt such that $R_2(k\Delta t)$ is zero for all $k > 1$, statistically independent samples cannot be drawn from a function having this power spectrum.

A random process having a nonbandlimited power spectrum of the form

$$G_3(\omega) = \frac{2}{\omega_0} \frac{1}{1 + \left(\frac{\omega}{\omega_0}\right)^2} \tag{13}$$

has an autocorrelation function

$$R_3(\tau) = e^{-\omega_0|\tau|} \tag{14}$$

which has no zeros for finite values of delay time, τ. Therefore statistically independent samples do not exist.

As another example consider a random process having a nonbandlimited power spectrum of the form

$$\begin{aligned} G_4(\omega) &= \frac{a}{\pi}, \qquad 0 \leq |\omega| \leq \frac{\pi}{a} \\ &= \frac{a}{2^2\pi}, \qquad \frac{\pi}{a} < |\omega| \leq \frac{2\pi}{a} \\ &\ \ \vdots \qquad\qquad \vdots \\ &= \frac{a}{n^2\pi} \qquad \frac{(n-1)\pi}{a} < |\omega| \leq \frac{n\pi}{a}. \\ &\ \ \vdots \qquad\qquad \vdots \end{aligned} \tag{15}$$

This has an autocorrelation function

$$R_4(\tau) = \left[\sum_{n=1}^{\infty} \frac{\cos (2n-1)\frac{\pi\tau}{2a}}{n^2}\right] \frac{\sin \frac{\pi\tau}{2a}}{\frac{\pi\tau}{2a}}. \tag{16}$$

If Δt is chosen equal to $2a$, $R_4(k\Delta t)$ is zero for all $k \geq 1$. Thus statistically independent samples can be chosen from a function having this power spectrum even though it is not bandlimited.

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Transient and Steady-State Behavior in Linear and Nonlinear Systems*

The author of a recent communication concerning the nature of transient and steady-state behavior in linear and nonlinear systems,[1] I feel, has made several conceptual errors. In analyzing the linear system depicted in Fig. 1, the statement is made that the total response, *i.e.*, transient plus steady state, is given by the convolution integral

$$\begin{aligned} g_2(t) &= \int_0^t k(t-\tau)g_1(\tau)d\tau \\ &= \int_0^t k(\tau)g_1(t-\tau)d\tau, \end{aligned} \tag{1}$$

where $g_1(t)$ and $g_2(t)$ are the forcing and response functions, respectively, and $k(t)$ is the impulse response of the system.

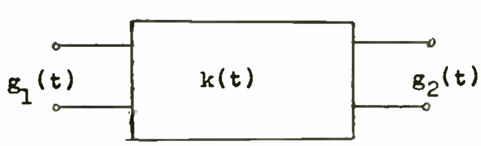

Fig. 1—A linear lumped parameter system.

The statement is not quite accurate for it does not take into account the effect of initial conditions, *i.e.*, the values of

$$g_2(0+)\left(= \lim_{t\to 0_+} g_2(t)\right)$$

and

$$\left.\frac{d^k g_2(t)}{dt^k}\right|_{t=0+}, \quad k = 1, 2, \cdots, n-1; \text{ (n is the order of the system.)}$$

The situation may be analyzed by considering the differential equation of the system of Fig. 1, *i.e.*, the differential equation relating input and output. It is given by

$$\begin{aligned} L_n g_2(t) &= a_n(t)\frac{d^n g_2}{dt^n} + a_{n-1}(t)\frac{d^{n-1}g_2}{dt^{n-1}} \cdots \\ &\quad + a_1(t)\frac{dg_2}{dt} + a_0(t)g_2 \\ = L_m g_1(t) &= b_m(t)\frac{d^m g_1}{dt^m} + b_{m-1}(t)\frac{d^{m-1}g_1}{dt^{m-1}} \cdots \\ &\quad + b_1(t)\frac{dg_1}{dt} + b_0(t)g_1. \end{aligned} \tag{2}$$

This is the general expression relating $g_1(t)$ and $g_2(t)$ for any linear lumped parameter system. For a restricted (though quite important) class of systems, the coefficients, a_i, b_j, are constants. If the coefficients are nonconstant, we assume that they are reasonably well behaved. Note that the Laplace transform of the right side of (2) gives the zeros of the transfer function

$$K(s) = \frac{G_2(s)}{G_1(s)}.$$

and, of course, the left hand side yields the poles.

Considering the homogeneous or "force-free" differential equation,

$$\begin{aligned} a_n(t)\frac{d^n g_2}{dt^n} &+ a_{n-1}(t)\frac{d^{n-1}g_2}{dt^{n-1}} \cdots \\ &+ a_1(t)\frac{dg_2}{dt} + a_0(t)g_2(t) = 0, \end{aligned} \tag{3}$$

we know that it will have a nontrivial solution, *i.e.*, a solution not identically zero, only if at least one of the initial conditions is not zero. These initial conditions indicate the initial state or motion (or energy) of the system and give rise to the *transient* or *force-free* response of the system. Only if *all* initial conditions are zero will there be no transient. When considering initial conditions we must account for discontinuities (occurring at $t=0$) in the forcing function $g_1(t)$ and its derivatives.

By a *fundamental set of solutions*, $\{\phi_n(t)\}$, we mean n linearly independent solutions of (3). Any linear combination of the $\{\phi_n\}$ is again a solution of (3). Thus

$$\Phi_T = c_1\phi_1(t) + c_2\phi_2(t) + \cdots + c_n\phi_n(t)$$

is a solution of (3) and is *by definition* the transient response of the system. The c_i ($i=1, 2, \cdots, n$) are arbitrary constants and reflect the initial conditions. If all initial conditions vanish, then the c_i all become zero and, again, we see that $\Phi_T \equiv 0$.

The total response of the system is given by the complete solution of (2), which is

$$\begin{aligned} g_2(t) &= \sum_{i=1}^{n} c_i\phi_i(t) + \int_0^t H(t;\tau)L_m g_1(\tau)d\tau \\ &= \sum_{i=1}^{n} c_i\phi_i(t) + \int_0^t k(t;\tau)g_1(\tau)d\tau. \end{aligned} \tag{4}$$

$H(t;\tau)$ is the Green's function (or impulse response) associated with the operator L_n of (2), and L_m is the right-side operator of

* Received by the IRE, November 12, 1959.

(2). In the general case, *i.e.*, nonconstant coefficients, we cannot write $k(t;\tau)\equiv k(t-\tau)$, since the differential equation is not invariant to a shift in time origin.

The term containing the integral on the right in (4) is the particular integral, *i.e.*, the solution for a particular $g_1(t)$. *It is the steady-state response.* In general, we can say that the initial conditions of the steady-state response are nil. The advantage of defining steady-state and transient responses in this manner is that this method of definition logically follows from the Laplace transform method of solving the differential equation. Actually, however, this definition of steady-state is somewhat arbitrary, since it specifies that this response and its derivatives be zero at $t=0$.

It is concluded, therefore, that the total response of the linear system is given by (4) not (1), since (1) does not consider arbitrary initial conditions. Eq. (1) simply gives the steady-state response.

The error is compounded somewhat when the integral (1) is split into two parts, one being called the transient and the other the steady-state solution. Thus

$$\int_0^t k(\tau)g_1(t-\tau)d\tau = \int_0^\infty k(\tau)g_1(t-\tau)d\tau - \int_t^\infty k(\tau)g_1(t-\tau)d\tau. \quad (5)$$

The author[1] identifies the first integral on the right as the steady-state response, while the second integral is considered to be the transient. This is incorrect (or at least incomplete) because the first integral reduces to the left side of the equation in view of the fact that $g_1(t-\tau)\equiv 0$ when $\tau>t$, and the second integral is identically zero for precisely this same reason.

Before leaving the analysis of linear systems, it is of interest to note that the impulse response is quite closely related to the transient response—the only question being that of total initial conditions. If the initial conditions arise solely from the discontinuities of the forcing function and its derivatives, *i.e.*, the system is at rest at $t=0-$, then the transient and impulse responses are identical (this is, perhaps, the main advantage to introducing the concept of impulse and singular functions).

We may again define the transient response for nonlinear systems as the force-free response arising as a result of a "state of unrest." Thus, as in the linear case, the transient response serves to carry the arbitrary constants which reflect the initial conditions. For example, when we analyze a triode feedback oscillator using van der Pol's equation, we see that a perfectly good self-sustained transient occurs due to "shock excitation," *i.e.*, as a result of an initial tube current arising from electron excitation. The steady-state response is, of course, the response to the forcing function. The term "complete solution" is essentially meaningless since the sum of the two responses is not a solution of the differential equation. As usual, discontinuities in the forcing function and its derivatives occurring at $t=0$ must be reflected as initial conditions.

[1] A. A. Wolf, "The significance of transients and steady-state behavior in nonlinear systems," PROC. IRE, vol. 47, pp. 1785–1786; October, September, 1959.

It is misleading to partition the response of a nonlinear system into two (or more) parts and associate one with the steady-state and another with the transient response. Determining each response is a problem of its own, and the sum of the two need not be a solution. If, as the author states, the total instantaneous response of any nonlinear system is:

total instantaneous response = steady-state response − transient response + cross-product response,

then one has the (nontrivial) problem of attempting to identify these parts from a known solution remaining consistent with definitions.

Perhaps the most general method of attacking nonlinear equations is via the perturbation and Poincaré-Bendixson theories making use of metric topology.[2]

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[2] E. A. Coddington and N. Levinson, "Theory of Ordinary Differential Equations," McGraw-Hill Book Co., Inc., New York, N. Y.; 1955.

The Behavior of Nonlinear Oscillating Systems in the Presence of Noise*

In view of the recent interest in the behavior of nonlinear oscillating systems in the presence of noise,[1] I should like to call attention to the paper by Rytov[2] on this subject. While several methods are available for analyzing the behavior of oscillators in the presence of noise, it seems that the symbolic differential equation method of Rytov is straightforward and of considerable generality; furthermore, the solution so obtained may be systematically improved to any degree of accuracy. In the present note, this method will be outlined and used to derive some simple results for oscillators with a single tuned circuit.

The general form of the equation of motion for a weakly nonlinear oscillating system with one degree of freedom and with a small random driving force is the following:

$$\frac{d^2x}{dt^2} + x = \mu f\left(x, \frac{dx}{dt}, t, \mu\right) + \mu^2 F(t), \quad (1)$$

where the free oscillation frequency of the system has been normalized to 1; μ is a small parameter; $f(x, dx/dt, t, \mu)$ contains, in general, nonlinear terms; and $F(t)$ is the noise term. The essential idea in the method is the introduction of a second independent time variable τ (slow time)

$$\tau = \mu t \quad (2)$$

and, hence, in the (t, τ) coordinates

$$\frac{d}{dt} = \frac{\partial}{\partial t} + \mu \frac{\partial}{\partial \tau}. \quad (3)$$

* Received by the IRE, January 19, 1960.

[1] J. A. Mullen, "Background noise in nonlinear oscillators," PROC. IRE, to be published.

[2] S. M. Rytov, "Fluctuations in oscillating systems of the Thomson type I and II," *Soviet Phys.*, vol. 2, pp. 217–235; March, 1956; *J. Exper. Theoret. Phys. (USSR)*, vol. 29, pp. 304–333; September, 1955.

In terms of t and τ, the form of the solution of (1) is judiciously chosen as

$$x = R\cos(t-\phi) + \mu\sum_n \{P_n \cos n(t-\phi) + Q_n \sin n(t-\phi)\} \quad (4)$$

where R, P_n, Q_n, and ϕ are functions of τ and μ only. Eq. (4) implies: 1) that x is approximately the solution of a conservative oscillator since R and ϕ are slowly varying in time; 2) that, using (3), the derivatives of any function in the (t, τ) coordinates are of the same order of magnitude, and hence, an iterative procedure (in the small parameter μ) can be used directly on (1); 3) that the zeroth-order solution of (1) is a simple harmonic oscillation of frequency 1, the first-order solution is the van der Pol solution[3] of a nonlinear oscillator in the absence of noise, and the second-order solution is the Berstein solution[4] of an oscillating system in the presence of fluctuation. Since these solutions have been independently established (with much less economy of effort), the validity and utility of the idea is reasonably assured.

With (3) and (4), solution of (1) proceeds by expanding all functions of t in a Fourier series of terms of cos $n(t-\phi)$ and sin $n(t-\phi)$ and all functions of μ in a power series of u. By equating coefficients of like powers of μ of the corresponding Fourier coefficients, the various orders of differential equations of R and ϕ in τ and their fluctuating driving terms can be obtained. These equations will be linear and have constant coefficients; therefore, they can be solved without much trouble. Two results follow immediately from this procedure, insofar as the fundamental component of the oscillator output is concerned: 1) $F(t)$ can be written in the form

$$F(t) = F_{\parallel}(\tau)\cos(t-\phi) - F_{\perp}(\tau)\sin(t-\phi), \quad (5)$$

where $F_{\parallel}(\tau)$, and $F_{\perp}(\tau)$, and $\phi(\tau)$ are functions of slow time, or, in terms of the spectrum of the noise $F(t)$, only input noise (*i.e.*, directly from the impressed noise voltage generator) with its frequency in the neighborhood of the oscillator fundamental frequency can produce FM or AM noise in the oscillator output; 2) however, low-frequency noise can affect the oscillator output only through its effect on the Fourier coefficients of $f(x, dx/dt, t, \mu)$, or, in other words, through parametric excitation (or gain fluctuation). These statements hold also in the case of oscillating systems with more than one degree of freedom such as the klystron.[5]

The present note results from a friendly discussion with Dr. J. A. Mullen while the author was at the Research Division, Raytheon Company.

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[3] See for example, J. J. Stoker, "Nonlinear Vibrations in Mechanical and Electrical Systems," New York University, New York, N. Y.; 1950.

[4] I. Berstein, "On fluctuations in the neighborhood of periodic motions of an auto-oscillating system," *Compt. Rend. (Doklady) Acad. Sci. URSS*, vol. 20, pp. 11–16; 1938.

[5] C. L. Tang, "A Study of Ion Oscillations and Noise in Klystrons," Res. Div., Raytheon Co., Waltham, Mass., Tech. Rept. No. R-51; November, 1959.

Frequency-Temperature-Angle Characteristics of AT- and BT-Type Quartz Oscillators in an Extended Temperature Range*

The frequency-temperature-angle characteristics of an AT- or BT-cut quartz oscillator operating in the temperature range $-50°$ to 80°C are usually described by a power series of the third order in the temperature. For special applications, low temperatures, *e.g.*, that of liquid nitrogen (-196°C) and high temperatures up to 300°C are required. In an extended temperature range, $-200°$ to 250°C, consideration has to be given to whether the third order approximation is satisfactory or higher order terms in the power series have to be considered. Frequency measurements for AT-cut quartz oscillators in the temperature range from $-215°$ to 250°C have recently become available.[1] Fig. 1 (solid line) shows the measured temperatures T_μ (μ = min, max) of zero temperature coefficient of frequency vs orientation angle θ for AT-type oscillators made from natural quartz taken from Phelps.[1]

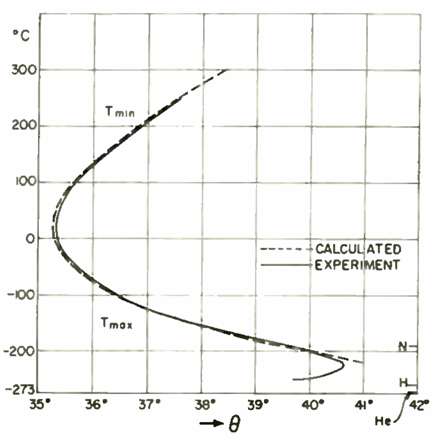

Fig. 1—Observed and calculated temperature of zero temperature coefficient of frequency vs orientation angle θ for AT-cut.

When f is the frequency at an arbitrary temperature and f_0, that of the temperature T_0 (see Bechmann[2]),

$$a_0(\theta) = \frac{1}{f_0}\left(\frac{\partial f}{\partial T}\right)_\theta, \quad b_0(\theta) = \frac{1}{2f_0}\left(\frac{\partial^2 f}{\partial T^2}\right)_\theta,$$
$$c_0(\theta) = \frac{1}{6f_0}\left(\frac{\partial^3 f}{\partial T^3}\right)_\theta, \tag{1}$$

are the temperature coefficients of the first, second, and third order related to the temperature T_0 and to the orientation angle θ_0, the maximum and minimum frequencies, using the third-order equation, occurs at temperatures

$$T_\mu - T_0 = \frac{1}{3c_0}(-b_0 \pm \sqrt{b_0^2 - 3a_0c_0}).^3 \tag{2}$$

The corresponding frequency deviations are given by

$$\left(\frac{\Delta f}{f}\right)_\mu = \frac{\pm 2\sqrt{b_0^2 - 3a_0c_0}^3 + 2b_0^3 - 9a_0b_0c_0}{27c_0^2}. \tag{3}$$

The measured values shown in Fig. 1 have been used to determine a new set of the three temperature coefficients of frequency and their derivatives with respect to the orientation angle θ_0. The new values are:

AT-cut, $\theta_0 = 35°15'$, $T_0 = 20$°C.

$$\left.\begin{aligned} a_0 &= 0, \\ \frac{\partial a_0}{\partial \theta} &= -5.15\cdot 10^{-6}/°\text{C}, °\theta, \\ b_0 &= 0.39\cdot 10^{-9}/(°\text{C})^2, \\ \frac{\partial b_0}{\partial \theta} &= -4.7\cdot 10^{-9}/(°\text{C})^2, °\theta, \\ c_0 &= 109.5\cdot 10^{-12}/(°\text{C})^3, \\ \frac{\partial c_0}{\partial \theta} &= -2.0\cdot 10^{-12}/(°\text{C})^3, °\theta, \end{aligned}\right\} \tag{4}$$

and are in good agreement with those given in Bechmann.[2] The values T_μ, calculated from (2) and using the values in (4) for the temperature coefficients, are also plotted in Fig. 1 (dotted line) and a very satisfactory agreement in the temperature range from $-200°$ to 250°C is obtained.

At the temperatures T_μ,

$$\frac{1}{f}\frac{\partial f}{\partial T} = 0,$$

and, in the vicinity of these values, the power series reduces to

$$\frac{1}{f}\frac{\partial f}{\partial T_\mu} = b_\mu(T - T_0),^3 \tag{5}$$

where

$$b_\mu = \sqrt{b^2 - 3ac}.$$

The value b_μ as function of the orientation angle θ calculated with the values in (4) is shown in Fig. 2. With increasing the orientation angle θ the range of a given frequency deviation decreases.

It can be seen that the third order expression for the frequency-temperature-angle characteristic satisfactorily describes the behavior of an AT-cut within the limits of $-200°$ to 250°C. Beyond this range higher order terms must be considered.

Similarly, frequency measurements of BT-quartz oscillators in an extended temperature range lead to the temperature coefficients:

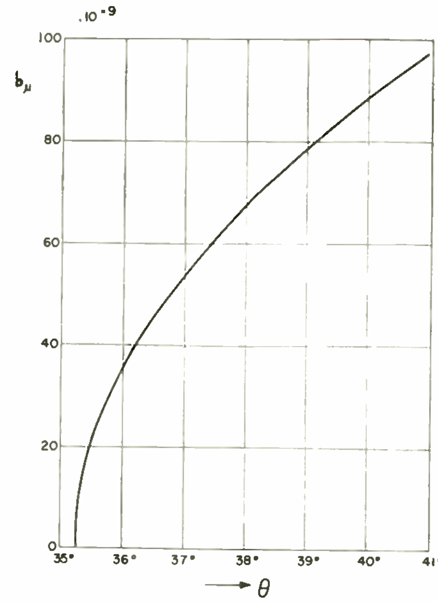

Fig. 2—Calculated values of the parabola constant b_μ as function of the orientation angle θ corresponding to the zero temperature coefficient curve of Fig. 1.

BT-cut, $\theta_0 = -49°12'$, $T_0 = 25$°C.

$$\left.\begin{aligned} a_0 &= 0, \\ \frac{\partial a_0}{\partial \theta} &= 1.8\cdot 10^{-6}/°\text{C}, °\theta, \\ b_0 &= -40\cdot 10^{-9}/(°\text{C})^2, \\ \frac{\partial b_0}{\partial \theta} &= 2.0\cdot 10^{-9}/(°\text{C})^2, °\theta, \\ c_0 &= -128\cdot 10^{-12}/(°\text{C})^3, \\ \frac{\partial c_0}{\partial \theta} &= 10\cdot 10^{-12}/(°\text{C})^3, °\theta, \end{aligned}\right\} \tag{6}$$

The value $\partial c_0/\partial\theta = 38\cdot 10^{-12}$ for the BT-cut previously given[2] is too high.

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* Received by the IRE, January 28, 1960.

[1] F. P. Phelps, "Stability of quartz resonators at very low temperatures," *Proc. Eleventh Annual Symp. on Frequency Control*, Fort Monmouth, N. J., pp. 256–276; May, 1957.
Also, F. P. Phelps, R. D. Goodwin, and A. H. Morgan, "Investigation of Stability of Quartz Resonators at Low Temperatures," National Bureau of Standards, Boulder, Colo., Second Quarterly Rept., U. S. Army Signal Corps Contract No. R-56-0034-SC-91; July 1 to September 30, 1956.

[2] R. Bechmann, "Frequency-temperature angle characteristics of AT-type resonators made of natural and synthetic quartz," Proc. IRE, vol. 44, pp. 1600–1607; November, 1956.

[3] In the reference to Bechmann, *ibid.*, in (5), the root should read: $\sqrt{b_0^2-3a_0c_0}$ instead of $\sqrt{b_0^2+3a_0c_0}$.

Electromagnetic Theory from a Mathematical Viewpoint*

The introduction of Maxwell's equations to students is always a fascinating subject. This little note gives a possibly not novel but certainly seldom used alternate. It started with the author wondering about physics and physical laws in general. Why should certain equations like Maxwell's equations be so successful and so unshakable that they have withstood nearly a century of tremendous revolutions in physics? One is told that they are based on a few experimental facts, the few experimental facts available at the time of their conception. Such general

* Received by the IRE, February 1, 1960.

understanding of the physical world with so few basic facts to start from seems to be bordering on the unlikely. Is it possible that the universality of such theories is not due to their author's genial feelings for the physical world but to a mathematical necessity which makes those theories nearly the only ones logically available? Is it possible to disregard completely the experimental world and still come out with only few alternatives in the laws of nature? In this case one may only truly wonder at the docility of the experimental world to human logic. Of course, logically necessary but as yet unrevealed relations may be found through the gifted perusal of experimental facts. It should however be the hard way, the way which demands genius and luck.

In this frame of mind, the author tried to derive Maxwell's equations from analytical geometry only. To his amazement, wh ch may be due to a lack of knowledge, he was successful.

To introduce Maxwell's equations one starts with coordinates and entities represented by functions of the coordinates. When the coordinate system is projective, the reality of the physical entities is determined by their invariance to transformations of coordinates. If we do not introduce tensors, the only invariants are scalars and vectors.

A physical law is represented by a mathematical relation between the function representing the physical entities. It may be assumed that before trying very complex relations, one wishes to try the very simple ones. Thus, one may restrict oneself to linear relations and use more complex ones only if the experimental world makes it necessary. The use of only linear relations permits one to use only first order operators. One can always introduce as many new physical entities as necessary to replace linear equations containing higher order operators by sets of equations containing only first order operators. To equate the results of an operation on a scalar or vector to a scalar or vector, the result of the operation must be a scalar or vector. There are only two first-order operators in analytical geometry which have this property: the gradient of a scalar and the divergence of a vector. There is an exception if and only if the number of coordinates is three. Then the curl of a vector, which is in general a tensor of the second order, reduces into a vector. After having exhausted the relations possible with functions of coordinates only, one can introduce new variables which do not transform in a projective manner. In analytical geometry they are parameters. The only first-order operator involving a parameter is the first-order partial derivative in that parameter. We are now ready to investigate all different possibilities provided we also assume that two entities or more which only and always appear in the same linear sum form only one entity represented by that linear sum.

The scheme now is to assume ever increasing numbers of scalars and numbers of vectors and to write the different relations possible between them. With two scalars and no parameters, for instance, the only relation is that a linear sum of their gradients must vanish which reduces the assumptions to one gradient free scalar. Two vectors, no parameter, and a number of coordinates different from three reduces to one divergence-free vector. The first interesting case is made of one scalar and one vector, in this case either:

$$aE + b\nabla U = 0 \tag{1}$$

or

$$a\nabla \cdot E + bU = 0 \tag{2}$$

with E a vector, U a scalar, a, b universal constants. The first one is trivial and relates a potential to a field. The second one however is Newton's field equation with E the field due to gravity and U the mass density. With two vectors and three coordinates, we get two divergence free vectors which can be related by two linearly independent relations of the form:

$$a\nabla \times E + b\nabla \times H + cE + dH = 0. \tag{3}$$

E and H are the vectors. The constants a, b, c, d are universal constants. If c and d are assumed dimensionless, then a and b are universal lengths. If universal lengths are not assumed possible, then (3) yields curl-free vectors. If, in this case, one introduces the parameter time, the vectors are still divergence free but (3) becomes:

$$a\nabla \times E + b\nabla \times H + cE + dH + e\dot{E} + f\dot{H} = 0 \tag{4}$$

(the dot for time derivative). Assuming e and f dimensionless, c and d are universal frequencies and a and b are universal velocities. If one restricts (4) to one universal constant and chooses a universal velocity one obtains the set of homogeneous Maxwell's equations in free space.

The next step is to introduce different media. Each medium is separated from its neighbor by a surface or boundary. The form of the homogeneous set of Maxwell's equations is such that, if one assumes the same laws with the same vectors E and H in all media, one can deduce from the equations themselves that E and H will be continuous at each boundary and that the solution will not recognize that the media exist. To introduce media we must thus assume one or more entities which are in some way different from medium to medium. We can, for instance, assume that the free space E is continued in a different medium in two different ways as follows.

If S is zero in free space and one within a given medium, one can write two vectors:

$$E = E(1 - S) + ES \tag{5}$$

and another vector:

$$D = E(1 - S) + DS. \tag{6}$$

One can, of course, do or not do the same for H. If one now introduces the new vectors in (7) and still assumes only one universal velocity, one obtains the homogeneous set of Maxwell's equations for current and space charge free media. It should be noted that boundary conditions do not have to be stated separately and are included automatically through the differentiation of S.

The introduction of a scalar in the preceding vector world can only arise through the modification of a divergence equation. Thus one can, for instance, introduce ρ by:

$$\nabla \cdot D = \rho. \tag{7}$$

This introduction forces a modification of one of the curl equations and, in order to keep within first-order operators, also the introduction of a vector i defined by:

$$\nabla \cdot i + \dot{\rho} = 0. \tag{8}$$

Along the way we have made very simple choices and discarded other simple possibilities. One may wonder if those simple possibilities should not be investigated and possibly used to explain some new phenomena.

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Poisson, Shannon, and the Radio Amateur*

The article by Dr. Costas[1] concerning the relative merits of DSB and SSB present a rather bleak picture of the capabilities of the armed forces in planning and controlling their usage of the frequency spectrum. While the radio amateur may shift frequency to avoid interference at will, it is certainly not true that military operators have the same privileges. Mere contemplation of such an undisciplined system would turn the hair of a Division Signal Officer gray instantly.

Actually, Army radio operators are trained to work through interference and jamming, and in no case to stop trying to work the circuit or to be led off the assigned frequency. Frequency changes are made only on order of the net control station (NCS), and are authorized by the appropriate frequency allocation board or officer. Military radio nets not strictly controlled in frequency and procedure are, of course, easy prey for enemy electronic warfare. Considering tactical radio nets, the picture is further improved by noting that distance ranges are inversely proportional to the amount of radio equipment used. Thus, while there are an extremely large number of Radio Sets AN/PRC-6[2] in the infantry division, their range is only about 1–3 miles.

The author compares the circuit capacity for narrow-band and broad-band systems and notes that when the average transmitting duty cycle factor is low, 10 per cent or less in his example, the broad-band system becomes superior. In the military situation, a low duty cycle factor occurs in a static situation, during buildup prior to offensive operation, or otherwise, when radio operation is required to be a minimum. In these cases, the commander does not particularly care what his channel capacity is, since the majority of his traffic is handled by wire and messenger. However, when he really needs the capacity, as in an offensive or break-out operation, the commander demands (and had better get) maximum capacity. Thus the military system should not be designed like

* Received by the IRE, December 28, 1959.
[1] J. P. Costas, "Poisson, Shannon, and the radio amateur," PROC. IRE, vol. 47, pp. 2058–2068; December, 1959.
[2] FM "Handy-Talkie" radio set, 47 to 55.4 mc in 43 channels spaced 200 kc apart.

the civilian telephone system, which will break down at a certain overload point, but must handle the most severe traffic conditions.

In connection with jamming, it is true that narrow-band systems are more easily jammed. Since our own and other armies will use the same portions of the spectrum for similar types of operation, it then becomes necessary for the enemy to employ a large number of spot jammers, rather than use barrage jamming, if he is not to disrupt his own communications. If he is willing to do this at all, he will not be hesitant about raising his power level enough to jam whatever DSB systems we may have. Jamming will not be nearly as effective against a well-disciplined military radio net as it would be, for example, against radio amateurs.

In the strategic-type military communications systems, such as ACAN (Army Command and Administrative Network), frequency control is even more rigid, since consideration must be given to possible effects on systems of our and other armed services and other governments. The total available bandwidth is limited by propagation characteristics, and high-capacity, high-traffic reliable channels are desired. Here again, narrow-band techniques would seem to be preferable.

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Author's Comment[3]

I find that there are two main difficulties involved in answering Mr. Early's letter. First, it is quite obvious that his knowledge of Army communications is much more detailed than mine. The second difficulty I face is of an historical nature. In past conflicts we have found ourselves to be the predominant users of frequency spectrum due to our superiority in either technical know-how or mass-production capability, or both, with respect to our opponents. Thus, seldom in the past have we been seriously bothered by massive amounts of either unintentional enemy interference or jamming. It is not surprising, then, that self-generated interference has turned out to be a most serious communications problem and that operational methods, such as net operation under strict frequency discipline mentioned by Early, were developed to alleviate the difficulty. But times change and so do the problems which confront us. No longer can we expect to achieve an overwhelming technological superiority over possible opponents, and with regard to the frequency spectrum, we could well find ourselves to be an equal or even less than equal partner in its future usage. One must also consider the distinct future possibility of a sizeable opponent capability with respect to jamming. Thus the situation changes and our problems and their solutions change accordingly. This historical situation presents a problem to me since it is understandably difficult to excite any large-scale interest in problems which have not yet been experienced but which have only a certain probability of existing in the future. It seems only normal that we tend to concentrate on solutions to old problems in the belief, I suppose, that the dominant problems of the past will also be the dominant problems of the future.

With regard to Early's first paragraph, we can certainly agree that our armed forces are capable of planning and controlling their usage of the frequency spectrum. It should be kept in mind, however, that *their usage* of the spectrum is *all* that they can plan and control, since the spectrum itself is not the exclusive property of either side. In a military environment the only spectrum usage you will *get* is that which you can *take*, not that which is given to you. This seems simple enough, yet acceptance of this premise dictates, in my opinion, a radical departure from prior art techniques in military communications systems planning.

Turning now to Early's second paragraph, we must admit that if Army radio operators are trained to work through interference and jamming, then there is no problem. Perhaps a more accurate statement would be that training certainly helps but only to a degree. Since Early admits that narrow-band systems (systems which have a low ratio of transmission bandwidth to data rate) are easily jammed, he must also concede that narrow-band circuits or nets could be disabled by enemy countermeasures in spite of operator skill. It then becomes inconceivable to me that the net members would remain on the useless frequency making persistent but futile attempts to clear traffic. I would suspect, rather, that an alternate frequency would be available and known to all net members and that a transfer would be made to that frequency by the net. When the jammer catches up with them, the net again must move to a second alternate frequency, and so forth. Note that in the above example (if accurate) narrow-band systems are used, but in order to obtain some degree of military capability these systems are employed in such a way as to cause more spectrum space to be used than is "absolutely necessary." (In this case we have a rather primitive form of frequency jumping.) This process of initially designing equipment to occupy the narrowest possible bandwidth and then using the equipment operationally in such a manner as to spread the signal spectrum occurs quite often and has always puzzled me. Many people of my acquaintance are horrified at the idea of buying any system which uses, say, twice the bandwidth of system X, yet when confronted with the problem of jamming they cheerfully point out that system X can switch sidebands or jump to another channel. I have always wondered why, if the other sideband or the other channels are available, there is concern over bandwidth in the first place? Or, if system X must be used, per circuit, on many different frequencies in order to survive in the expected environment, how can a bandwidth saving be claimed for this system?

Early, in paragraphs 2 and 5, seems to relate frequency rigidity with operational reliability. I do not understand this, for it seems to me that mobility is the key to survival for communications circuits as well as for military units in modern combat operations. This "mobility" of communications circuits can be obtained in many different ways, some better than others, but this vital ingredient of survival can never be possessed by a system which stays in one channel and which has a low ratio of transmission bandwidth to data rate (*i.e.*, a bandwidth-conserving system).

In paragraph 3, Early brings up a truly serious problem. Since I have shown in my paper that the average per-circuit capacity varies inversely as band loading, this means that commanders on both sides will have maximum per-circuit capacity when they need it least (during periods of relative military inactivity) and minimum per-circuit capacity when they need it most (during a breakout and its containment, for example). I admit that I have not come up with pleasant results but maybe the man who said "war is hell" was right. I have tried to point out in the paper that capacities will be limited and that commanders must be aware of this and plan their procedures accordingly. I have further tried to show that narrow-band systems may at first appear to give high traffic capacities, and their use may be entirely satisfactory in exercises conducted under carefully controlled conditions. This may be quite deceptive, however, and I think it only wise to try to estimate as accurately as possible the capacities which will actually be available in practice. The results in the paper for the average per-circuit capacity during periods of high communications activity may be discouragingly low, but I think this gives a more accurate picture of the true situation than may be obtained from reading the data-rate specification on the name plate of a piece of narrow-band communications equipment.

The comparison drawn by Early between military systems and civilian telephone systems is somewhat unfortunate since there is contained the implication in his statement that the telephone systems break down under overload due to some shortcoming in design while the military systems keep working by virtue of superior planning. This certainly is not the case and I am sure Early intended no such comparison. The point is simply that there is a certain total practical capacity available in the frequency spectrum which, like it or not, must be shared by *all* users. One can demand more capacity for one's own use but getting it is another matter.

In raising the question of the relative jamming immunity of SSB and DSB systems, Early points out that an enemy would no doubt be willing to expend the extra effort required to spot-jam whatever DSB circuits we may have. This is certainly true, but such an argument ignores the fundamentals of the problem. To begin with there is no such thing as a system which is absolutely immune to jamming. If an enemy has unlimited resources and is determined, he can jam any system. The trick then is to design the communications system in such a way that an unreasonably large effort is required on the part of the jammer. This is really a battle of resources; one attempts to make the ratio of jammer cost to communications

[3] Received by the IRE, January 22, 1960; revised manuscript received, February 1, 1960.

equipment cost so high as to discourage the potential jammer. The basic DSB system shows some anti-jam advantage over SSB but there is no dramatic difference in the two systems. After all, how can there be any great difference since DSB uses only twice the bandwidth of SSB? In a hostile environment bandwidth is power; conserve one and you must expend more of the other. DSB as a modulation technique represents only a first step in the right direction. By using DSB in place of SSB you cut your cost and you raise the jammer's cost. Now, as a next step, couple the DSB modulation technique with certain baseband processing tricks so that really wide transmission bandwidths (in relation to data rate) result, and then you get anti-jam capabilities of real significance.

In spite of what has just been said, I do not wish to minimize the advantage gained by even a two-to-one increase in transmission bandwidth. In a recent series of on-the-air tests, DSB showed a significant and consistent advantage over SSB for voice transmission over an HF path. Part of this advantage is derived from speech clipping (which raises average power for a fixed peak power limitation) and speech pre-emphasis—techniques that can easily be applied to DSB but are not as directly acceptable in SSB. Another important DSB advantage became evident when unintentional interference appeared on the channel. The word scores in DSB remained high because the DSB receiver could meet band conditions by choosing one of three modes of reception: upper, lower, or double-sideband. This same interference, during many runs, caused an almost complete drop-out when SSB was used, due to the fact that the transmissions were "blind" with no communications available from the receiver back to the transmitter. Except for the fact that these test results were not reported, they represented at least a partial vindication of my position that DSB is superior to SSB for general-purpose military use. This is a battle that I have apparently lost and I do not mean to bore Early by bringing up dead issues. I only wish to point out that even a two-to-one increase in transmission bandwidth, if properly executed, can yield very handsome operational rewards.

The subject of world-wide strategic networks such as ACAN is a topic on which I am not qualified to speak. All I know about these systems is what I read in the trade press and there have been so many proposals and so many alpha-numeric designations for different systems that I have long since given up trying to keep track. I can say this much about such systems in general: If they are of the "high-capacity" variety, meaning that the ratio of transmission bandwidth in cycles per second to the data rate in bits per second is low (say near one or two), then these systems must have in the order of a +10- to +20-db signal-to-interference ratio in order to operate. If the signal-to-interference ratio gets down to, say, the neighborhood of 0 db, these systems simply die. Thus, "high-capacity" systems need a controlled, nonmilitary type of environment in which to operate. I would not class such systems as military systems but I would say, rather, that they are basically commercial systems in use by the military. For such strategic applications Early says narrow-band techniques are preferable. I can only agree that their use is convenient because of the diplomatic situation involved with frequency allocation. On the other hand, if one must request a clear channel in order to obtain proper operation for a particular system, then that system is not a military system in the true sense.

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Letter from Mr. Early[4]

In reply to Dr. Costas, I believe the Army is very much aware of the problem of jamming. Current doctrine[5] requires all radio operators to "observe radio discipline at all times—when jammed, keep calm, keep trying, keep operating" among other AJ techniques. We can safely assume that at least the friendly radio system will be well-disciplined, and we have quite a few AJ tricks available. Why, then, buy a broadband technique which will be swamped by our own use during peak traffic periods, without any help from the enemy?

Charles D. Early, Jr.

[4] Received by the IRE, January 21, 1960.
[5] Department of the Army, "Defense Against Electronic Jamming," FM-11-151; June, 1955.

Three-Port Ring Circulators*

A compact three-port ring circulator using a total nonreciprocal phase shift of 180 degrees has been developed. Circulators that use a transmission line in the form of a ring do not require hybrids, can be synthesized in three ways, are easily analyzed, and can be built in waveguide, coaxial line, or any other type of transmission line in which nonreciprocal devices can be constructed. Vartanian[1] has proposed a symmetrical ring circulator using three nonreciprocal 180-degree phase shifters. A similar device is described by Kock.[2]

The phase-shift parameters for the ring circulator are easily determined by referring to Fig. 1; the circulator ports are numbered from 1 to 3. We define θ as the phase shift between the adjacent ports that are indicated by the subscript. Direction of power flow is indicated by the sequence of numbers in the subscript—that is, θ_{23} is the phase shift of a signal traveling by the most direct path (in this case, counterclockwise) from port 2 to port 3.

* Received by the IRE, January 29, 1960.
[1] P. N. Vartanian, "Theory and Applications of Ferrites at Microwave Frequencies," Sylvania Electronic Defense Lab., Mountain View, Calif., Rept. E15, pp. 119–126; April, 1956. Obtainable from ASTIA as Rept. No. AD 101888.
[2] W. E. Kock, "Signal Routing Apparatus," U. S. Patent No. 2,794,172; May, 1957.

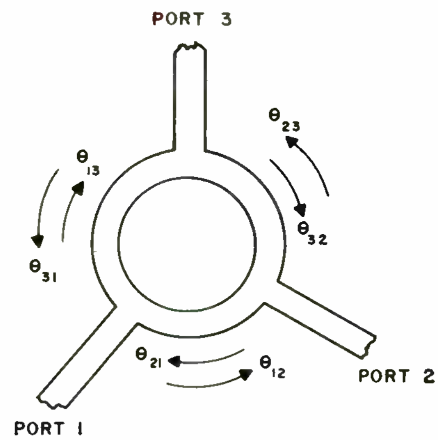

Fig. 1—Three-port ring circulator.

Assuming excitation of the circulator at port 1:

a) to transmit to port 2, $\theta_{12}=\theta_{13}+\theta_{32}+2m\pi$;
b) to isolate port 3, $\theta_{13}=\theta_{12}+\theta_{23}+(2n+1)\pi$;

where m and $n=0, \pm 1, \pm 2, \cdots$.

Similar expressions are written for excitation at ports 2 and 3. The circulator is assumed lossless. There are three solutions for $m=0$.

I. Three Nonreciprocal Phase Shifters

This is the symmetrical case, that is, all clockwise, and counterclockwise, phase shifts are equal. Thus, $\theta_{12}=\theta_{23}=\theta_{31}$ and $\theta_{13}=\theta_{32}=\theta_{21}$. The solution is $\theta_{cw}=(2n+1)\pi/3$ for the clockwise phase shift between ports, and $\theta_{ccw}=(2n+1)2\pi/3$ for the counterclockwise phase shift between ports. Thus, for $n=0$, $\theta_{cw}=60$ degrees and $\theta_{ccw}=120$ degrees [see Fig. 2(a)]. The circulator becomes more frequency sensitive as n increases, since the line lengths increase, and the circulator action depends upon phase addition and cancellation for its operation. On the other hand, a lower limit is placed upon n, since the ferrite phase shifter, which has a nonreciprocal phase shift $\Delta\theta=\theta_{ccw}-\theta_{cw}$, requires a finite line length for physical realization.

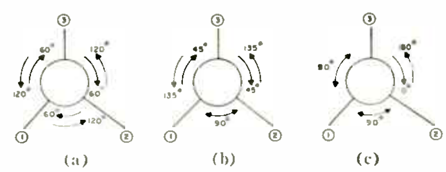

Fig. 2—Phase shifts required to synthesize three types of ring circulators. (a) Type I circulator, symmetrical structure requiring three nonreciprocal phase shifters. (b) Type II circulator, requiring two nonreciprocal phase shifters. (c) Type III circulator, requiring one nonreciprocal phase shifter.

II. Two Nonreciprocal Phase Shifters

Let $\theta_{12}=\theta_{21}$, $\theta_{13}=\theta_{32}$, and $\theta_{23}=\theta_{31}$. The solution is $\theta_{12}=\theta_{21}=(2n+1)\pi/2$, $\theta_{13}=\theta_{32}=(2n+1)\pi/4$, and $\theta_{23}=\theta_{31}=(2n+1)3\pi/4$. Solution for $n=0$ results in two nonreciprocal phase shifters of $\Delta\theta=90$ degrees each [see Fig 2(b)].

III. One Nonreciprocal Phase Shifter

Here, the solution is $\theta_{12}=\theta_{21}=\theta_{13}=\theta_{31}=(2n+1)\pi/2$, $\theta_{32}=n\pi$, and $\theta_{23}=(2n+1)\pi$. The solution for $n=0$ results in one nonreciprocal phase shifter of $\Delta\theta=180$ degrees [see Fig. 2(c)].

In all three cases, the total nonreciprocal phase shift is 180 degrees. This is in agreement with Carlin's conclusion[3] that the minimum number of gyrators necessary for circulator action is equal to half the rank of the circulator impedance matrix. For a three-port circulator, the rank is 2. Therefore, we are using the minimum nonreciprocal phase shift required for physical realization.

Fig. 3 shows a symmetrical ring circulator of Type I that has been constructed for operation at X-band. The unit uses 100-ohm strip transmission line matched into Type N coaxial connectors. Dielectric phase-shifters (not shown) are used to optimize the reciprocal phase-shift. The nonreciprocal phase shifters use rectangular General Ceramics R-1 ferrite bars, in Rexolite mounting brackets, located adjacent to the center-conductor of the strip transmission line (see Fig. 4). Nonreciprocity is obtained by the TEM-mode-distorting mechanism described in Duncan, *et al.*[4] To obtain high nonreciprocity at X-band, no additional dielectric, besides that provided by the ferrite itself, was required.

Test results at 9300 mc for the Type I ring circulator shown in Fig. 3 were: a forward loss of 0.4 db, an isolation of 23 db, a SWR of 1.16, and a bandwidth of 50 mc. It is believed that performance can be improved with additional effort.

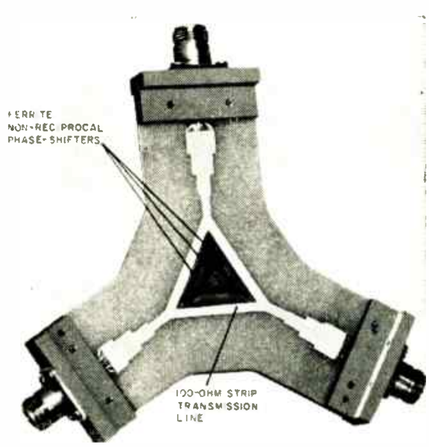

Fig. 3—Ring circulator (top ground plane removed).

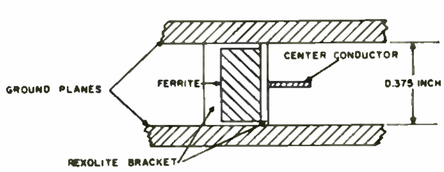

Fig. 4—Configuration of nonreciprocal phase shifter.

The realizability of the three-port ring circulator has been checked using network analysis. The characteristic impedance of the ring required for matched inputs was computed. It was found to equal the characteristic impedance of the input lines, since power does not flow in the portion of the ring containing the isolated port. One can cascade three-port circulators into a ring as proposed by Auld.[5] Using the Type III circulator shown in Fig. 2, it is possible to construct an n-port cascaded-ring circulator similar to Fig. 11 in Auld, but using a common ferrite ring or post (see Fig. 5).

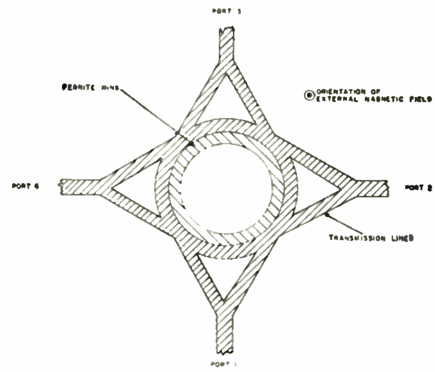

Fig. 5—Four-port cascaded-ring circulator.

The authors wish to thank S. Okwit and W. D. White of AIL for valuable discussions.

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[3] J. J. Carlin, "Principles of Gyrator Networks," *Proc. Symposium Modern Advances in Microwave Techniques*, Polytechnic Institute of Brooklyn, Brooklyn, N. Y., pp. 175–204; November, 1954.

[4] B. J. Duncan, L. Swern, K. Tomiyasu, and J. Hannwacker, "Design Considerations for Broad-Band Ferrite Coaxial Line Isolators," Proc. IRE, vol. 45, pp. 483–489; April, 1957.

[5] B. A. Auld, "Synthesis of Symmetrical Waveguide Circulators," IRE Trans. on Microwave Theory and Techniques, vol. MTT-7, pp. 238–246; April, 1959.

Models of the Atmospheric Radio Refractive Index*

In the above article,[1] Bean and Thayer studied the correlation between the gradient of the refractive index and the refractive index at ground level. By using a large number of readings of U. S. origin, the authors showed that a good representation of the variations of the refractive index with altitude is obtained with an expression of the form

$$N(h) = N_0 \exp ah \quad \text{(with the usual notation).}$$

the value of the coefficient a varying with the climate.

It is well known that for radio communication beyond the horizon, the variations of the gradient of the refractive index are correlated with the variations of electric field. It is tempting, when the gradient of the index is unknown, to use the value N_0 as a radio-climatic parameter. This has led one of the authors to write several articles on the subject. Unfortunately, this method is not one which can be generalized and it can only be used for a limited part of the world. This can easily be shown.

If s is the specific humidity (measured in grams of water-vapor per kilogram of air), we can write, using the usual units,

$$N = 77.6 \frac{P}{T}\left(1 + 7.7 \frac{s}{T}\right).$$

For an average value, the quantity P/T varies exponentially with altitude irrespective of the climatic zone chosen; the chief reason for the generalization of this law is that the relative changes in T are small. Hence,

$$N = K_1\left(1 + 7.7 \frac{s}{T}\right) \exp K_2 h,$$

K_1 and K_2 being constants.

In order to be able to write $N = N_0 \exp ah$, we must have one of the following two conditions fulfilled:

$$1 + 7.7 \frac{s}{T} = \text{constant} \tag{1}$$

$$1 + 7.7 \frac{s}{T} = K_3 \exp K_2 h. \tag{2}$$

and either one suffices. We will now look at each separately.

Condition (1) can be satisfied in two different ways:

a) $7.7s/T \ll 1$ which gives $s \leq 10$ g/kg approx.

This condition for s means the temperature is a maximum and therefore, for an average value, the altitude is a minimum. In order to satisfy this condition from sea level upwards, the ground temperature must be less than 14°C and this condition suffices. Figs. 1 and 2 show the zones for which this hypothesis is admissible.

b) $s/T = c^t$.

For the first few kilometers above ground-level, T decreases by about 10 per cent for an average value. [For higher altitudes $s < 10$ g/kg, and we have case a).] s then must decrease by about 10 per cent. This decrease is very small compared with the possible changes in s. Hence, we can say that s is virtually constant. In any case, measurement accuracy is of the order of 10 per cent for this parameter.

This almost constant value does not exist outside zones subject to strong atmospheric turbulence between ground level and the region where $s = 10$ g/kg. Such conditions only exist during daytime above overheated territories such as the Sahara, the Arizona desert, etc.

Condition (2) is never rigorously satisfied, because regions where $7.7\ s/T \gg 1$ simply do not exist. However, in order to satisfy the requirements of experimental physics, we can consider an exponential decrease of s/T as being sufficient. This elimi-

* Received by the IRE, November 16, 1959.

[1] B. R. Bean and G. D. Thayer, "Models of the atmospheric radio refractive index," Proc. IRE, vol. 47, pp. 740–755; May, 1959.

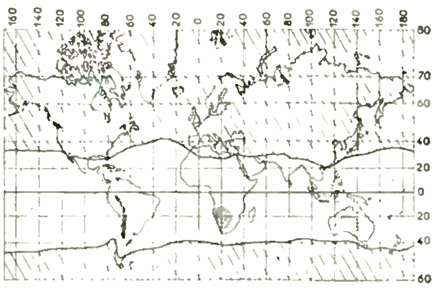

Fig. 1—January ▧ $N(h)$ exponential, day and night. ▨ $N(h)$ exponential, daytime.

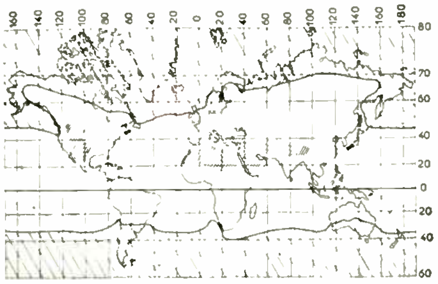

Fig. 2—July ▧ $N(h)$ exponential, day and night. ▨ $N(h)$ exponential, daytime.

nates all the zones in which the diffusion of water vapor is prevented by a stratum of subsidence at almost constant altitude, by a region subject to systematic nocturnal cooling, by the existence of a monsoon climate, etc. In fact, only certain temperate regions can be considered as being in this category.

To sum up, apart from the cases shown in Figs. 1 and 2, there may be accidental correlations between N_0 and ΔN, but they cannot be generalized. Hence, the exceptions cited by Bean and Thayer, namely California and the Dakar region, are not special cases but simply examples from regions where their theory is not applicable, and many similar cases can be found in Australia.[2] One of the results of the geographical limitation of the exponential law for N is that it makes impossible the "transfer to sea level of N_s" for a large part of the earth's surface.[3]

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Authors' Comment[4]

As we interpret Misme's letter, he reaches the following conclusions concerning three separate problems:

1) Only in those regions where the exponential distribution is likely may the refractive index gradient be obtained from the surface value of the index, N_s.
2) As a consequence of 1) he expects the variations of beyond-the-horizon radio fields to be correlated with N_s only for the regions of exponential profiles.
3) He would also expect these regions of "exponentiality" to delineate properly the regions wherein one may reduce the surface value of the index to a sea-level value.

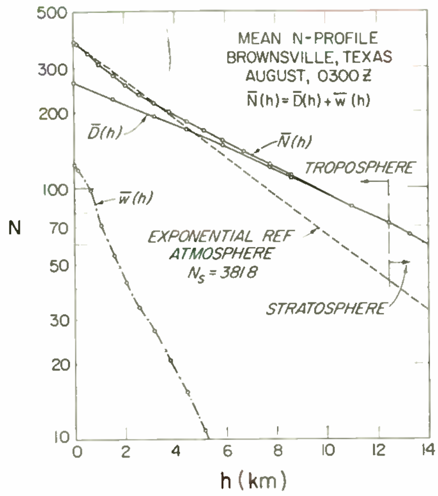

Fig. 1—Typical variation with height of the air density and water vapor components of the refractive index for a warm and humid climate (Brownsville, Texas; August).

We agree that our model atmospheres will fit more precisely in regions of uniform low atmospheric water vapor content. We also agree that even on the average the atmosphere is never exactly an exponential function of height. We do not believe Misme's conclusions as to the regions of "nonexponentiality" to be justified since, to arrive at these conclusions, he must apparently assume:

1) the air is always saturated with water vapor;
2) the essentially arbitrary value of 10 gm/kg for the specific humidity at the earth's surface separates regions where the refractive index does and does not decrease exponentially;
3) the water vapor component of the refractive index must decrease with height according to certain preconceived principles.

Since Misme's criterion depends solely upon the surface value of the specific humidity, it is not necessarily a relevant measure of the height dependence of N. Misme has overlooked the fact that we have examined in detail the *actual* height variation of the refractive index and that our model atmospheres are based upon the world's largest and most extensive collection of refractive index profiles. Indeed, to agree with Misme's regions of "nonexponentiality" would require us to abandon the data.

We can, however, test both Misme's hypothesis and our model by comparison with independent data from regions outside of our original data sample and inside Misme's regions of $s \geq 10$ g/kg where he presumes the atmosphere is not exponential.

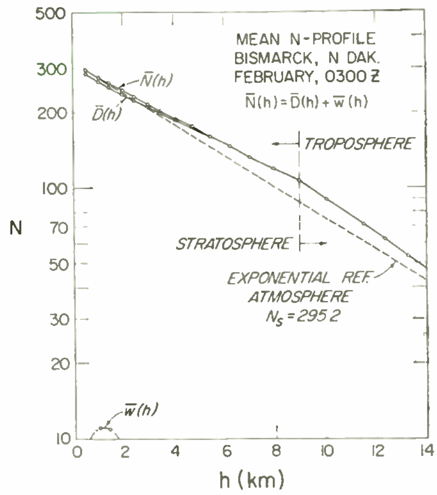

Fig. 2—Typical variation with height of the air density and water vapor components of the refractive index for a cold and dry climate (Bismarck, N. D.; February).

These data are presented in terms of his three conclusions given above.

Refractive Index Height Structure

What we are really concerned with is not Misme's surface conditions but rather the height variation of N and, more particularly, as Misme has pointed out, the two height variations of the water vapor and air density components of the refractive index.

One may write

$$N = (n - 1)10^6 = D + W \tag{1}$$

where with the usual units

$$D = 77.6\frac{P}{T} \quad \text{and} \quad W = \frac{3.73 \times 10^5 e}{T^2}.$$

Examples of the average variation of N, D, and W with height are given for warm, humid conditions on Fig. 1 and for cold, dry conditions on Fig. 2. It is seen immediately that D decreases very nearly exponentially with height from the surface to the tropopause. Above the tropopause the refractive index decreases more rapidly due to the isothermal nature of the stratosphere, but still in an exponential fashion. The water vapor component, W, is observed to be always less than D and to decrease much more rapidly with height. It appears from an examination of these figures, and many similar charts for other locations in the U. S., that, whenever it is large, W also tends to decrease exponentially with height. Assuming this to be the case we may use as our model

$$N(h) \cong D_0 \exp\left\{-\frac{h}{a}\right\} + W_0 \exp\left\{-\frac{h}{c}\right\}$$
$$\cong N_s \exp\left\{-\frac{h}{b}\right\} \tag{2}$$

where $a > b > c$. A rough estimate of the average values of a, b and c yields $a \simeq 9$ km, $c \simeq 3$ km and $b \simeq 7$ km. Misme concludes that an exponential decrease of both the air density and the water vapor components is a sufficient condition for an exponential decrease of N with height; the data of Figs. 1 and 2 indicate that this sufficient condition

[2] W. S. Ament, "Airborne radiometeorological research," Proc. IRE, vol. 47, pp. 756–761; May, 1959.

[3] See "Influences radioclimatiques sur les liasons transhorizon," *Onde Élect.*, vol. 399, pp. 116–123; January, 1960.

[4] Received by the IRE, February 1, 1960.

is satisfied in nature. We have determined the constant b in (2) as a function of N_s by fitting this model to the data only over the first kilometer and, in this way, have obtained a good fit to the data over the first few kilometers. It should be recalled in this connection that most of the influence of refractive index structure upon radio wave propagation occurs in these first few kilometers.

The general trend of the data of Figs. 1 and 2 is towards that of an exponential distribution with height. The data for the winter months, Fig. 2, are in agreement with Misme's hypothesis that we expect an exponential height distribution for that time. However, the data for Brownsville, where $s = 18.2$ gm/kg, also appear to be essentially an exponential function of height in the important lower atmosphere. Gerson's[5] mean refractive index profiles for tropical maritime air and monsoon air masses of India also tend toward an exponential height distribution even though they represent regions and meteorological conditions for which Misme concludes an exponential decrease of N with height would not exist.

Independent data have recently become available which allow a comparison of our model with N profiles from regions much different than the U. S. These consist of 5 years of individual radiosonde observations. Figs. 3 and 4 show mean profiles prepared from data for Canton Island and Isachsen, N. W. T., Canada. These two stations represent tropical and arctic conditions, respectively, and should represent a good test of our U. S. derived atmospheres.

Canton Island, in the South Pacific Ocean, practically on the equator (02° 46″S, 171°42″W), has a year-round value of $s \sim 20$ gm/kg and thus does not meet Misme's criteria for a climate with an exponential N atmosphere. Even in this extreme case N tends to decrease exponentially with height and our model exponential atmosphere predicted solely from N_s lies within one standard deviation of the mean N profile to an altitude of 4.5 kilometers indicating essential agreement over the critical lower portion of the atmosphere. Similar agreement was obtained for Isachsen, N. W. T., which represents the extreme opposite of climatic conditions from Canton Island, and despite the intense radiation inversion characteristic of the long arctic night, the model lies within one standard deviation of the average N value from the surface to 3 kilometers.

Although our model was developed using data only from the continental United States we have subsequently not found any regions where it has not proved to be a reasonable representation of the actual N structure in the first few kilometers. For example, summer time data from Florida and the states adjoining the Gulf of Mexico are fully comparable with that of Canton Island while the northern prairies in winter yield N profiles similar to that of Isachsen. Work is of course continuing on the development of better model atmospheres. These future models will probably be developed along the lines of the bi-exponential representation of N as mentioned above.

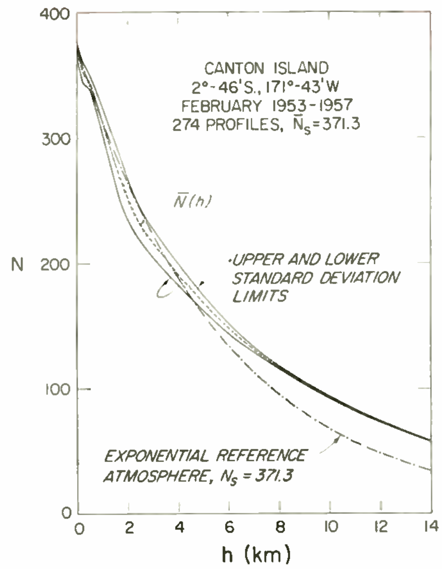

Fig. 3—Five year mean N profiles for Canton Island compared with the exponential reference atmosphere predicted from N_s alone.

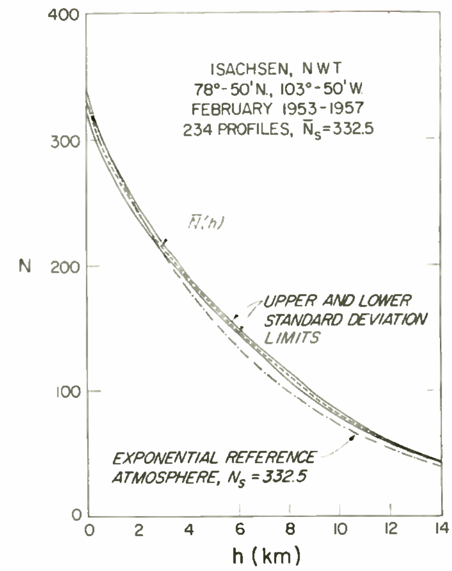

Fig. 4—Five year mean N profiles for Isachsen, N. W. T., compared with the exponential reference atmosphere predicted from N_s alone.

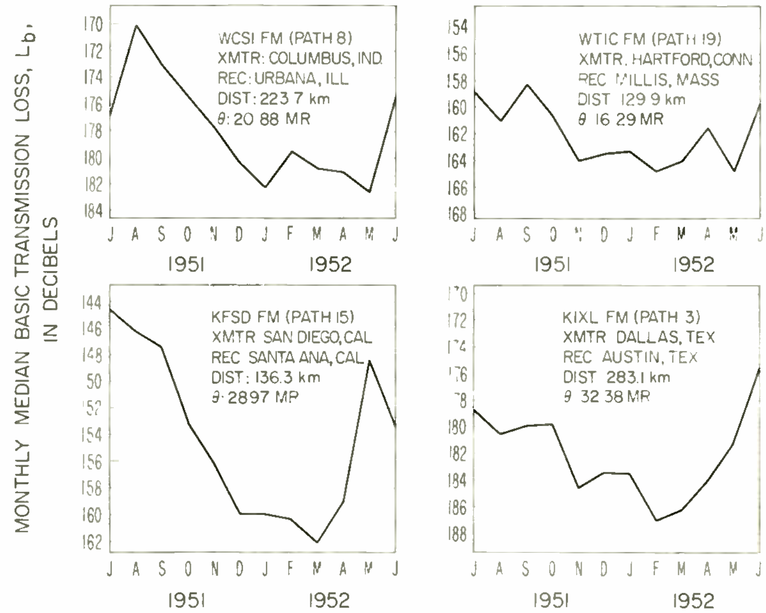

Fig. 5—Annual cycles of monthly median basic transmission loss for four climatically different radio paths.

Correlation of Radio Field Strengths and N_s

Misme's hypothesis that $s < 10$ g/kg is a necessary condition for a good correlation between N_s and beyond-the-horizon radio fields appears to be contradicted by a growing body of reported correlations to the contrary for regions where $s > 10$ g/km. Correlations of the order of 0.7 to 0.9 between the annual cycles of N_s and electric field strengths have been reported for areas that lie within the region of $s > 10$ g/km for at least part of the year. Such correlations have been reported for the east coast of the United States,[6] for the Gulf region of Texas,[7] for the Florida-Caribbean area,[8] for Argentina to Uruguay,[8] for several paths in the Mediterranean Sea,[8,9] and from Tokyo to

[5] N. C. Gerson, "Variations in the index of refraction of the atmosphere," *Geofis. pura e Appl.*, vol. 13, pp. 88–101; March–April, 1948.

[6] G. W. Pickard and H. T. Stetson, "Comparison of tropospheric reception at 44.1 mc with 92.1 mc over the 167-mile path of Alpine, New Jersey, to Needham, Massachusetts," Proc. IRE, vol. 38, p. 1450; December, 1950.

[7] B. R. Bean, "Some meteorological effects on scattered radio waves," IRE Trans. on Communications Systems, vol. CS-4, pp. 32–38; March, 1956.

[8] R. E. Gray, "The refractive index of the atmosphere as a factor in tropospheric propagation far beyond the horizon," 1957 IRE National Convention Record, pt. 1, pp. 3–11.

[9] L. Bonavoglia, "Correlazione fra fenomeni meteorologico e propagazione oltre l'orizzonte sul Mediterraneo," *Alta Frequenza*, vol. 27, pp. 815–824; December, 1958.

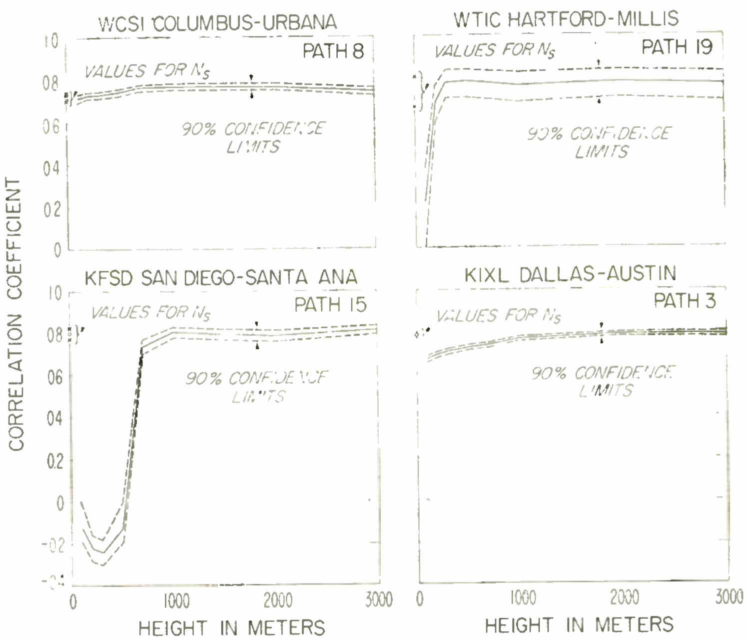

Fig. 6—Comparison of correlation coefficients of transmission loss with N_s and with the gradients determined by the difference between N_s and the value of N at the indicated heights.

TABLE I

Deviation of Estimated 8-Year Means of N_s (Calculated from Contour Charts) from Actual 8-Year Means for 20 Randomly Distributed U. S. Weather Bureau Stations

Test Station	August 0200			
	Height	Actual 8-yr mean value of N_s	Deviation*	
			N_s map	N_0 map
	Meters	N units	N units	N units
Sacramento, Calif.	7	329.6	1.0	−1.8
Portland, Ore.	8	337.7	19.7	6.0
San Diego, Calif.	11	348.1	16.1	3.5
Mobile, Ala.	66	376.0	6.0	0.6
Fresno, Calif.	86	326.2	5.2	4.2
Boston, Mass.	89	347.5	− 7.5	−0.4
Grand Rapids, Mich.	210	340.5	− 1.5	0.1
Columbia, Mo.	239	348.7	− 2.3	−2.5
Minneapolis, Minn.	255	338.5	− 0.5	2.7
Cincinnati, Ohio	271	344.1	− 2.9	−2.8
Des Moines, Iowa	294	343.1	− 1.9	−0.1
Pendleton, Ore.	455	300.9	2.9	−3.1
Billings, Mont.	1088	285.6	5.6	1.2
Burns, Ore.	1262	271.3	−15.7	−4.4
Salt Lake City, Utah	1288	279.5	8.5	4.4
Reno, Nev.	1340	277.6	−29.4	1.6
Pocatello, Idaho	1355	269.7	− 3.3	0.0
Denver, Colo.	1625	276.6	− 1.4	0.3
Colorado Springs, Colo.	1882	272.4	− 6.6	1.1
Flagstaff, Ariz.	2131	261.4	−36.6	−2.2
Root-mean-square deviation			13.0	2.7

* Deviation equals the actual long-term mean minus the value obtained from map contours.

Osaka in Japan.[10] It is unfortunate that data on the variation of radio fields are not available to compare with du Castel and Misme's published meteorological data for the Dakar region.[11] A long-term radio-meteorological experiment in the Dakar region would certainly yield valuable information as to coastal modification of tropical air and the effect of these modifications upon radio fields.

A further critical test of Misme's hypothesis was made by comparing the annual cycle of radio fields recorded along the Southern California coast with the annual cycles of both N_s and a variety of N gradients. We note from Fig. 5 that the annual cycle of transmission loss in this region follows the same general pattern as those recorded in other parts of the United States. This is quite remarkable considering the great differences in climate represented by these four examples. It is even more remarkable (see Fig. 6) that N_s yields as good a correlation for this path as any of the usual measures of N gradient. The fact that this conclusion is the same for all four radio paths despite their dissimilar climatic zones is also significant.

[10] M. Onoe, M. Hirai, and S. Niwa, "Results of experiment of long-distance overland propagation of ultra-short waves," *J. Radio Res. Labs. (Japan)*, vol. 5, pp. 79–94; April, 1958.

[11] F. du Castel and P. Misme, "Elements de radio climatologie," *Onde Elec.*, vol. 37, pp. 1049–1052; November, 1957.

Mapping of Sea-Level Values of N from N_s

This problem is quite different from the preceding ones. The basic argument here is whether more accurate estimates of N_s can be obtained from maps in which the systematic dependence of N_s on station elevation has been removed than can be obtained from simple maps of the surface value N_s. This elevation dependence of N_s was removed in the past[12] by assuming that the fictitious sea-level value, N_0, is related to the station value by

$$N_0 = N_s \exp (ah_s)$$

where h_s is the altitude of the weather station above sea level. This exponential form is a natural consequence of the equation for the molecular refractivity of air.[13] The practical consequences of the use of N_0 is that the fascinating departures of N from average values, due to climatic differences and weather storm systems, are emphasized.[12,14,15]

A further critical test of Misme's hypothesis was carried out by preparing charts of N_0 and N_s for summer nights in the U. S. when $s>10$ g/kg. These charts were prepared from only 42 of the 62 U. S. Weather Bureau stations for which 8-year means of N_s are available. The remaining 20 stations, selected by Misme during the course of our CCIR work, were used as a test sample by estimating their 8-year mean values of N_s from separate N_0 and N_s contours. The individual deviations of the values obtained from the two contour maps are listed in Table I. By comparing the root-mean-square (rms) deviations of 13.0 N units obtained by estimating N_s from the N_s contours with the 2.7 N-unit rms obtained by estimating N_s from N_0 contours, one concludes that it is 5 times more accurate to estimate N_s from contours of N_0 than from those of N_s itself. We must also conclude that it is not only possible to map N_0 but that it is quite profitable to do so.

In conclusion, the above experimental evidence indicates that it is reasonable to assume an exponential model for practical applications. It is gratifying that our simple exponential model works as well as it does, especially in such apparently anomalous regions as Southern California, the Arctic, and the equatorial zone. Extensive tables of refraction effects for the CRPL Exponential Reference Atmosphere profiles are now available.[16]

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[12] B. R. Bean and J. D. Horn, "The radio refractive index near the ground," *J. Res. NBS*, vol. 63D, pp. 259–271; November–December, 1959.

[13] B. R. Bean and R. M. Gallet, "Applications of the molecular refractivity in radio meteorology," *J. Geophys. Res.*, vol. 64, pp. 1439–1444; October, 1959.

[14] B. R. Bean and L. P. Riggs, "Synoptic variation of the radio refractive index," *J. Res. NBS*, vol. 63D, pp. 91–97; July–August, 1959.

[15] B. R. Bean, L. P. Riggs, and J. D. Horn, "Synoptic study of the vertical distribution of the radio refractive index," *J. Res. NBS*, vol. 63D, pp. 249–254; September–October, 1959.

[16] B. R. Bean and G. D. Thayer, "CRPL Exponential Reference Atmosphere," Natl. Bur. of Standards, Monograph No. 4. For sale by the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. ($0.45).

Effects of Resistance in Avalanche Transistor Pulse Circuits*

In using avalanche transistors to generate fast-rising high-current pulses,[1] one generally encounters a degradation of performance due to the presence of resistance in the circuit. A typical pulse circuit is shown in Fig. 1, where the resistance R includes the device lead and contact resistance, the bulk resistance of the semiconductor material, and any resistance inserted for the purpose of sampling the current. The purpose of this note is to obtain a quantitative estimate of the effects of R. Intuitively one expects that the resistance will cause a degradation of both peak current and 10 to 90 per cent rise time. The analysis confirms the intuitive expectation: for the case considered, the presence of 50 ohms resistance caused a reduction in peak current of about 30 per cent and an increase in 10 to 90 per cent rise time of about 20 per cent.

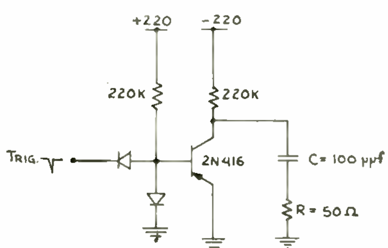

Fig. 1—Typical avalanche transistor pulse circuit.

It is to be emphasized that the goal of the analysis is to obtain an *estimate* of the effects of R. In the course of the analysis it will be convenient to borrow heavily from the results for the case for which $R=0$.[2] For that case, the circuit model and the device model, with charge variables, are shown in Figs. 2 and 3. The assumptions, approximations, and results are given in Tables I and II. We shall make use of an additional important result: for a fixed base width, the peak current always occurs at the same voltage, V_p, and the stored charge at that voltage is linearly distributed, regardless of the configuration of the external circuit. Qualitatively, the effect of R is to decrease the amount of charge stored as the current rises to its peak, thus decreasing the value of the peak current.

For the charge model of the transistor, shown in Fig. 3, it is clear that the incremental stored charge is given by

$$dQ_S = (1 - 1/M)dQ \qquad (1)$$

regardless of the external circuit configuration. When the external circuit consists of a

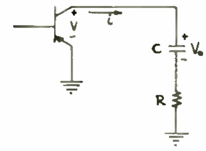

Fig. 2—Model for the avalanche transistor pulse circuit.

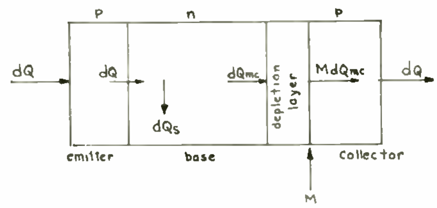

Fig. 3—Charge model for the avalanche transistor.

series RC as in Fig. 2, the collector-to-emitter voltage V is given by

$$V = V_0 + iR + (1/C)\int dQ. \qquad (2)$$

Designating the incremental stored charge for this case to be dQ_{SR} and solving (2) for dQ and substituting in (1), one obtains

$$dQ_{SR} = Cdv(1 - 1/M) - RCdi(1 - 1/M). \ (3)$$

A gross approximation is now made to facilitate the analysis: the voltages across the resistor and the capacitor are assumed small enough so that the multiplication remains essentially at its initial value until peak current is reached. This approximation is valid only for small values of R and initial values of M near unity. The approximation permits the use of

$$1 - 1/M = (V_0/V_B)^n$$

in (3), which then becomes

$$dQ_{SR} = CdV(1 - 1/M) - RCdi(V_0/V_B)^n.$$

It will be noted that the first term of the right member of (4) is the incremental stored charge for the case $R=0$; (4) demonstrates that R reduces the stored charge. Since the peak current occurs at the same voltage, V_p, for $R=0$ and for $R\neq 0$, (4) may be integrated as V goes from V_0 to V_p; *i.e.*, as i goes from 0 to I_{pR}, where I_{pR} is the peak current for $R\neq 0$. The result is

$$Q_{SRP} = Q_{SP} - RCI_{pR}(V_0/V_B)^n$$

where Q_{SRP} is the stored charge at peak current for $R\neq 0$. Since the stored charge is linearly distributed at peak current, $I_p = Q_{SRP}/\tau$. Taking the ratio I_{pR}/I_p, one obtains

$$(I_{pR}/I_p) = 1/[1 + (RC/\tau)(V_0/V_B)^n]. \ (5)$$

To obtain the effects of R on the rise time, the approximation is made that the actual 10 to 90 per cent rise times for the two cases bear the same ratio as their estimated maximum rise times. The correlation between calculated and experimental results

TABLE I

Approximations and Assumptions for the Case $R=0$

1) Bias currents are negligible in comparison to pulse currents, so that the circuit model for the circuit of Fig. 1 is as shown in Fig. 2.
2) During the pulse, the emitter junction voltage is zero.
3) The emitter efficiency is unity.
4) $C \gg C_c$ (the collector capacitance), and C_c is thus neglected.
5) No recombination occurs in the base region.
6) Avalanche multiplication occurs instantaneously at the metallurgical collector junction. The multiplication factor s given by $M = 1/[1-(V/V_B)^n]$ where V is the collector-to-emitter voltage, and V_B is the breakdown voltage.
7) The nondepleted base width remains fixed.

TABLE II

Important Results for the Case $R=0$

1) The stored charge at peak current is

$$Q_{SP} = \int_{V_0}^{V_p} C(1 - 1/M)dV$$

where V_0 is the initial voltage and V_p is the voltage at peak current.
2) Because the stored charge is linearly distributed, the peak current is $I_p = QSP/\tau$ where $\tau = w^2/2D_{eff}$; w is the nondepleted base width, and D_{eff} is the effective diffusion constant for minority carriers in the base region.
3) The estimated maximum 10 to 90 per cent rise time of the current pulse is T_C est. max. $= 1.6C(V_0 - V_p)/I_p$.

indicates that this is a reasonable approximation. Denoting by T_R the 10 to 90 per cent rise time for the case $R\neq 0$, one obtains

$$\frac{T_R \text{ actual}}{T_C \text{ actual}} = \frac{T_R \text{ est. max.}}{T_C \text{ est. max.}} = \frac{I_p}{I_{pR}}\left[1 - \left(\frac{I_{pR}R}{V_0 - V_p}\right)\right]. \qquad (6)$$

Table III shows calculated and measured results for a 2N416 transistor with $R=50$ ohms and $C=100$ $\mu\mu$f.

TABLE III

Experimental and Calculated Results for a 2N416 Transistor With $R=50$ Ohms and $C=100$ $\mu\mu$F

	Peak current ratio, I_{pR}/I_p	10 to 90 per cent rise-time ratio, T_R/T_C
Calculated	0.62	1.18
Experimental	0.69	1.18

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*Received by the IRE, February 5, 1960. This work was performed at Stanford Electronics Laboratories under Office of Naval Research Contract Nonr 225(24), NR 373 360.

[1] G. B. B. Chaplin, "A method of designing transistor avalanche circuits with application to a sensitive transistor oscilloscope," presented at the IRE-AIEE Transistor and Solid-State Circuits Conference, Philadelphia, Pa.; February 20–21, 1958.

[2] D. J. Hamilton, J. Gibbons, and W. Shockley, "Physical principles of avalanche transistor pulse circuits," Proc. IRE, vol. 47, pp. 1102–1109; June, 1959.

Isolator Effect on Cascaded Reflex Klystron Amplifiers*

Introduction

It has been known that a reflex klystron 2K25 can be used as a regenerative amplifier.[1-3] The gain and sensitivity of a receiving system can be increased if two or more reflex klystron amplifiers are cascaded. One of the simplest ways is cascading the amplifiers with ferrite isolators; however, the isolation of most practical isolators is imperfect. Most practical isolators have some reflections. These slight reflections and leakage powers have strong influence upon the over-all performance of cascaded reflex klystron amplifiers.

Principle of Isolator Application

When the reflex klystron amplifiers are cascaded by a coupling circuit which contains a ferrite isolator, the amplifier performance is influenced by the degree of isolation. If the isolation is perfect, then no feedback is possible. The amplifier will be stable, but high gain due to feedback cannot be expected. If the isolation is poor, then the feedback between stages will be too great and the amplifier system will tend to oscillate and become unstable. Because, as shown in Fig. 1, the reflex klystron 2K25 amplifier used in this experiment is a one-port amplifier, a considerable portion of the amplified power reflects back to the input waveguide. Therefore, the degree of isolation must be chosen so that the amplifier system can have stable, high-gain operation. Several isolators must be cascaded in order to increase the isolation.

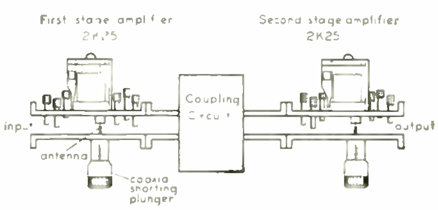

Fig. 1—Schematic diagram of cascaded 2K25 reflex klystron amplifier.

When several isolators are cascaded, the isolation is increased but, at the same time, the phase shift of the over-all isolator system will be changed. Thus it is desirable to use the isolators together with a variable phase shifter to obtain proper phase shift. If the isolator is used with a variable phase shifter, the position of the isolator in the coupling circuit makes a difference in the amplifier performance. When the isolator is arranged as shown in Fig. 2(a), the first-stage amplifier is directly influenced by residual reflections from the phase shifter. In this arrangement, the second stage is not strongly influenced by the residual reflections from the phase shifter since the source side impedance of the second stage is kept almost constant by the isolator. On the other hand, changing the setting of the phase shifter changes the load impedance of the first stage.

On the contrary, when the isolator is as shown in Fig. 2(b), the first-stage amplifier is protected against the spurious reflections from the variable phase shifter. The first-stage amplifier must amplify small input signals and, at times, the magnitude of the spurious reflections would be comparable to the input signals. Therefore, it is important to protect the first stage by the isolator. In this arrangement, however, the second stage is not protected against the spurious reflections.

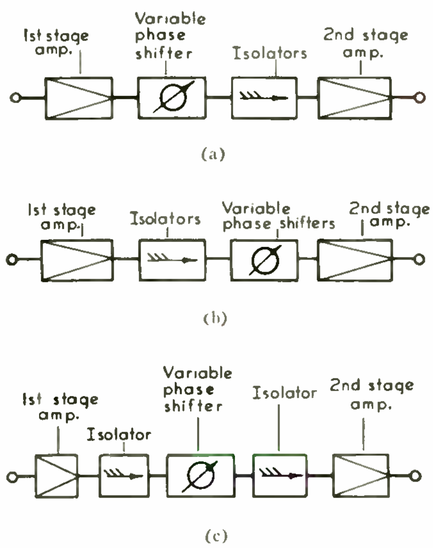

Fig. 2—Various kinds of cascaded reflex klystron amplifiers. (a) Phase shifter-isolator coupling. (b) Isolator-phase shifter coupling. (c) Isolator-phase shifter-isolator coupling.

It is therefore necessary to arrange two isolators as shown in Fig. 2(c), in order to protect both stages.

Hence, an isolator serves two functions when used in the coupling network: 1) to obtain proper isolation, and 2) to obtain the proper impedance for stable high-gain amplification.

Experimental Results

Various kinds of cascading methods were investigated to study the isolator effects on the cascaded 2K25 reflex klystron amplifier performance.

The input vs output characteristics of various kinds of isolator coupled amplifiers are shown in Fig. 3. For reference, the characteristics of a direct-coupled amplifier are also shown in Fig. 3. When only one isolator was used, the gain and linearity improvements were not significant in comparison with the direct-coupled amplifier. When two isolators were used, the increased isolation made the system stable and made the adjustment of the individual amplifier for high gain easier.

When the isolators were used together with a phase shifter, similar effects were observed. The experimental results are shown in Fig. 4. These experiments were done in a frequency range of 9359–9365 mc.

Phase characteristics of cascaded reflex klystron amplifiers are shown in Fig. 5. In these various kinds of coupling arrangements, the coupling sequences of individual components varied, but one thing common

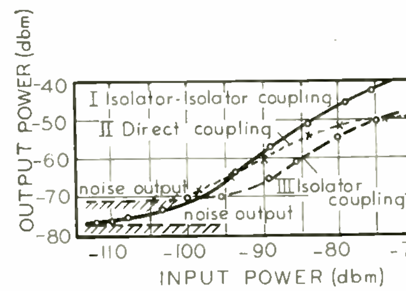

Fig. 3—Input vs output characteristics of isolator-coupled amplifiers.

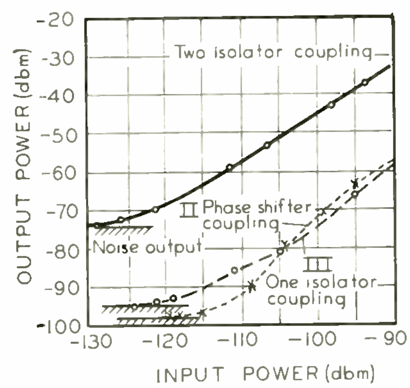

Fig. 4—Input vs output characteristics of various kinds of isolator-phase shifter-coupled amplifiers.

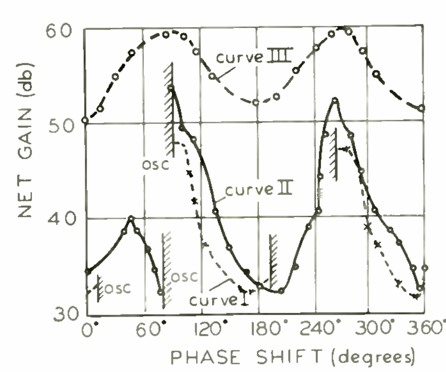

Fig. 5—Phase characteristics of cascaded reflex klystron amplifiers. Curve I, –*––*–, phase shifter-isolator-isolator coupling. Curve II, —o—o—, isolator-isolator-phase shifter coupling. Curve III, – – o – – o – –, isolator-phase shifter-isolator coupling.

* Received by the IRE, January 21, 1960; revised manuscript received, February 4, 1950. A part of this work was performed while the author was at the Dept. of E. E., Univ. of Wisconsin, Madison.

[1] K. Ishii, "X-band receiving amplifier," *Electronics*, vol. 28, pp. 202–210; April, 1955.

[2] K. Ishii, "One-way circuit by the use of a hybrid T for the reflex klystron amplifier," Proc. IRE, vol. 45, p. 687; May, 1957.

[3] K. Ishii, "Noise Figure of Internal Cavity Reflex Klystron Amplifier," Master's thesis, University of Wisconsin, Madison; 1957.

in these three kinds of arrangements was that one phase shifter and two isolators were employed in the coupling circuit.

When the isolators were used after the phase shifter, the first-stage amplifier, which had to amplify small input signals, was not protected against spurious reflections. The gain was relatively low and there were wide oscillation regions in phase shifter setting (see curve I). On the other hand, when the isolators were used in front of the phase shifter, the amplifier performance was improved as shown by curve II in Fig. 3. Finally, when the isolators were used in both sides of the phase shifter, the circuit adjustment of the amplifiers for high gain was easier. Thus, high gain and smooth phase characteristics were obtained as shown by curve III. The isolators showed very strong influence on the directivity of the amplifier. The directivity was defined as a ratio of the forward to the backward power from the amplifier. The directivity depends on the gains of the individual stages, the impedance ratio of the input and output circuit of each stage, and the transmission characteristics of the cascading circuit.

The employment of the isolators showed some effect on stability of the cascaded reflex klystron amplifier. In this case, the stability was expressed in terms of average-gain fluctuation over a 10-minute period after several hours of long-time stability test.

When the two reflex klystron amplifiers were coupled together directly without using the isolators, the gain fluctuation was 0.08 db over a 10-minute period. But when one isolator was used for coupling, the gain fluctuation became 0.066 db/10 minutes and finally, when two isolators were used for coupling, the gain fluctuation was decreased to 0.025 db/10 minutes.

An examination was made of the effects of the isolators upon the frequency bandwidth and repeller voltage margin; however, no significant effect was observed.

Conclusions

The employment of the isolators in the coupling circuit of the cascaded reflex klystron amplifiers made the amplifier system stable, and gave it high gain, high sensitivity, wide dynamic range and high directivity. These effects were significant when two isolators instead of one were used in the coupling circuit. It was necessary to put the isolator immediately after the first-stage amplifier to obtain good amplifier performance.

Acknowledgment

The author extends his thanks to Prof. E. H. Scheibe of the University of Wisconsin and S. Krupnik and L. Heiting of Marquette University for their assistance.

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The Compatibility Problem in Single-Sideband Transmission*

K. H. Powers has written an interesting paper[1] discussing the compatibility problem in single-sideband transmission. I am pleased to have the opportunity to comment on his paper and this problem in general.

The C.C.I.R. has accepted a proposal by the U. S. to study our compatible single-sideband technique, and in doing so, the following definition was adopted in the Ninth Plenary Assembly at Los Angeles in 1959:

> "A Single-Sideband transmission is considered to be compatible if it can be received on the existing conventional double-sideband receivers without any modifications whatsoever and with satisfactory quality of reception."

This definition highlights the obvious fact that the term compatibility means compatible with existing AM receivers. In his paper, Powers fails to emphasize the tremendous importance of compatibility. The problem is to make new systems compatible with equipment now in widespread use and the question of compatibility with nonexistent equipment is of secondary interest.

It would appear that Powers' system fails to solve the compatibility problem for several reasons.

First, if negligible distortion is to be achieved, square law demodulation is required.

Second, since he proposes a controlled carrier system which has a variable average amplitude, normal AGC equipment would fail.

Finally, because of the variable average amplitude required, conventional AM transmitters would need considerable alteration such as the installation of a direct-coupled modulator. Therefore, his system does not appear to be compatible—neither with AM transmitters nor with existing AM receivers.

The Compatible Single-Sideband (CSSB) system has proven to be compatible by both careful laboratory measurements and numerous listening tests. Sideband rejection figures of over 30 db have been reported in a number of published papers. Confirming reports were also made to the FCC by stations performing experiments on this system. Thus, there appears to be little question as to the practicability of producing a CSSB wave.

It is interesting to note that the Soviets have spent considerable time and effort producing what they call "optimum modulation" which is an attempt at producing a compatible single-sideband wave. Their reports show undesired sideband radiation in the order of 18 db, which is some 12 to 14 db worse than what has been accomplished in this country. The Russians point out that this technique thickens the spectrum energy concentration in the desired sideband and that the over-all spectrum is approximately one-half the spectrum of conventional double-sideband. They also point out that use of such a system has been shown to improve the intelligence of words by 30 per cent, and phrases by 35 per cent, when the circuit was contaminated by impulse noise.

Mr. Powers devotes an appreciable portion of his paper to the description of a wide-band time-delay phase-shift network. The material is presented as one would present completely new material without any references to prior art. It is not common to reference old patents, but certainly his paper should at least have been brought up-to-date and have referenced a recent excellent paper by G. G. Gouriet and G. F. Newell entitled "A Quadrature Network for Generating Vestigial-Sideband Signals" in *Proc. of the IEE*, Part B, dated May, 1960.

Mention is made by Dr. E. L. C. White, in the same issue of *The Proceedings of the IEE*, that such networks are very old and he cites a 1931 patent which describes the general time-delay filter technique. Incidentally, practical use of such a network for sharp selectivity required in SSB operation is quite questionable at the present time because of the very large delay problem (in the millisecond range).

In conclusion, I would like to say that I feel that Mr. Powers' proposal does not solve the basic and important compatibility problem. Many of his general comments are correct but, of course, the CSSB system does not violate basic concepts. His hasty conclusions demonstrate the danger of insufficient care in the use of information theory techniques to prove that some new systems are impossible. A statement implying that something is impossible must always be made with the greatest of care.

By the way, I would like to point out that we have proposed two basic single-sideband systems. One is the CSSB system above discussed. The second is a much older technique which has found extensive commercial high-power use during the past seven years. This is the envelope elimination and restoration system for amplifying conventional SSB waves with high-efficiency class-C modulated amplifiers. The SSB wave thus produced may be received on any conventional AFC or suppressed-carrier SSB receiver. The extensive discussions published in Proceedings of the IRE have obscured the fact that there are two complete separate and distinct techniques.

I hope to offer a paper for publication in the next few months describing in some detail the basic concepts of compatible single-sideband system.

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* Received by the IRE, July 6, 1960.

[1] K. H. Powers, "The compatibility problem in single-sideband transmission," this issue, p. 1431.

Contributors

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He has been with the Department of Electronics, Weizmann Institute of Science, Rehovoth, Israel, since 1953, with the exception of one year, 1959, which was spent as a research associate in the Department of Electrical Engineering, University of Illinois, Urbana.

Dr. Aharoni is a member of the Israeli Physical Society and the American Physical Society.

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Lloyd V. Berkner (A'26–M'34–SM'43–F'47) was born in Milwaukee, Wis., on February 1, 1905. He received the B.S. degree in electrical engineering in 1927 from the University of Minnesota, which honored him with the Distinguished Alumni Award in 1952, and studied physics for two years, 1933–1935, at George Washington University. He has been awarded honorary Doctorate degrees by Brooklyn Polytechnic Institute, 1955; Uppsala University, Sweden, 1956; University of Calcutta, India, 1957; Dartmouth College, 1958; University of Notre Dame, 1958; University of Edinburgh, Scotland, 1959; Columbia University, 1959; University of Rochester, 1960.

L. V. Berkner

While still an undergraduate, he was engineer-in-charge at radio station WLB-WGMS, Minneapolis. For one year after graduation, he worked as an electrical engineer for the Airway Division of the U. S. Bureau of Lighthouses. He was an engineer with the first Byrd Expedition to the Antarctic in 1928–1930, and was awarded the U. S. Special Congressional Gold Medal, the Silver Medal of the Aeronautical Institute, and the Gold Medal of the City of New York for his services. For three years thereafter, he was on the staff of the National Bureau of Standards. From 1933 to 1941, he was a physicist with the Department of Terrestrial Magnetism of Carnegie Institution of Washington, and during 1940–1941, he was a consultant to the National Defense Research Committee.

An aviator in the Naval Reserve since 1926, Dr. Berkner was called to active duty as head of the Radar Section, Bureau of Aeronautics in 1941. He directed the Bureau's Electronics Materiel Branch from 1943–1945, and served on the U.S.S. *Enterprise* in 1945. He has held the rank of Rear Admiral, USNR, since 1955.

During 1946–1947, he was Executive Secretary of the Research and Development Board and remained a consultant to the Board until 1951. He was head of the Section on Exploratory Geophysics of the Atmosphere, Department of Terrestrial Magnetism, Carnegie Institution from 1947 to 1951. Since 1951, he has been President of Associated Universities, Inc., New York, N. Y., an educational institution which operates such research facilities as Brookhaven National Laboratory under contract with the Atomic Energy Commission and the National Radio Astronomy Observatory under contract with the National Science Foundation.

Dr. Berkner has held various offices in government, industry, and education. In the State Department, he served as Special Assistant to the Secretary of State and Director of the Foreign Military Assistance Program in 1949, and Chairman of the International Science Steering Committee and author of the document "Science and Foreign Relations," in 1949–1950. He was a consultant on special projects at M.I.T. during 1950–1952, and a consultant to the National Security Resources Board during 1952–1953. At present, he is on the Board of Directors of the Long Island Biological Association and of Texas Instruments, Inc., is charman of the Committee on Rockefeller Public Service Awards, and is a consultant to the President's Science Advisory Committee.

He received the Science Award of the Washington Academy of Sciences in 1941; Commendation Ribbon, Secretary of the Navy, in 1944; Honorary Officer, Order of the British Empire in 1945; U. S. Legion of Merit, in 1946; and Alumni Recognition Award of Acacia Fraternity, in 1954.

Dr. Berkner is Treasurer of the National Academy of Sciences and Chairman of its Space Science Board, Vice-President of the Special Committee on the IGY, and an ex-officio member of the Executive Committee of the U. S. National Committee for the IGY.

He is also President of URSI, Past President of the International Council of Scientific Unions, and a former member of the Executive Committee of the International Union of Geodesy and Geophysics.

He is President of the American Geophysical Union, and a Fellow of the American Academy of Arts and Sciences, AIEE, the American Physical Society, the Arctic Institute of North America, and the New York Academy of Sciences. He is a member of the AAAS, the American Philosophical Society, the Council on Foreign Relations, the Philosophical Society of Washington, the Washington Academy of Sciences, Acacia, Eta Kappa Nu, Plumb Bob, Scabbard and Blade, Theta Tau, the Cosmos Club, Explorers Club, and Century Association.

Abroad, he holds membership in the Royal Swedish Academy of Sciences, the Swedish Academy of Arts and Sciences, the Physical Society of India, the New Zealand Electronics Institute, and the Royal Society of Arts (England).

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John W. Bremer was born in Schenectady, N. Y., on September 30, 1932. He received the B.S.E.E. and M.S.E.E. degrees in 1955, both from the Massachusetts Institute of Technology, Cambridge. His thesis topic was concerned with multiple nonlinearities in control systems.

J. W. Bremer

He joined the General Electric Company in 1955, where he was first engaged in control systems analysis, including supervision of an analog computer facility; and then, in 1956 he became project engineer for Cryotron Development, located at the G.E. Research Laboratory. He is now holding the same position at the G.E. Computer Laboratory in Palo Alto, Calif.

Mr. Bremer is a member of the AIEE, Sigma Xi, and Eta Kappa Nu.

❖

W. A. Edson (M'41–SM'43–F'57) was born in Burchard, Neb., on October 30, 1912. He received the B.S.E.E. and M.S.E.E. degrees in 1934 and 1935, respectively, from the University of Kansas, Lawrence. The following two years were spent at Harvard University on a fellowship. Upon receiving the D.Sc. degree in communication engineering in 1937, he joined the systems development department of Bell Telephone Laboratories.

W. A. Edson

In 1941, he joined Illinois Institute of Technology, Chicago, as assistant professor of electrical engineering. He became professor of physics at the Georgia Institute of Technology, Atlanta, in 1945, professor of electrical engineering in 1946, and was director of the School of Electrical Engineering from 1951 to 1952.

In 1952, he joined Stanford University as a research associate in the Electronics Laboratory and acting professor of electrical engineering. In 1954 he joined the newly-founded General Electric Microwave Laboratory at Stanford, where he is now manager of the Klystron Subsection.

Dr. Edson is a member of the American Physical Society, the AAAS, and the AIEE.

❖

Harold H. Edwards (M'59) was born in Delhi, N. Y., on July 14, 1932. He received the Bachelor of Engineering Physics degree from Cornell University, Ithaca, N.Y., in 1955.

H. H. Edwards

After graduation, he joined the General Electric Company's Advanced Engineering Program, completing the three-year course in 1958. He then joined the GE Light Military Electronics Department's advanced engineering group in Ithaca, N. Y. He has been temporarily assigned to the GE Research Laboratory to develop thin-film superconductive devices for computer circuits.

❖

E. H. Frei (M'58) was born in Vienna, Austria, on March 2, 1912. He received the Ph.D. degree from the University of Vienna, Austria, in 1936.

E. H. Frei

He has been with the Department of Electronics, Weizmann Institute of Science, Rehovot, Israel, since 1948, with the exception of two years (1950–1952) which were spent with the Computer Project of the Institute for Advanced Study, Princeton, N. J. In 1958 he was made acting head of the Department of Electronics. At present, he is spending his sabbatical leave at the Stanford Research Institute, Menlo Park, Calif.

Dr. Frei is a member of the Israeli Physical Society and the American Physical Society.

❖

Marcel J. E. Golay (SM'51–F'60) was born in Neuchatel, Switzerland, on May 3, 1902. He attended the Gymnase Scientifique of Neuchatel where he received the B.S. degree in 1920, and the Federal Institute of Technology in Zurich where he received the License in Electrical Engineering in 1924. He attended the University of Chicago, where he obtained the Ph.D. degree in physics in 1931.

From 1924 until 1928, he was at the Bell Telephone Laboratories. After a short association with the Automatic Electric Company, Chicago, Ill., he entered the civil service, and was a member of the Signal Corps Engineering Laboratories at Fort Monmouth, N. J., until 1955. He is now serving as consultant to the Philco Corporation, Philadelphia, Pa., and to the Perkin-Elmer Corporation, Norwalk, Conn.

M. J. E. Golay

Dr. Golay is a member of the American Physical Society, the Optical Society of America, the American Rocket Society, and the Society for Applied Spectroscopy.

❖

Nick Holonyak, Jr. (S'51–A'55–M'59) was born on November 3, 1928 in Zeigler, Ill. He received the B.S. degree in electrical engineering in 1950, the M.S. degree in 1951, and the Ph.D. degree in 1954 from the University of Illinois, Urbana. While a graduate student at the University of Illinois, he was a teaching assistant, a research assistant in microwave tubes and semiconductor electronics, and held the Texas Instruments Fellowship in transistor physics. He joined the transistor development department of Bell Telephone Laboratories in 1954 and worked on diffused-impurity silicon devices. He was inducted into the U. S. Army in 1955 and served with the Signal Corps at Fort Monmouth, N. J. and in Japan. In 1957 he joined the Advanced Semiconductor Laboratory of the General Electric Company and has been involved in studies of power and signal *p-n-p-n* devices and tunnel diodes.

N. Holonyak, Jr.

Dr. Holonyak is a member of the American Physical Society, the Mathematical Association of America, Sigma Xi, Eta Kappa Nu, Tau Beta Pi, Pi Mu Epsilon, and Phi Kappa Phi.

❖

Jess J. Josephs was born in New York, N. Y. on January 4, 1917. He obtained his B.A. and M.S. degrees and the Ph.D. degree in physical chemistry at New York University, New York. After serving in the U. S. Navy as a chief engineer on an LST in World War II, he successively taught at Northwestern University, the University of Chicago, and Boston University. From 1950 to 1952 he was Assistant Director of the Upper Atmosphere Research Laboratories at Boston University. He is now at Smith College in Northampton, Mass, where he is Associate Professor of Physics and Chairman of the Department of Physics. He was a member of the summer staff at the Massachusetts Institute of Technology, Lincoln Laboratory, Lexington, Mass., in 1956, also 1957 and 1959. This past summer he was also with the Mitre Corporation, Bedford, Mass., where he is currently a consultant.

J. J. Josephs

His research and studies have carried him into the fields of infrared spectroscopy, reaction kinetics, electroluminescence and ferroelectricity.

Dr. Josephs is a member of the American Physical Society, American Association of Physics Teachers, American Chemical Society, Sigma Xi, and Phi Lambda Upsilon.

❖

Aditya K. Kamal (M'57) was born on July 5, 1927, in Kairana, India. He received the B.Sc. degree from Benares Hindu University, Benares, India, in 1946, the M.Sc. degree from Allahabad University, India, in 1948, the D.I.I.Sc. diploma from the Indian Institute of Science, India, in 1951, and the Doctor of Engineering degree from Paris University, Paris, France, in 1957

A. K. Kamal

From 1952 to 1954 he was employed as a staff engineer with the Marconi Wireless Telegraph Co., Ltd., Chelmsford, England, in their VHF development group. In 1954 he joined the Compagnie Generale de Telegraphic sans Fil, Paris, France, in the department of Applications of Radar. In 1957 he went to India to take the position of senior scientific officer and became the head of the Special Circuit Division at the Electronics Engineering Research Institute, Pilani, India. In 1959, after spending three months with Ligne Télégraphique et Téléphonique, Paris, France, as a senior engineer in the area of millimeter wave diode research, he came to the U. S. Since then he has been an assistant professor in the School of Electrical Engineering, Purdue University, Lafayette, Ind., where he directs the Millimeter Wave Research Laboratories.

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Harold B. Law (SM'46–F'52) was born at Douds, Iowa, on September 7, 1911. He received the B.S. degree in liberal arts and the B.S. degree in education, both in 1934, from Kent State University, in Kent, Ohio. He received the M.S. and Ph.D. degrees in physics from the Ohio State University, Columbus, in 1936 and 1941, respectively.

H. B. Law

Dr. Law taught elementary mathematics at schools in Maple Heights, Ohio, and in Toledo, Ohio, in 1935 and 1937–1938, respectively. In 1941, he joined the Radio Corporation of America, Camden, N. J., and in 1942 transferred to the David Sarnoff Research Center in Princeton, N. J. He has engaged in research on pickup tubes and, more recently, on color kinescopes.

Dr. Law is a member of the American Physical Society and of Sigma Xi.

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I. Arnold Lesk (S'47–A'52–M'56–SM'59) was born in Regina, Sask., Canada, on January 26, 1927. He received the B.Sc. degree in engineering physics from the University of Alberta, Edmonton, in 1948 and the Ph.D. degree in electrical engineering from the University of Illinois, Urbana, in 1951. At the University of Alberta he won several awards, including a gold medal in engineering. He held a Galvin Fellowship for one year and a half-time teaching assistantship for two years at the University of Illinois. His theses topics there were concerned with thyratrons and oxide semiconductors.

I. Arnold Lesk

In November, 1951, he joined the Electronics Laboratory of General Electric Company, Syracuse, N. Y., where he has worked on semiconductor devices and materials. In 1957 he was named consultant, semiconductor devices and techniques, in the Advanced Semiconductor Laboratory of the Semiconductor Products Department.

Dr. Lesk is a member of the American Physical Society, the Electrochemical Society, Sigma Xi, Eta Kappa Nu, Phi Kappa Phi and Pi Mu Epsilon.

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James A. Mullen (S'50–A'55–SM'57) was born on May 28, 1928 in Malden, Mass. He received the B.S. degree in physics in 1950 from Providence College, Rhode Island, after two year's service in the U. S. Army. He attended Harvard University, Cambridge, Mass., and received the M.S. degree in 1951 and the Ph.D. degree in 1955 in the field of random processes.

J. A. Mullen

Since the spring of 1955 he has been employed by the Research Division of Raytheon Co., Waltham, Mass., where he has been concerned with noise in electron tubes and communication systems.

Dr. Mullen is a member of Sigma Xi, the Society for Industrial and Applied Mathematics, and the Operations Research Society of America.

Vernon L. Newhouse (M'55) was born in Mannheim, Germany, on January 30, 1928. He received the B.S. degree in 1949, and the Ph.D. degree in 1952, both from the University of Leeds in England.

V. L. Newhouse

From 1951 to 1954, he was employed by Ferranti in Manchester, England, where he developed one of the earliest magnetic core memories, afterwards used in the Ferranti Mark II computer. From 1954 to 1956, he worked with RCA in Camden, N. J., where he headed the group responsible for the circuits of the Bizmac Computer System. After the completion of Bizmac, he developed a ferrite plate memory and participated in the discovery of the domain wall viscosity effect. Since 1957, he has been employed by the General Electric Research Laboratory, Schenectady, N. Y., where he is working on the physics and computer applications of superconductive films.

Dr. Newhouse is a member of the American Physical Society.

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Kerns H. Powers (S'48–A'52–M'57–SM'59) was born on April 15, 1925 in Waco, Texas. After serving in the U. S. Navy from 1942 to 1946, he entered the University of Texas, Austin, where he received the B.S. and M.S. degrees in electrical engineering in 1951. In that year, he joined the research staff of RCA Laboratories, Princeton, N. J., where he was engaged in research on color television and high resolution radar. In 1953, he was awarded an Industrial Fellowship in Electronics for graduate study at Massachusetts Institute of Technology, Cambridge, where he received the Sc.D. degree in 1956.

K. H. Powers

He then returned to RCA Laboratories, where he is now engaged in analytical studies in communication theory. He is also an Adjunct Professor of Electrical Engineering at Newark College of Engineering.

Dr. Powers is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

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Edward G. Ramberg (M'53–SM'53–F'55) was born in Florence, Italy, on June 14, 1907. Having entered Reed College, Portland, Ore., in 1922, he transferred two years later to Cornell University, Ithaca, N. Y., from which he received the B.A. degree in 1928. Continuing his study at Cornell, he became Heckscher Research Assistant in the physics department there until 1930. He then went to the University of Munich, Germany, from which he received the Ph.D. degree in 1932.

E. G. Ramberg

Dr. Ramberg's undergraduate work at Cornell was interrupted by a two-year employment as a scientific aide in optical computation with Bausch and Lomb Optical Company, Rochester, N. Y. After returning from Munich, in 1932, he went back to Cornell as a research assistant and worked on the theory of X-ray satellites and line widths, until 1935. He then joined the Electronic Research Laboratory of RCA Manufacturing Company, Camden, N. J., as a junior engineer. In 1942, he was transferred as research physicist to the RCA Laboratories at the David Sarnoff Research Center in Princeton, N. J. From 1943 to 1945, Dr. Ramberg worked in Civilian Public Service. Some of this time he was engaged in work on electronic aids for the blind, under the Office of Scientific Research and Development, at Haskins Laboratory, New York, N. Y. Having returned to the RCA Laboratories in Princeton, in 1946, he has worked in the fields of electron microscopy, electron optics and optics related to television, theory of thermoelectric refrigeration, and other problems in electron physics. In 1949, he interrupted this work to serve as visiting professor at the University of Munich.

Dr. Ramberg is a Fellow of the American Physical Society. He is also a member of the Electron Microscope Society of America, the American Association of Physics Teachers, the American Association for the Advancement of Science, and of Sigma Xi and Phi Beta Kappa.

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Jerome J. Tiemann (M'59) was born on February 21, 1932, in Yonkers, N. Y. He received the B.Sc. degree in physics from the Massachusetts Institute of Technology, Cambridge, in 1953 and the Ph.D. degree from Stanford University, Stanford, Calif., in 1960. His dissertation was concerned with high energy nuclear physics.

J. J. Tiemann

In 1957 he assumed his present position in the semiconductors studies section of the General Physics Research Department at the General Electric Research Laboratory, Schenectady, N. Y. His fields of interest include fundamental research in physics, the physics and electrical properties of devices, device development, and circuit applications. Since 1958, most of his work has been connected with the tunnel diode.

Dr. Tiemann is a member of the American Physical Society and Sigma Xi.

Books

The Other Side of the Moon, issued by the USSR Academy of Sciences, translated by J. B. Sykes

Published (1960) by Pergamon Press, 122 E. 55 St., N. Y. 22, N. Y. 36 pages. 8 Illus. 8¼×10¾. $2.50.

The book contains a short, nontechnical description of the third Soviet Space Rocket launched on October 4, 1959. It also contains printed halftone reproductions of the photographs made in the Space Rocket of the "other side" of the moon.

The apparatus, techniques and the orbital path of the rocket are described qualitatively, but in some detail. The photographic apparatus and the radio transmission capabilities are presented. The point at which roll was eliminated and photo stabilization was achieved for proper photography is given.

Although 35 mm film was used and lens diameters of 1½ inches and 2 inches are given indirectly, and the radio transmission capability of 1000 lines per frame is claimed, the resolution is much poorer than would be expected under these conditions. This poor resolution is not explained.

C. H. Hoeppner
Radiation, Inc.
Melbourne, Fla.

Electrons and Phonons, by J. M. Ziman

Published (1960) by Oxford University Press, 417 Fifth Ave., N. Y. 16, N. Y. 523 pages+30 index pages+xiv pages. 166 figures. 6¼×9¼. $13.45.

Graduate students and others interested in gaining a rather deeper view of solid state physics will welcome this book. From a pedagogical standpoint, it is one of the clearest and most literate textbooks on the subject that has yet appeared. Its subject matter, as its title implies, is concerned with those properties of solids which are determined by electrons and lattice vibrations (phonons).

The book is developed in a logical and detailed manner, the emphasis being placed on physical interpretation rather than mathematical rigor. It begins by presenting the ideas and formal apparatus necessary for understanding the behavior of lattice vibrations. A corresponding discussion for electrons along the lines of traditional single particle theory and a brief review of band theory is followed by a comprehensive treatment of phonon-phonon, electron-electron, and electron-phonon interactions as well as a discussion of interactions with various imperfections in the solid, such as point imperfections, dislocations, and stacking faults. These are necessary ingredients for the "Theory of Transport Phenomena in Solids," that is, of those aspects of solid state theory dealing essentially with electric and thermal conduction, the central theme and the announced subtitle of the book. The formal transport theory built around the Boltzmann equation is presented from the usual heuristic viewpoint with relatively little attention paid to the foundations. A thorough discussion of variational methods in transport theory, tying these methods to well-known thermodynamic principles, will be found useful, since such techniques are becoming increasingly important. In the area of applications of the theory, the chapters on lattice thermal conduction in insulators and metals and electronic conduction in metals are to be highly recommended. In contrast, the discussion of semiconductors is much more cursory, many important theoretical contributions in this area being completely omitted. Indeed, those expecting a treatise covering all important aspects should take note of the statement made in the chapter on mobility in semiconductors: "From the point of view of the electronics industry what we have to say is largely of academic interest."

It is also to be regretted that several important theoretical advances of the last five years are essent ally neglected. There is an all too brief mention of galvanomagnetic effects in high fields, and quantum mechanical transport theory in general. Also, the theory of superconductivity which falls within the purview announced by the book's title is completely omitted. Understanding of these more difficult areas would undoubtedly have been considerably furthered if the author had chosen to deal with them, since his ability for presenting lucid expositions of frequently abstract subject matter is quite considerable.

H. Ehrenreich
General Electric Research Lab.
Schenectady, N. Y.

Introduction to Electrical Engineering, by Robert P. Ward

Published (1960) by Prentice-Hall, Inc., Englewood Cliffs, N. J. 362 pages+6 index pages+2 appendix pages+x pages. Illus. 6¼×9¼. $8.50.

Professor Ward intended this book as an introductory text to "electrical science in engineering." It includes the subject material normally contained in an early electrical engineering course. However, greater emphasis is placed on the electric circuit and magnetic concepts.

An introductory chapter precedes fundamental electrical concepts and laws of the electric circuit. This is followed by electrical networks limited to linear resistances and dc sources only, since capacitance and inductance are taken up in a later phase of the book. This is a good chapter, but could stand some expansion in the early part on network graphs. Construction of indicating instruments is discussed before getting into the uses of ammeters, voltmeters, and bridges. Characteristics of metallic conductors are given consideration.

The next five chapters are concerned with magnetic concepts, ferromagnetic materials, the magnetic circuit, electromagnetic forces and induction, and motional electromotive force. This area was particularly well organized and is one of the strong points of the book.

Electric field concepts are covered well in the next chapter. Here his approach to capacitance through the electric field is straightforward. Capacitance is followed by two chapters, one on conduction in liquids and gases and the final one on transient response of simple circuits.

The author covered a considerable amount of material in a matter of some three hundred and fifty pages. Since he intended this as an introduction to electrical engineering, the short space was adequate for this type of coverage.

On the whole, he did an excellent job of organizing his material and presenting it in an "easy to read" manner. He was generally brief and to the point.

Throughout the book are very good selections of problems, frequently preceded by example problems worked out in detail. At the end of each of twelve of the fourteen chapters is a group of study questions. The example problems, the problems scattered throughout the book and the study problems enhance the value of the book for classroom work.

Joseph J. Gershon
DeVry Technical Institute
Chicago, Ill

Experiments in Electronics, by W. H. Evans

Published (1959) by Prentice-Hall, Inc., 70 Fifth Ave., New York 11, N. Y. 334 pages+38 appendix pages+ix pages. Illus. 4×6¼. $6.95.

A representative and substantial sampling of the material encountered in an introductory course is to be found in the one hundred laboratory exercises (two experiments devoted to each of fifty different subjects) which comprise the present work. Emphasis is upon the uses of electron devices in circuits, with secondary attention paid to the physics of the devices themselves.

The general pattern of presentation of each experiment is this: Purpose (a sentence or two), Information (from one or two to sometimes several paragraphs of general comments, but not included in all experiments), Preliminary (questions or problems to direct thought to pertinent aspects of the subject matter), Performance (the actual instructions as to what to do), Calculations and Conclusions, References, and Materials Required (in a few cases citing manufacturer and type number). It is evident that this work is intended for use in conjunction with a textbook and other reference works, and that the laboratory exercises are intended to supplement classroom lectures and problems.

The selection of material is sufficiently broad to afford a year or so of laboratory experiments with good variety. Included are experiments which afford familiarization with commonly encountered electronic instruments, and with the fundamental properties of tubes, transistors, and solid state rectifiers and their important applications in amplifier, oscillator, wave-shaping, light sensing and power-supply circuitry.

Included is a good selection of transistor experiments—thirteen of the fifty pairs—and proper emphasis is placed upon biasing.

The transistor inverter power supply circuit might well be added. The semiconductor junction diode receives less than its fair share of attention; and with recent interest and effort centered upon parametric amplifier circuits and tunnel diodes, these topics must surely be included in another edition of this work. Too, the categorical exclusion of microwave experiments on the grounds, expressed in the Preface, of "diversity of available equipment" hardly seems justified.

In general, the presentations are terse and to the point. Not all the details are spelled out, leaving room for the exercise of judgment by the student. This is commendable, for it is perhaps more important, even at this level, that the student be able to decide what is worth doing than it is that he be able to do it when instructions are laid before him. Included in the book is a good assortment of graphical data on vacuum tubes, transistors, thyratrons and phototubes. The value of this volume is obvious, especially to the student left to work out laboratory exercises with a minimum of help and to the instructor organizing a new laboratory.

CARL R. WISCHMEYER
Rice University
Houston 1, Tex.

Nuclear Fusion, Dr. William P. Allis, Ed.

Published (1960) by D. Van Nostrand Co., Inc., 120 Alexander St., Princeton, N. J. 471 pages+12 index pages+5 appendix pages+viii pages. Illus. 6¼×9¼. $12.50.

One of the major milestones in progress on the control of hydrogen fusion for production of nuclear power was the Second International Conference on Peaceful Uses of Atomic Energy, held in Geneva in September, 1958. Some months later, the United Nations published a complete Proceedings, in 33 volumes, of which volumes 31 and 32 were devoted to the 111 papers on controlled fusion The present Van Nostrand book is one of a 3-volume condensation, under the title, "Second Geneva Series on the Peaceful Uses of Atomic Energy," and under the general editorship of Dr. J. G. Beckerley. This volume is edited by Professor Allis, of the Massachusetts Institute of Technology, a well-known research leader in plasma physics. In the book, Professor Allis has carefully selected 28 of the 111 papers on controlled fusion; he has edited and/or condensed these papers, divided them among 11 chapter headings and written a very short critical discussion for each chapter, in which many of the omitted papers are mentioned (by number).

It is well known that this Conference saw the first public release of the vast amount of U.S.A. research on controlled fusion, and that the U.S.A. contributed about 60 per cent of the papers on the subject. The papers which Professor Allis has selected comprise 16 from the U.S.A., 8 from the U.S.S.R., and 4 from the U.K., a fair representation of the major contributions. A few of the more important U.S.A. papers were omitted on the grounds that they have received wide circulation in technical journals. Unfortunately, no detailed reference is given to such journal publication, so the reader will find it a chore to locate an omitted paper. Even more strange, this reviewer could find no mention anywhere in the book of the official Proceedings, in which *all* the papers were published in full. An Appendix lists these papers but gives no indication of how a reader can obtain them. In connection with references, each chapter has a long list, many of which refer to U. S. Atomic Energy Commission reports, conference reports, etc. These are important original references, but are not normally available in an average technical library.

The editor deemed it desirable for brevity and clarity to make changes and omissions in the original texts, thus giving the present book a somewhat more unified presentation, and probably easier to read than the originals, in many cases. On the other hand, the omissions from original texts make for strange reading in a few instances, as for example, the survey paper by Tuck on Los Alamos work. In the original, the important Scylla experiment was fully described; in Professor Allis' book, the word Scylla is used repeatedly in the condensation of Tuck in Chapter I, but there is no indication whatever of the nature of this piece of apparatus, or the details of the experiment. There is also no reference to where this material can be found. A knowledgeable reader will find, through some effort, that Scylla is described in Chapter VII, whose title is Ring Discharges. An omission in the book is the paper by Trubnikov and Kudryavtsev, which was the most exciting of the Conference, since it predicted that electron cyclotron radiation might prevent any successful deuterium reaction, in a magnetic field. Later calculations by others have shown that, although the numerical results exaggerated the effect, the phenomena may be of major importance.

To summarize this reviewer's comments, the book is not a substitute either for a survey paper, or for Volumes 31 and 32 of the complete Proceedings. However, it is a very convenient summary of a group of important papers. The editing and references represent a great deal of work by the editor, helpful in many instances, and a hindrance in others.

E. W. HEROLD
Varian Associates
Palo Alto, Calif.

Solid State Physics in Electronics and Telecommunications, Vol. I, Semiconductors, Part 1, M. Diserant and J. L. Michiels, Eds.

Published (1960) by Academic Press Inc., 111 Fifth Ave., N. Y. 3, N. Y. 638 pages+xxiii pages. Illus. 6¼×10. $16.00.

This book represents the first of two volumes recording the proceedings on semiconductors of an international conference on solid state physics in electronics and telecommunications sponsored by the International Union of Pure and Applied Physics in Brussels during June, 1958. The papers are divided into six categories: Preparations of semiconductors and allied problems; Properties of semiconductors; Solid state theory; Effects of intense electric fields in semiconductors; Noise in semiconductors; and Surface phenomena. Authors from all over the world are represented, and a reading knowledge of French and German is desirable since, of a total of about 61 papers, 14 are in French and 7 are in German.

A review of the papers indicates them to be of uniformly high quality and to cover a wide range of topics which are virtually required reading for workers in the growing field of semiconductor circuits. The quality of the printing and illustrations is good. The majority of the papers are well annotated with biographical references.

R. P. BURR
Circuit Research Co.
Glen Cove, L. I., N. Y.

RECENT BOOKS

Bell, D. A., *Electrical Noise, Fundamentals and Physical Mechanism.* D. Van Nostrand Co., Inc., 257 Fourth Ave., N. Y. 10, N. Y. $9.75. A unified reference and source book on electrical noise, concerned primarily with the mechanism of noise in physical devices which are of interest to physicists and electrical engineers.

Chapman, Alan J., *Heat Transfer.* The Macmillan Co., 60 Fifth Ave., N. Y. 11, N. Y. $9.00. A teachable textbook in the field of heat transfer, presenting the fundamental approach in a vigorous manner. Presupposes a knowledge of mathematics and the fundamental principles of thermodynamics and fluid dynamics. Written for use in an introductory course in heat transfer taught to senior or first-year graduate students.

Gaynor, Frank, *Aerospace Dictionary.* Philosophical Library Inc., 15 East 40 St., N. Y. 16, N. Y. $6.00. An up-to-date reference work of essential terminology in space exploration. Includes every abbreviation in use, together with abbreviations of missiles and vehicles and their designations.

Halliday, David and Resnick, Robert, *Physics for Students of Science and Engineering.* John Wiley and Sons, Inc., 440 Fourth Ave., N. Y. 16, N. Y. $6.00. A beautifully illustrated and highly readable textbook for a general physics course for engineers and scientists. Parts I and II are available in separate editions, and in a combined edition as well. For a more detailed description of the general features of this set, see the review of Part I (F. Herman, PROC. IRE, July, 1960).

Neal, J. P., *Electrical Engineering Fundamentals.* McGraw-Hill Book Co., Inc., 330 W. 42 St., N. Y. 36, N. Y. $8.50.

Oldfield, R. L., *Radio-Television and Basic Electronics, Second Edition,* American Technical Society, 848 East Fifty-Eighth St., Chicago 37, Ill. $4.95. A learner-oriented method is used throughout the book. Electrical principles, electronic phenomena, functioning of electronic equipment, differences between Monophonic and Stereophonic High Fidelity Systems are among the subjects discussed.

Shiere, Alexander, Ed., *Low-Frequency Amplifier Systems.* John F. Rider Publisher, Inc., 116 W. 14 St., N. Y. 11, N. Y. $1.80. Deals with the subject matter by keeping the mathematical treatment to a minimum without simplifying the material to a point which might mislead the reader. Review questions at the end of each chapter.

Scanning the Transactions

Tunnel diodes, with their extremely high frequency limit, are finding important use in high-speed computer logic circuits. Unlike transistor switching circuits, tunnel diode circuits are essentially bilateral and require special circuit arrangements to obtain a unilateral characteristic for the transmission and amplification of digital signals. This is the same situation which was encountered by the inventors of the parametron. These same workers, and their colleagues in Japan, have applied the principles of the parametron circuits to the tunnel diode case, and have developed experimental logic circuits with clock rates of 30 mc. Their work shows that with presently available germanium tunnel diodes, a clock rate of 100 mc should be possible, and with the future development of improved tunnel diodes, a 1000-mc rate may become feasible. (E. Goto, *et al.*, "Esaki diode high-speed logical circuits," IRE Trans. on Electronic Computers, March, 1960.)

The first known wave-type amplifier apparently pre-dates the traveling-wave amplifier by more than a half century. A recently published book makes reference to a lecturer who in 1890 described an acoustic amplifier which made use of growing waves similar in some respects to those encountered in traveling-wave tubes, double-stream amplifiers and other present-day electron beam devices. The amplifier was attributed to Chichester Bell, cousin of Alexander Graham Bell. In this amplifier, a thin jet of water was directed against a small rubber diaphragm which was coupled to a horn. Any wave on the stream of water would cause the diaphragm to vibrate, producing a sound from the horn. When a tuning fork or music box was pressed to the water nozzle, growing waves were set up in the stream of water which produced a volume of sound from the horn sufficient to be heard through a lecture hall. (J. R. Pierce, "An interesting wave amplifier," IRE Trans. on Electron Devices, April, 1960.)

Radar spelled backwards is radar. There is no way of telling how many times this word has been inadvertently spelled in reverse by careless authors, editors and typesetters. It is doubtful that anyone makes this mistake deliberately. Some words, however, have been spelled backwards on purpose, and have achieved some measure of fame as a result—Serutan, for example. In the world of electronics, the mho has become a widely known, frequently used, and valuable upside-down companion to the ohm. Some have gone so far as to propose the term yrneh as the inverted counterpart of the upright henry. It was perhaps inevitable, therefore, that we should eventually find ourselves using a HTIMS chart. As the name implies, a HTIMS chart is an inverted Smith chart. It is used for plotting impedances that have negative resistance components. In one form, the two types of charts are placed back to back to form a "double SMITH-HTIMS chart." Although this gives a most common name a most uncommon appearance, it also provides the engineer with a valuable addition to his bag of tools. (R. L. Kyhl, "Plotting impedances with negative resistance components," IRE Trans. on Microwave Theory and Techniques, May, 1960.)

What is a matched filter? A filter that is made to be responsive to a particular signal waveform by designing its impulse response to be the time reverse of the waveform is said to be matched to the waveform. The use of such filters in detecting signals is substantially equivalent to the operation of cross-correlation detection, and it has been known for 15 years that these procedures were desirable for extracting a simple waveform from noise. However, it seems that in a wide variety of more complicated cases, matched filters continue to play a role as part of the optimum form of signal processing equipment. Some examples of these more complicated situations are multipath communication channels, radar detection or estimation in the presence of noise, clutter, or uncertainties in target parameters, and others. A special issue of the Information Theory Transactions is devoted to experimental and theoretical work on matched filter communication and detection. A tutorial survey of the field is followed by eight contributed papers on various pieces of current work. Among these papers are to be found a number of discussions of equipment and systems that have been built according to the guidelines for optimality laid down by the theoretical studies. (Special Issue on Matched Filters, IRE Trans. on Information Theory, June, 1960.)

Eight U. S. computer experts, recently returned from Russia, have compiled a comprehensive 49-page report on Soviet computer technology. Their visit, which followed a similar visit by a Russian delegation to the U. S., was arranged by the U.S.S.R. Academy of Sciences and the U. S. National Joint Computer Committee. The latter organization represents the three computer societies of the United States, namely, the IRE, AIEE, and Association for Computing Machinery. This unusual document covers the history of the delegation, itinerary, Russian organizations, machines, applications, circuits, components, education, and even Chinese developments. Edited by the chairman of PGEC, who was one of the delegates, the report has now been published by the PGEC in their March Transactions. (H. W. Ware, *et al.*, "Soviet computer technology—1959," IRE Trans. on Electronic Computers, March, 1960.)

Monograph on radar detection theory. Occasionally important papers appear that are just too long for most technical journals, and yet are not of interest to commercial publishers. The Professional Group on Information Theory plans to make an occasional such paper available to its members in the form of Monographs—extra issues in the Transactions series. The first Monograph is a reprint of the classic series of papers by J. I. Marcum and P. Swerling on signal detectability in radars. These studies came out as a series of RAND reports beginning in 1947, and have been in continuing demand. A particularly valuable part of the volume is the extensive series of tables and graphs of functions that are always cropping up in problems of detecting and integrating sinewave signals in noise. (J. I. Marcum and P. Swerling, "Studies of target detection by pulsed radar," IRE Trans. on Information Theory, April, 1960.)

Abstracts of IRE Transactions

The following issues of TRANSACTIONS have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

Sponsoring Group	Publication	Group Members	IRE Members	Non-Members*
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* Libraries and colleges may purchase copies at IRE member rates.

Antennas and Propagation

Vol. AP-8, No. 3, May, 1960

Editorial Comment—(p. 234)

Propagation at 36,000 MC at Near-Vertical Incidence—W. L. Flock and W. D. Hershberger (p. 235)

Fading characteristics at 36 kmc over a line-of-sight path in the Los Angeles basin have been shown to be closely correlated with meteorological conditions, particularly with the relatively persistent, low-level temperature inversion of the area. No positive evidence of the influence of atmospheric pollutants has been found, but it has been shown that suitably located microwave paths can be of value for locating and monitoring temperature inversions when they are accompanied by sufficient variation in water-vapor content. The relation of diurnal variations in propagation characteristics to diurnal variations in the temperature inversion and in atmospheric turbulence indicate that the refraction mechanism is the predominant one in causing the observed large fading amplitudes. The view is further stength-ened by the relatively noncritical relation of fading to the proximity of the inversion layer.

Two Statistical Models for Radar Terrain Returns—L. M. Spetner and I. Katz (p. 242)

A statistical approach to radar backscattering from terrain is taken in this paper. The normalized radar cross section, σ_0, has been computed for two different terrain models. The value of σ_0 is obtained for both models as a function of grazing angle, θ, and radiation wavelength, λ. The first model is a distribution of isolated independent scatterers such as corner reflectors. For such surfaces, a wavelength dependence for σ_0 is obtained, and, depending upon the density of scatterers and their average size, the theoretical results indicate that the local dependence of σ_0 on λ can be as λ^{-6}, λ^{-4} or λ^{-2}. For such surfaces, σ_0 is independent of θ.

Where reflection occurs from specularly reflecting facets on the surface and where the distribution of surface slopes is Gaussian, the θ dependence turns out to be of the form

$$e^{-k \cot^2 \theta \, 2s_0^2}$$

where s_0 is the standard deviation of the surface-slope distribution. The precise form of σ_0 depends upon the space spectrum of the slopes. Two cases are worked out, one where such a spectrum is flat out to some cutoff, and the other where the space spectrum has a single peak at a particular wave number. In either case, for small enough λ, σ_0 varies as λ^{-2}. As the wave-length becomes large compared to the facet size, the facet no longer behaves as a specular reflector and instead becomes more like an isotropic scatterer.

For any particular wavelength one may expect that the radar return be the result of the addition of two types of backscattering. The large facets will behave as specular-type reflectors, while the smaller facets will act as the isotropic scatterers discussed in the first model.

Radar Terrain Measured Return at Near-Vertical Incidence—A. R. Edison, R. K. Moore, and B. D. Warner (p. 246)

An experimental program to investigate the reradiation properties of terrain at near-vertical incidence was carried out. Data were obtained at 415 and 3800 mc, using narrow-pulse type radar at altitudes of 2000 to 12,000 feet over many different target areas.

At frequencies over 400 mc, most terrain acts as a scatterer of energy even at near-vertical incidence with a backscattering "radiation pattern" that drops off rapidly as the angle of incidence is increased. An exception occurs for heavily wooded areas which appear as nearly isotropic scatterers.

At 415 mc, the radar cross section per unit area at vertical incidence ranges between values of 0.7 for woods to approximately 4 for city targets, while at 3800 mc, the variation ranges between values of 0.8 for woods to about 18 for some city targets. If the ground were a lossless isotropic scatterer, the radar cross section per unit area would be 2 at vertical incidence.

For a wide beam antenna, the fading range between the level exceeded by 95 per cent of the return pulses and the level exceeded by only 5 per cent of the return pulses is generally between 12 and 17 db, except for a few very smooth areas which give considerable specular (nonfading) type of return and have a smaller fading range.

A New Mathematical Approach for Linear Array Analysis—D. K. Cheng and M. T. Ma (p. 255)

It is well known that linear antenna arrays are representable mathematically by polynomials. However, even for the simplest case of a uniform array, properties of its radiation pattern are conventionally analyzed by examining the transcendental form of the array factor and some of its important characteristics have been determined only approximately. For a more general array, a closed form of the associated polynomial is usually not obtained and the analysis becomes quite difficult. This paper proposes a new approach for linear array analysis. Basically, the current distribution in the discrete elements of a linear array is considered as the sampled values of a continuous function. Known relations in Z transforms developed for sampled-data systems can then be used to express the array polynomail in a closed form. Mathematical techniques for determining important properties of the array pattern are developed. Typical examples illustrating the applications of this new approach are given.

Log Periodic Dipole Arrays—D. E. Isbell (p. 260)

A new class of coplanar dipole arrays is introduced. The antennas described provide unidirectional radiation patterns of constant beamwidth and nearly constant input impedances over any desired bandwidth. The broad-band properties are achieved by making use of the principles of log periodic antenna design. Models are discussed which are capable of providing 8- to 9-db directive gain with an associated input standing wave ratio of 1:2:1 on a 75-ohm feeder, and this performance is independent of frequency. The free-space properties of several of these arrays have been measured and the results are presented. The antenna configuration is simple, permitting practical methods of fabrication, and the design should prove useful in many applications. It makes possible, for example, the construction of "all-wave" rotatable beams of very low cross section for use in the HF to UHF spectrum.

Beam Pointing Errors of Long Line Sources—M. Leichter (p. 268)

The relation between the statistics of the antenna beam pointing direction and the phase and amplitude errors at the source has been obtained to first order in the mean-square errors, under certain restrictions, for long line sources. It is shown that when the desired phase at the source is a constant, the results are, to first order, independent of the amplitude errors. When the desired amplitude is also constant, there is a simple formula for computing the allowable rms-phase error at the source when the pointing direction is required to lie in a given angular range with a given probability. When the amplitude distribution corresponds to the Taylor-modified $(\sin x)/x$ pattern, the allowed rms-phase error is obtained from the constant-amplitude case by a multiplicative factor which depends only on the one parameter characterizing the Taylor distribution. This function is plotted for the range corresponding to sidelobe ratios of 13.2 to 40 db. At 40 db the allowed rms-phase errors are about three fourths of the allowed rms-phase errors at 13.2 db (constant amplitude) for the same uncertainty in the pointing direction. The results are applied to a hypothetical example and to an actual "Mills Cross" for illustrative purposes.

Mutual Impedance Effects in Large Beam Scanning Arrays—P. S. Carter, Jr. (p. 276)

An analysis is presented of the driving point impedance of the elements in a flat array of infinite vertical height but finite horizontal width. It is assumed that each of the elements is fed by a separate amplifier, having infinite internal impedance, and that the amplifiers can be phased to direct the beam at various positions in space. The radiating elements considered are infinitely long wires spaced on half-wavelength centers, each backed by a conducting ground plane spaced one-quarter wavelength from the elements. Values of driving point impedance are computed for the 61 elements in arrays 30 wavelengths wide. Values of the driving point impedance near the center of the array are found to agree closely with the values computed for infinitely wide arrays while the driving-point impedance of elements near the edge of the array are found to deviate considerably from the values at the center.

Mutual Coupling Effects in Large Antenna Arrays: Part I—Slot Arrays—S. Edelberg and A. A. Oliner (p. 286)

A periodic structure approach is presented for the analysis of the impedance properties of large antenna arrays. The method is applied to a two-dimensional array of slots, in which each slot is fed by a separate waveguide and the array radiates into a half-space. The slot spacing and progressive phasing in the array may be arbitrary, however. The periodic structure approach permits a waveguide-type analysis of the half-space which automatically includes all mutual coupling effects. Both the susceptance and conductance of the slot are evaluated for arbitrary scan angle, and the effects associated with the appearance of higher order beams are considered.

Spiral Antennas—Walter L. Curtis (p. 298)

The radiation fields of the Archimedes spiral are derived by approximating the spiral with a series of semicircles. The calculated patterns are shown to have excellent correlation with experimentally determined patterns. It is shown that the high-frequency limit is determined by the feed configuration and that the low-frequency limit occurs when the outside diameter is a little greater than a half wavelength.

Ground Constant Measurements Using a Section of Balanced Two-Wire Transmission Line—Erik J. Kirkscether (p. 307)

When an open-circuited section of unshielded balanced two-wire transmission line is introduced perpendicularly into earth (or some sample under test), the electrical characteristics of the latter may be found by simple input-impedance measurements. By laboratory sample measurements the classical short- and open-circuited method can be used. Some exact and approximate procedures are presented and their utility and practical limitations discussed. Some precautions as to how possible errors and inexactnesses in the measurements and following calculations may be avoided are given. As an example, a typical earth sample is tested in a frequency range from 0.6 to 400 mc, with graphic representation of the most important electrical constants: conductivity, dielectric constant, attenuation, velocity of propagation, etc., which exhibit great variations in the frequency range cited. The measurement method presented seems to be adequate to use in small mobile equipment, with which the ground in general can be tested in its original site and under natural conditions without the necessity of being removed.

The Archimedian Two-Wire Spiral Antenna—Julius A. Kaiser (p. 312)

A pair of equally excited but oppositely sensed Archimedean two-wire spirals situated close to one another in the same plane—a doublet—is used to generate a linearly polarized field in which the direction of polarization and phase are controlled or varied independently of each other by rotation of the spiral radiators. An array of these doublets can be made to scan by rotation of the several spiral elements; an eight-doublet array which was made to scan over an 83° sector with small amplitude variation is discussed. A doublet fed from a ring network can be employed as a polarization diversity circuit. A virtual doublet is achieved by placing a single spiral in a right angle trough. A preliminary scanning array comprising four spirals in a trough was made to scan ±36°. The possibility of using a parasitic spiral in conjunction with a driven spiral for obtaining linear polarization of variable direction and phase is indicated. Also, a brief, simplified analysis of the two-wire Archimedean spiral is presented, which leads to the concept of higher-order modes of radiation.

Coupled Leaky Waveguides I: Two Parallel Slits in a Plane—S. Nishida (p. 323)

Theoretical expressions are derived for the effects of mutual coupling between two parallel leaky wave antennas located in an infinite plane. The leaky wave antennas treated are slitted rectangular waveguides, the propagation constants of which are modified by the coupling. It is shown that the attenuation constants are influenced significantly but that the phase constants are changed only slightly, so that the coupling is different from that between neighboring surface wave lines. The nature of the coupling effects are illustrated by numerical calculations.

Experimental Study of Diffraction Reflector—J. H. Provencher (p. 331)

A microwave antenna has been designed and constructed on the principles of parageometric optics formulated by Toraldo di Francia and on principles similar to those of Fresnel rings and the diffraction grating. Its surfaces are all zones of cones and are simple to construct.

The chosen design parameters were incorporated in two K-band ($\lambda = 1.24$ cm, $\lambda = 1.22$ cm) models. Experimental results show good agreement with theory. Scanning characteristics are superior to those of the paraboloidal or spherical reflector, and spherical aberration and coma are minimized, and the effects of astigmatism are minimized by using a "compromise focus."

Communications (p. 337)

Announcement of the NBS Bibliography on Radio Propagation (p. 349)

Contributors (p. 350)

Electron Devices

Vol. ED-7, No. 2, April, 1960

An Interesting Wave Amplifier—J. R. Pierce (p. 73)

This paper concerns the theory and structure of an amplifier involving growing waves.

High Average Power Dissipation in Helix Tubes—Arthur H. Iverson (p. 74)

By precisely shrinking the glass envelope around the helix to a controlled depth, thus providing an exceptionally low thermal impedance to the outside of the envelope, and passing a suitable coolant over the envelope, it has been found that the power dissipation capability of helix circuits can be extended by more than one order of magnitude. A qualitative analysis indicates that the heat transfer characteristics should be as good as the experimental results indicate.

Targets for Storage and Camera Tubes—H. R. Day, H. J. Hannam, and P. Wargo (p. 78)

Several new information storage targets have been developed for use in storage tubes and camera tubes. One, a double-sided mosaic, is useful primarily in storage tubes. The others, based upon the use of thin self-supported films of magnesium oxide, can be used in many tubes. Having first found application in image orthicons, they provide markedly improved performance and longer life.

Theoretical Power Output and Bandwidth of Traveling-Wave Amplifiers—H. Sobol and J. E. Rowe (p. 84)

Expressions are developed to calculate the theoretical power output of traveling-wave amplifiers using any type of RF structure. Calculations are made for helix-type tubes and it is shown how to calculate the power output of tubes using more dispersive structures in terms of calculations made for helix tubes.

The principal factors accounting for higher power output of dispersive structures are presented and discussed. The gain and bandwidth of forward-wave helix amplifiers are derived from the small-signal theory as functions of frequency and it is shown that the gain in db times the frequency bandwidth is a constant as a function of helix length for high $\gamma_0 a'$ and the gain times the bandwidth squared is a constant for low $\gamma_0 a'$.

Correction to "Thermally-Induced Cracking in the Fabrication of Semiconductor Devices"—T. C. Taylor (p. 94)

Contributions to the Knowledge of Excess Noise—H. P. Louis (p. 95)

This paper describes measurements on the noise behavior of a long, magnetically focused electron beam with a convergent Pierce-type electron gun having a shielded cathode. It is found that the excess noise phenomenon is initiated by a scallop-beam amplification and is essentially dependent on the nonlaminarity of electron flow in the beam. The beam reaches a state of high thermal equilibrium which may be caused by strong electron-electron interaction. An estimation shows that this should be theoretically possible.

High Performance Impedance Transformation with the EFP-60 Secondary-Emission Pentode—N. S. Nahman and E. J. Martin, Jr. (p. 99)

The unique characteristics of the EFP-60 secondary-emission pentode make it possible to utilize this tube in a novel cathode follower circuit in which the impedance transformation action is "enhanced" by connection of the secondary-emission dynode back to the cathode. An analysis of this circuit in two slightly different forms indicates the possibility of achieving cathode-follower action with virtually unity gain and consequently high input-to-output impedance transformation ratio. The results of the analysis are verified by experimental data. Some aspects of the circuit as applied to millimicrosecond pulse work are discussed, and conclusions are reached with respect to optimization of tube characteristics for this particular type of application.

Contributors (p. 105)

Program of 1959 Electron Devices Meeting (p. 107)

Electronic Computers

Vol. EC-9, No. 1, March, 1960

A. A. Cohen, Chairman, 1960–1961 (p. 1)

High Density Digital Magnetic Recording Techniques—A. S. Hoagland and G. C. Bacon (p. 2)

The merit of any high density detection method is ultimately dependent on the "resolution" characteristic of the magnetic recording components. Justification of readback waveform synthesis through "single pulse" superposition is given. A comprehensive, yet general readback simulation program is described which will automatically, for any characteristic pulse, simulate all possible readback signal patterns and test them for specified reading logic as a function of bit density. Amplitude, phase, peak, etc., sensing are compared and the influence of parameter variation on performance indicated. Good correlation with experiment has been realized and has greatly reduced time at the bench. The significance of pulse waveform is clearly revealed and this study has provided a guide to head design (ring and probe), permitting the optimization of a total recording system for high-density storage.

The Optimal Organization of Serial Memory Transfers—Arthur Gill (p. 12)

This paper is concerned with the optimal compilation of programs whose function is to transfer words of information from one location in a serial memory to another. The most important optimization tool is the "timing schedule," which facilitates the analysis of various transfer schemes and the determination of the fastest one. The procedure described for optimizing serial transfers is readily programmable for computer execution, and is directly applicable to a general class of transportation problems.

The Design of Diode-Transistor NOR Circuits—Dale P. Masher (p. 15)

Considerations leading to the adoption of diode-transistor NOR circuitry for a moderately fast data-processing system are outlined. The design of the basic circuit is treated in detail. Development of a unique set of compatible logic packages from the basic circuit is described. This set is unique in the sense that a single type of diode-transistor circuit is used to provide the great majority of logic and storage functions required in the system. This single circuit type, which functions as a NOR circuit, is embodied in two package types. One package provides a single gate with a fan-in of five. The other package provides two gates, each with a fan-in of two. The latter type may be externally connected to provide a set-reset flip-flop. Only two other package types are used. The first is a passive transfer circuit which greatly simplifies shift register logic, and the second is a delay package which is closely related to the basic NOR circuit.

Esaki Diode High-Speed Logical Circuits—E. Goto, K. Murata, K. Nakazawa, T. Moto-Oka, Y. Matsuoka, Y. Ishibashi, H. Ishida, T. Soma, and E. Wada (p. 25)

Logical circuits using Esaki diodes, and which are based on a principle similar to parametron (subharmonic oscillator element) circuits, are described. Two diodes are used in series to form a basic element called a twin, and a binary digit is represented by the polarity of the potential induced at the middle point of the twin, which is controlled by the majority of input signals applied to the middle point. Unilateral transmission of information in circuits consisting of cascaded twins is achieved by dividing the twins into three groups and by energizing each group one after another in a cyclic manner.

Experimental results with the clock frequency as high as 30 mc are reported. Also, a delay-line dynamic memory and a nondestructive memory in matrix form are discussed.

Magnetic Analogs of Relay Contact Networks for Logic—D. B. Armstrong, T. H. Crowley, U. F. Gianola, and E. E. Newhall (p. 30)

Two techniques are described for designing multi-apertured magnetic structures capable of realizing any specific logic function. The structures are made from rectangular hysteresis loop material. The designs are derived from the corresponding relay contact network by replacing each current carrying conductor in the relay circuit with its analog in the magnetic circuit, a flux-carrying conductor, replacing the emf with a pulsed mmf; and replacing each back contact with a saturable portion of the magnetic circuit in the topological equivalent of the contact network. A flux reversal through a saturated portion may then be blocked by means of an inhibiting current applied to a suitable winding, the current representing a logical input variable. Thus flux may be "steered" through the magnetic circuit in a manner analogous to the steering of current through the contact network. For planar structures the first technique may be used to obtain the analog of series-parallel and bridge type circuits; the second technique is suitable only for the analogs of series-parallel circuits. It is pointed out that the analog of a relay tree can be used as a standard structure suitable for realizing any Boolean function. Representative examples of both designs are shown, and experimental data are given.

The Determination of Carry Propagation Length for Binary Addition—George W. Reitwiesner (p. 35)

It is well known that the expected maximum length of nonzero carry propagation in the addition of two uniformly distributed binary numbers of n-digits each is less than $\log_2 n$. The propagation of both zero and nonzero carry is required in the employment of asynchronous self-timing addition. For the addition of two n-digit binary numbers which are uniformly distributed, a simple recursive algorithm is readily derived for the exact determination of the expected maximum length of zero or nonzero carry propagation.

Regular Expressions and State Graphs for Automata—R. McNaughton and H. Yamada (p. 39)

Algorithms are presented for 1) converting a state graph describing the behavior of an automaton to a regular expression describing the behavior of the same automaton (section 2), and 2) for converting a regular expression into a state graph (sections 3 and 4). These algorithms are justified by theorems, and examples are given. The first section contains a brief introduction to state graphs and the regular-expression language.

A Method for the Design of Pattern Recognition Logic—Sam D. Stearns (p. 48)

The general problem of pattern recognition is regarded as a problem wherein the recognition device is presented with a plane array of black-or-white elements and must decide to which general class (pattern) this array belongs.

A method for reducing the necessary amount of logic is presented. It is basically a method for reducing Boolean equations in many variables which contain large numbers of redundant or "don't care" terms.

The reduced logic is in the form of Boolean functions of the black-or-white elements. Some experimental results, in which this logic was mechanized with diodes, are discussed.

Optimization of Reference Signals for Character Recognition System—I. Flores and L. Grey (p. 54)

The role of signal structure in a signal discrimination system is discussed. The optimality criterion for reference signals for detection in the case of white Gaussian independent noise is defined. The need for normalization of the reference signals is demonstrated. A geometric interpretation is presented. Optimum classes are obtained and several examples cited. A theoretical optimum class of signals is derived against which any set of signals developed within given constraints may be rated.

Frequency-to-Period-to-Analog Computer for Flowrate Measurement—Ted W. Berwin (p. 62)

The Frequency-to-Period-to-Analog Computer is a special purpose nonlinear analog computer which accepts an ac voltage of varying frequency, acts upon the period of each cycle, computes the inverse of the time period, $e=1/T$, and holds the information for the period of the next cycle. Thus, the output voltage is a level which is proportional to the input frequency $f=1/T$ computed once for every cycle. The system is accurate to better than ± 0.5 per cent of 2/3 full scale. Application of the computer is described and results presented for fast readout and recording of gas and liquid turbine type flowmeters. Extensions of the circuits used can produce voltages proportional to ln t or $1/t^2$, for time t greater than a small positive number.

Soviet Computer Technology—1959—W. H. Ware, Editor; S. N. Alexander, P. Armer, M. M. Astrahan, L. Bers, H. H. Goode, H. D. Huskey, M. Rubinoff, W. H. Ware, Contributors (p. 72)

The paper presents a factual account of the trip of the 1959 U. S. technical delegation in computers to the Soviet Union. It includes the itinerary, descriptions of specific Soviet computers, descriptions of certain computing centers, a discussion of Soviet computer-oriented education, and a description of current circuit and component development. In appendices are given the instruction repertoire of the URAL-I and the URAL-II machines and an analysis of some magnetic cores. The paper is extensively illustrated and contains a bibliography of relevant Soviet documents.

Correspondence (p. 121)

Reviews of Books and Papers in the Computer Field (p. 130)

Abstracts of Current Computer Literature (p. 134)

PGEC News and Notes (p. 152)

Human Factors in Electronics

Vol. HFE-1, No. 1, March, 1960

Frontispiece and Editorial—Curtis M. Jansky (p. 2)

Man-Computer Symbiosis—J. C. R. Licklider (p. 4)

Man-computer symbiosis is an expected development in cooperative interaction between men and electronic computers. It will involve very close coupling between the human and the electronic members of the partnership. The main aims are 1) to let computers facilitate formulative thinking as they now facilitate the solution of formulated problems, and 2) to enable men and computers to cooperate in making decisions and controlling complex situations without inflexible dependence on predetermined programs. In the anticipated symbiotic partnership, men will set the goals, formulate the hypotheses, determine the criteria, and perform the evaluations. Computing machines will do the routinizable work that must be done to prepare the way for insights and decisions in technical and scientific thinking. Preliminary analyses indicate that the symbiotic partnership will perform intellectual operations much more effectively than man alone can perform them. Prerequisites for the achievement of the effective, cooperative association include developments in computer time sharing, in memory components, in memory organization, in programming languages, and in input and output equipment.

Pattern Recognition and Display Characteristics—W. R. Bush, R. B. Kelly, and V. M. Donahue (p. 11)

This paper reports experimental results of human operator performance in a visual recognition task. The work began with a method of generating families of complex patterns to simulate certain characteristics of visual sensor displays, such as radar and infrared returns. The experimental effort was directed toward establishing criteria for predicting human operator performance in a map-matching task. The operators' task was to recognize which of four patterns presented simultaneously with a reference pattern belonged to the reference pattern family. The measure of performance was the time in seconds taken by the operator to make a selection. Response times were more rapid when the reference pattern was less complex than the comparison than when the reference pattern was the more complex. Analysis of the display characteristics led to the selection of four physical measures to be used in predicting operator performance. These measures—pattern length, pattern density, and two measures of pattern complexity—correlated highly with response time, were not highly intercorrelated, and were applicable to natural sensor returns. The four measures were found to account for a high degree of the total variance. Regression equations were derived which predict performance from known values of the four measures.

The Use of Quickening in One Cöordinate of a Two-Dimensional Tracking System—J. W. Duey and R. Chernikoff (p. 21)

In a previous study, it was found that tracking error in one coordinate of a two-dimensional tracking system was affected by the dynamics used in the other coordinate. In particular, tracking performance progressively deteriorated as the dynamics in the two coordinates became more dissimilar. The present study seeks to extend these findings by determining how the introduction of quickening into one coordinate of a second-order, two-coordinate tracking system would affect the performance in the unquickened coordinate. In the light of the previous study, it might be expected that quickening one coordinate would degrade the performance of the other. On the other hand, the simplification of the tracker's task induced by quickening might effect an improvement in performance.

The findings suggest that a counterbalance of the above factors was achieved, since quickening one coordinate had no effect on performance in the other.

Desirable Push-Button Characteristics—Richard L. Deininger (p. 24)

This paper reports the results of studies in a series concerning the characteristics of push-button keysets that people can operate quickly, accurately and conveniently. The studies investigated push-button arrangements, button top and lettering characteristics, and push-button force-displacement characteristics. Considerable latitude exists in the design of keysets if only keying performance is considered. The preference judgments were somewhat more selective, particularly for the force-displacement characteristics of the button mechanism.

The Relation of Electronic and Optical Gain to System Performance—S. Seidenstein and H. P. Birmingham (p. 30)

An experiment was conducted to investigate the effect of adjusting display gain upon man-machine system performance in a simple aided tracking system. Gain was varied in two ways: electrically by changing amplification, and optically by changing the distance from the scope to the eye. Manipulation of gain by each method produced similar changes in system performance. Over the ranges studied, system error decreased as display gain was increased. This result agrees with predictions based upon closed-loop control system theory and suggests the feasibility of including additional experimental variables within the theory.

Communications (p. 33)

Reviews of Current Literature (p. 35)

Information Theory

Vol. IT-6, No. 2, April, 1960

Special Monograph—Issue Studies of Target Detection by Pulsed Radar—J. I. Marcum and P. Swerling

A Statistical Theory of Target Detection By Pulsed Radar (Rand Research Memo. RM-754, December 1, 1947)—J. I. Marcum (p. 59)

This reports presents data from which one may obtain the probability that a pulsed-type radar system will detect a given target at any range. This is in contrast to the usual method of obtaining radar range as a single number, which can be taken mathematically to imply that the probability of detection is zero at any range greater than this number, and certainty at any range less than this number.

Three variables, which have so far received little quantitative attention in the subject of radar range, are introduced in the theory:

1. The time taken to detect the target.
2. The average time interval between false alarms (false indications of targets).
3. The number of pulses integrated.

It is shown briefly how the results for pulsed-type systems may be applied in the case of continuous-wave systems.

Those concerned with systems analysis problems including radar performance may profitably use this work as one link in a chain involving several probabilities. For instance, the probability that a given aircraft will be detected at least once while flying any given path through a specified model radar network may be calculated using the data presented here as a basis, provided that additional probability data on such things as outage time etc., are available.

The theory developed here does not take account of interference such as clutter or man-made static, but assumes only random noise at the receiver input. Also, an ideal type of electronic integrator and detector are assumed. Thus the results are the best that can be obtained under ideal conditions. It is not too difficult, however, to make reasonable assumptions which will permit application of the results to the currently available types of radar.

The first part of this report is a restatement of known radar fundamentals and supplies continuity and background for what follows.

The mathematical part of the theory is not contained herein, but will be issued subsequently as a separate report.[23]

Mathematical Appendix (RM-753, July 1, 1948) (p. 145)

In a previous report[28] a statistical theory of radar detection was presented in outline form. The mathematical details were omitted, in order that the main ideas and results might be made available as soon as possible.

This report contains the mathematics that led to the results presented in Ref. 28.

In addition, several subjects are briefly discussed that were not covered in Ref. 28. These are collapsing loss, antenna beam shape loss, the effect of signal injection, limiting loss, and moving target indication.

For references see page 264.

Probability of Detection for Fluctuating Targets (RM-1217, March 14, 1954)—P. Swerling (p. 269)

This report considers the probability of detection of a target by a pulsed search radar, when the target has a fluctuating cross section.

Formulas for detection probability are derived, and curves of detection probability vs range are given, for four different target fluctuation models.

The investigation shows that, for these fluctuation models, the probability of detection for a fluctuating target is less than that for a nonfluctuating target if the range is sufficiently short, and is greater if the range is sufficiently long.

The amount by which the fluctuating and nonfluctuating cases differ depends on the rapidity of fluctuation and on the statistical distribution of the fluctuations. Fig. 18, p. 307, shows a comparison between the nonfluctuating case and the four fluctuating cases considered.

Information Theory

Vol. IT-6, No. 3, May, 1960

Introduction—Paul E. Green, Jr. (p. 310)

An Introduction to Matched Filters—George L. Turin (p. 311)

In a tutorial exposition, the following topics are discussed: definition of a matched filter; where matched filters arise; properties of matched filters; matched-filter synthesis and signal specification; some forms of matched filters.

Joint Estimation of Delay, Doppler, and Doppler Rate—Phillip Bello (p. 330)

The methods of inverse probability have been used by Woodward and others to obtain lower bounds on variances of radar parameter estimators. Previous results on the lower bounds of variances of delay and Doppler estimators have assumed that the reflecting object travels with a constant line of sight velocity and does not cause scintillation in the radar return. Using the inverse probability approach, this paper derives expressions for the minimum variances of estimators of the delay, Doppler, and Doppler rate of a radar return assumed to consist of a long train of pulses with independent scintillation from pulses to pulse.

Processing Gains Against Reverberation (Clutter) Using Matched Filters—E. C. Westerfield, R. H. Prager, and J. L. Stewart (p. 342)

The Woodward ambiguity function is discussed in connection with the output of a matched filter. A formula for the treatment of sonar reverberation or radar clutter is set up in terms of the ambiguity function. This formula is applied to determine the effect of signal waveform on the output signal-to-reverberation power ratio of a matched filter for a simple distribution of randomized scatterers.

On New Classes of Matched Filters and Generalizations of the Matched Filter Concept—David Middleton (p. 349)

In this paper it is shown how the earlier concepts of the matched filter may be generalized by recognizing explicitly the decision-making character of most reception systems. Accordingly, when an approach making use of statistical decision theory is applied for both signal detection and extraction, a variety of new classes of matched filters (Bayes matched filters) can be defined. These can be described specifically in the critical situation of threshold reception, where system optimality is at a premium. It is shown, for incoherent reception in some important special instances, that matched filters based on maximizing output signal-to-noise ratio (the S/N matched filters of the earlier theory) are also optimum from the broader, decision viewpoint. The required optimum filters are themselves time-varying and nonunique, and thus permit a measure of design freedom. In all instances, realizable filters are possible, and it is shown how their weighting functions may be determined. Both discrete and continuous filtering on a finite interval, $(0, T)$, are considered.

Correlation Detection of Signals Perturbed by a Random Channel—Thomas Kailath (p. 361)

We show that the concept of correlation detection of deterministic signals in additive Gaussian noise can be extended in a natural manner to the detection of signals that are transmitted through a "Gaussian" random channel besides being corrupted by additive Gaussian noise. Such situations are typical in communication over scatter-multipath channels (with or without a specular component). In the deterministic case, the receiver essentially crosscorrelates the received signal with the signal before the additive noise was introduced. When a random channel is present, however, this latter signal, *i.e.*, the output of the random channel, is not known at the receiver. However, knowing the statistics of the channel and the noise, the receiver can make an estimate of it from the received signal on the hypothesis that a particular signal was transmitted. The optimum receiver then crosscorrelates this estimate with the received signal

A Matched Filter Communication System for Multipath Channels—Steven M. Sussman (p. 367)

A matched filter communication system is described whose underlying principles are based on the Rake. The point-to-point synchronous teletype system employs complex lumped parameter networks to generate and receive a pair of long-duration, broadband signals representing Mark and Space respectively. The receiver contains a pair of matched filters whose output is a narrow pulse when the matching waveform is applied. One advantage of the system, arising from the long duration of the signals, is an increase in energy per teletype baud when operating under a peak power limitation. Another is that multiple propagation paths due to ionospheric reflection are resolved by the broadband signals, resulting in the appearance of the multipath pattern at the output of the Mark or Space matched filter. Recombination of paths is achieved by means of a recirculating delay line tuned to the teletype baud rate in conjunction with parallel multiplier-integrators in the Mark and Space channels. The combination acts as a self-adjusting correlation detector for the multipath pattern.

A Matched Filter Detection System for Complicated Doppler Shifted Signals—Robert M. Lerner (p. 373)

A matched filter system is described which was designed to detect complicated signals subject to a wide range of possible Doppler shifts. A 100 tap band-pass delay line was used in conjunction with a resistor weighting matrix to synthesize signals and filter characteristics. The system could handle a signal with a duration-bandwidth product of 100 over a range of Doppler frequency shifts 17 times the reciprocal of the signal duration. A theoretical discussion of the Doppler effect is given, making use of conjugate functions or Hilbert Transforms. Various engineering compromises which simplify the construction of matched filters are suggested. The performance of the resulting signal detection system was within 5 db of that of an ideal theoretical model.

Optical Data Processing and Filtering Systems—L. J. Cutrona, E. N. Leith, C. J. Palermo, and L. J. Procello (p. 386)

Optical systems, which inherently possess two degrees of freedom rather than the single degree of freedom available in a single electronic channel, appear to offer some advantages over their electronic counterparts for certain applications. Coherent optical systems have the added property that one may easily obtain many successive two-dimensional Fourier transforms of any given light amplitude distribution, or, by use of astigmatic optics, one-dimensional transforms can be obtained. Therefore, most linear operations of an integral transform nature are easily implemented. The optical implementation of integral transforms which are of importance to communication theory is discussed; the general problems of optical filter synthesis and multichannel computation and data processing are introduced, followed by a discussion of potential applications. Astigmatic systems, which permit multichannel operation in lieu of two-dimensional processing, are treated as a special case of general two-dimensional processors. Complex input functions are discussed with relation to their role in coherent optical systems.

Quaternary Codes for Pulsed Radar—George R. Welti (p. 400)

A class of quaternary codes is described, and an algorithm for generating the codes is given. The codes have properties that make them useful for radar applications: 1) their autocorrelation consists of a single pulse, 2) their length can be any power of two, 3) each code can be paired with another code (its mate) of the same class in such a way that the cross-correlation of mates is identically zero, 4) coded waveforms can be generated in a simple network the number of whose elements is proportional to the base-2 logarithm of the code length, and 5) the same network can be readily converted to a matched filter for the coded waveform.

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Microwave Theory and Techniques

Vol. MTT-8, No. 3, May, 1960

Administrative Committee, 1959–1960 (p. 272)

An Analysis of Four-Frequency Nonlinear Reactance Circuits—David K. Adams (p. 274)

Several advantages of multiple-frequency nonlinear reactance circuits are described in this paper. In particular, a circuit is considered in which a nonlinear reactance couples four basic frequencies: ω_0, ω_1, ω_2, and ω_3; these are so related that $\omega_2=\omega_0+\omega_1$ and $\omega_3=\omega_0-\omega_1$. Here, ω_0 is taken to be the power source or pump. It is found to be desirable to allow for the possible presence of the pump harmonic, $2\omega_0$, and individual cases are characterized by whether $2\omega_0$ is present or not. The major results are as follows: 1) Unlimited amplification gain is theoretically possible at frequencies higher than the pump, by reflecting negative input resistance at ω_2, but without relying on any effects due to pump harmonics. 2) Unlimited up- or down-conversion gains between ω_1 and ω_2 are theoretically possible in the additional presence of the first pump harmonic, but without reflecting negative input or output resistance. 3) Unlimited amplification gain is theoretically possible at frequencies both lower and higher than the pump fundamental, without reflecting negative input resistance.

Some Properties of Three Coupled Waves—Laszlo Solymar (p. 284)

The paper deals with the problem of three waves, 1, 2, and 3, in which waves 2 and 3 are coupled to wave 1 but not to each other. The general solution for the amplitudes of the waves is given in closed form. It is shown that for certain values of the parameters growing waves can exist. Numerical solutions for the location of the boundaries of the growing wave regions are plotted. It is shown furthermore that under certain conditions the power can be completely transferred from wave 1 to waves 2 and 3.

Examples on traveling-wave tubes, waveguide couplers, and backward-wave oscillators illustrate the applicability of the theory.

Noise Figures of Reflex Klystron Amplifiers—Koryu Ishii (p. 291)

The noise figure of the 2K25 reflex klystron amplifier was investigated. The noise figure of the reflex klystron amplifier depends on operating frequency, electronic impedance, circuit impedance, and operating electronic mode. Experimental results show that a noise figure of 5 db is possible under particularly carefully adjusted conditions. In order to obtain the low-noise figure, careful electronic tuning and the impedance adjustments are particularly important. Generally, relatively low noise figures were obtained when the electronic tuning was good. Noise figures of cascaded reflex klystron amplifiers were also investigated experimentally. Noise figures of the cascaded amplifier were generally higher than that of the single stage amplifier, but still low enough to use this reflex klystron amplifier as a preamplifier of a microwave receiver to increase the sensitivity of the receiving system.

On Measurements of Microwave $\bar{E}$ and $\bar{H}$ Field Distributions by Using Modulated Scattering Methods—Ming-Kuei Hu (p. 295)

The modulated scattering method of Justice, Rumsey, and Richmond for measuring $\bar{E}$ field distribution is extended to the measurement of $\bar{H}$ field distribution by using a loop scatterer formed by two diodes. This diode loop method has the particular advantage of eliminating the large and undesirable effect produced by the associated $\bar{E}$ field when measuring the $\bar{H}$ field.

A scattering analysis of the modulated diode loop is presented. It explains the principle of this new method and also supports the advantage mentioned above. A similar analysis for the modulated diode scatterer used in measuring E is also presented. It is believed that the explanation based upon this analysis for the E measurement is more satisfactory than that given by Richmond which is based upon a qualitative description of the diode scatterer.

Analysis of Certain Transmission-Line Networks in the Time Domain—W. J. Getsinger (p. 301)

Many linear components in nondispersive transmission line are made up solely of commensurate lengths of line of various characteristic impedances. Such components have impulse responses that are a series of equispaced impulses, and, as a result, their frequency responses can be written as a Fourier series. Given the period and coefficients of the Fourier series describing the frequency response, the time response of the circuit to any pulse can be written down immediately as a sum of replicas of the applied pulse, each replica having an amplitude given by the coefficient of a term in the series, and occurring at a time determined by the period of that term of the series.

The pulse responses of stepped transmission-line transformers, backward-coupling hybrids, and branch-line hybrids are determined and, after assuming a simple applied-pulse shape, are plotted.

Sets of Eigenvectors for Volumes of Revolution—J. Van Bladel (p. 309)

The electric and magnetic eigenvectors of a volume of revolution can be written in terms of two-dimensional scalar and vector functions. These functions are the eigenfunctions of certain linear transformations in the meridian plane. The form of the transformation is examined, and much attention is devoted to the orthogonality properties of their eigenfunctions and the calculation of their eigenvalues from variational principles.

A Printed Circuit Balun for Use with Spiral Antennas—R. Bawer and J. J. Wolfe (p. 319)

A novel printed circuit balun is described which is particularly well suited to applications where space is at a premium. The design utilizes unshielded strip transmission line, but is readily adaptable to all of the common printed circuit transmission line techniques. When the balun is housed within the cavity of a spiral antenna, boresight error is virtually eliminated, ellipticity ratios of less than 2 db are maintained over an azimuth angle greater than $\pm 60°$, and the input standing-wave ratio is less than 2:1 over an octave frequency range. Experimental results are given and additional applications are described.

The P-I-N Modulator, an Electrically Controlled Attenuator for MM and Sub-MM Waves—F. C. De Ronde, H. J. G. Meyer, and O. W. Memelink (p. 325)

The construction and performance of a millimeter wave modulator are described. The main part of the modulator consists of a *p-i-n* germanium structure inserted into a rectangular waveguide. A modulation depth of 11 db could be obtained at frequencies up to 5 kc, this modulation being caused for the greatest part by attenuation.

Discontinuities in the Center Conductor of Symmetric Strip Transmission Line—H. M. Altschuler and A. A. Oliner (p. 328)

A systematic measurements program has been carried out to check the validity of theoretical formulas for the equivalent circuit parameters of a variety of discontinuities in the center conductor of symmetric strip transmission line. These theoretical formulas have been in part previously available and are in part new or modified. Results indicate that, in general, these formulas are adequate for most engineering purposes and that certain of the network parameters can be neglected.

A Variational Integral for Propagation Constant of Lossy Transmission Lines—Robert E. Collin (p. 339)

By assuming that the current on a lossy transmission line flows in the axial direction, only a variational integral for the propagation constant can be readily obtained. This variational integral shows that the usual power loss method of evaluating the attenuation constant is valid for general transmission lines. This variational integral also shows that the perturbation of the loss-free phase constant is due to the increase in magnetic field energy caused by penetration of the field into the conductors.

Measurement of Bandwidth of Microwave Resonator by Phase Shift of Signal Modulation—D. S. Lerner and H. A. Wheeler (p. 343)

Bandwidth is measured by transmission of a signal with sine-wave modulation through a microwave resonator under test. The modulation frequency is adjusted so that the envelope is delayed 45° with respect to the input, indicating that the two sideband frequencies are separated by the half-power bandwidth. The resonance ratio (Q) is then equal to the ratio of carrier frequency over twice the modulation frequency. This depends on observations of these frequencies and the modulation phase shift, but not on the amplitude. It is insensitive to detuning or incidental frequency variation of the resonator or the signal. In a resonant cavity tested, an observed bandwidth of 30 kc at 700 mc indicated that $Q=23{,}300$.

A Y-Junction Strip-Line Circulator—U. Milano, J. H. Saunders, and L. Davis, Jr. (p. 346)

The theoretical approach to the three-port symmetrical circulator is reviewed and presented in a form valid for the most general waveguide case.

A strip-line Y-junction circulator is described and the performance of different units in the band 800–1600 mc is illustrated.

The new type of device described offers, for the low-frequency region of the microwave spectrum, advantages of simple design, light weight, and great compactness with respect to the classical types. When operated with a permanent magnet it gives—in a bandwidth of about 4 per cent—isolation greater than 20 db, insertion loss ≤ 0.4 db, and input VSWR ≤ 1.20.

A Wide-Band UHF Traveling-Wave Variable Reactance Amplifier—R. C. Honey and E. M. T. Jones (p. 351)

The techniques developed for designing periodically loaded traveling-wave parametric amplifiers using variable-reactance diodes are described in detail. An amplifier was built and tested with two different sets of eight diodes. The performance of the amplifier with each set of diodes agrees substantially with the theoretical predictions, the measured noise figures being about 1.2 db higher than the theoretical values in each case. The gain of the second amplifier varies from a minimum of 6.7 db to more than 13 db over the band from 550 to 930 mc, with a measured noise figure of 2.3 db for wideband noise inputs in the middle of the band, corresponding to about 4.9 db for single-frequency inputs.

On the Theory of Strongly Coupled Cavity Chains—M. A. Allen and G. S. Kino (p. 362)

A chain of identical cavity resonators coupled together through slots in their common walls forms a band-pass microwave filter. The pass band characteristics of such a system are determined by a combination of field theory and circuit theory. The fields in the cavities are expressed in terms of the normal modes of the uncoupled cavities. The fields in the neighborhood of a slot are determined by representing the slot as a transmission line. Irrotational components of the field in the cavities account for direct slot-to-slot coupling. The method successfully predicts both the dispersion characteristics and field distributions over large frequency ranges for many practical systems, such as slow-wave circuits for high-power traveling-wave tubes.

Microphony in Waveguide—I. Goldstein and S. Soorsoorian (p. 372)

This paper describes the mechanism of phase modulation by waveguide in the presence of a high intensity acoustic field. X-band rectangular was studied to determine:

a) Resonant frequency in a transverse vibrational mode.

b) Means of minimizing phase modulation.

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Reliability and Quality Control

Vol. RQC-9 (Follows PGRQC-16), April, 1960

Calculation of Average Failure-to-Failure Time of Equipment—M. A. Sinitsa (p. 1)

On Reservation by Placement Method—M. A. Sinitsa (p. 6)

Consideration of substitution techniques for reliable design is usually based upon statistical independence of failures of operating and spare components.

A method of design of a reliable system, consisting of many operating and spare components, for the case of statistical relation between failures of operating and spare components is considered. In a number of cases more reliable operation of systems, can be achieved by providing substituting reservation, instead of permanent switching, to a spare (or "hot" reserve) component. Effective usage of a single spare component substituted for several operating components is considered.

Some Results of Mathematical Reliability Theory—B. R. Levin (p. 14)

Purchasing Ability—E. J. Breiding (p. 19)

The contributing responsibility and a system for controlling component reliability as part of the procurement function are discussed. The system, which is particularly applicable to large-scale procurement, consists of three major controls:

1) Component Vendor Approval Procedure—a tool for obtaining and documenting bona fide sources capable of supplying components to specified time and quality requirements.
2) Vendor Delivery Performance—a purchasing control for maintaining and upgrading performance from selected vendors.
3) Vendor Quality Rating—a quality control tool to aid purchasing in maintaining and upgrading quality of products from selected vendors.

Diagnosis of Equipment Failures—J. D. Brule, R. A. Johnson, and E. J. Kletsky (p. 23)

This paper introduces several new concepts which are applicable to the problem of diagnosis of equipment failures. Following the definitions of an equipment, an element of the equipment, and the model of a test, a general diagram of a testing procedure is developed. The testing diagram is constructed in such a way that the various tests needed and the probability of failure of the elements are readily incorporated. While it is found that a completely general testing diagram becomes quite complicated even when the equipment under consideration is not intricate, a major simplification is obtained by introducing a simplified diagram with suitably restricted tests. This simplified testing diagram may be used repeatedly in order to find all the faulty elements of the equipment.

With reference to the testing diagram, it is possible to compute the minimum average cost of diagnosing the equipment. This appears to be the most useful measure of the efficiency of a test procedure. The order of magnitude of this optimization problem is discussed and solutions for two special cases are obtained by analogy with an optimum coding problem.

An Information Theory Approach to Diagnosis—R. A. Johnson (p. 35) (Abstract)

Reliability of Parallel Electronic Components—H. Walter Price (p. 35)

Electronic components are frequently connected in parallel as a measure to increase reliability. Whether the result of such a parallel connection results in an increase or a decrease in reliability, and the amount of such increase or decrease, is a function of the open-circuit failure probability and the short-circuit failure probability. Equations are derived which permit a determination of the increase or decrease of reliability when components are connected in parallel. Some curves are included to aid the circuit designer in this determination.

Evaluation and Prediction of Circuit Performance by Statistical Techniques—J. Marini and R. Williams (p. 40)

A method is described for predicting circuit performance to the extent that it is dependent on part performance. The basis for the prediction is the performance of parts as measured at test points fixed by the part specifications. Implicit in this method is the assumption that the distribution of part performance at the test points can be predicted from consideration of the specifications. Such an assumption is necessary to any attempted prediction of this nature.

An empirical equation giving circuit performance in terms of part performance as measured at the test points is assumed. The exact form of the equation is determined experimentally, by means of regression analysis of data consisting of sets of measurements of breadboard models of the circuit. The empirical equation is then used mathematically to calculate the distribution of the circuit performance from the assumed distributions of part performance.

The method has been applied successfully to predict the laboratory performance of an ac amplifier and a telemetering oscillator. In principle, the method can be extended to the prediction of equipment or system performance.

Reliability Using Redundancy Concepts—L. Depian and N. T. Grisamore (p. 53)

This paper introduces a new method of using redundancy to obtain reliable operation of electronic circuits. Switching circuits are used as examples to illustrate the method. A comparison is made between the majority logic method and the averaging method proposed in this paper. The comparison shows that the averaging method should provide a greater circuit reliability than the majority method if the components of each circuit have the same reliability.

How Can We Attain High Reliability of Complex Military Electronic Equipments?—Morris Halio (p. 61)

Each piece of military electronic equipment passes through various phases in its normal life cycle. These are planning, design and development, pilot production, manufacture, transportation, storage, operation and maintenance. Each of these stages is replete with opportunities for the introduction of unreliabilities. This paper points out the pitfalls which may be encountered and makes specific recommendations to avoid these so that the full amount of potential reliability may be realized in the final equipment.

What Price Unreliability?—Dana A. Griggin (p. 70)

Criteria for Determining Optimum Redundancy—R. E. Barlow and L. C. Hunter (p. 73).

Redundant circuits whose components may suffer either an open-circuit or a short-circuit type of failure are considered. A probabilistic model for such circuits is proposed. Two criteria for determining optimum redundancy are studied. A formula for obtaining the number of components which maximize reliability is derived for general failure distributions. A table for obtaining the number of components which maximize the expected life of the circuit is presented for the case of exponential failure.

Interval Estimation of Product Reliability by Use of the Noncentral t Distribution—R. E. Schafer (p. 77)

Is There Anything New in Reliability?—W. D. McGuigan (p. 81)

Contractual Aspects of Reliability—R. W. Smiley (p. 84)

Reliability Predictions, A Cast History—R. A. Davis and W. Wahrhaftig (p. 87)

Contractor Management Looks at Reliability Program Activities—W. B. Laberge (p. 90)

Design Information Interchange Among Co-Contractors—Martin Barbe (p. 93)

Human Factors in the Attainment of Reliability—R. S. Lincoln (p. 97)

Estimation from Life Test Data—Benjamin Epstein (p. 104)

A Customer Looks at the Reliability Program Activities—H. R. Powell (p. 108)

Correction to "Module Prediction"—George Hauser (p. 112)

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Electronic Technology*, (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these papers, 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.

Acoustics and Audio Frequencies...... 1517
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The Index to the Abstracts and References published in the PROC. IRE from February, 1959 through January, 1960 is published by the PROC. IRE, June, 1960, Part II. It is also published by *Electronic Technology* (incorporating *Wireless Engineer* and *Electronic and Radio Engineer*) and included in the April, 1960 issue of that Journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

UDC NUMBERS

Certain changes and extensions in UDC numbers, as published in PE Notes up to and including PE 666, will be introduced in this and subsequent issues. The main changes are:

Artificial satellites:	551.507.362.2	(PE 657)
Semiconductor devices:	621.382	(PE 657)
Velocity-control tubes, klystrons, etc.:	621.385.6	(PE 634)
Quality of received signal, propagation conditions, etc.:	621.391.8	(PE 651)
Color television:	621.397.132	(PE 650)

The "Extensions and Corrections to the UDC," Ser. 3, No. 6, August, 1959, contains details of PE Notes 598–658. This and other UDC publications, including individual PE Notes, are obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1, England.

ACOUSTICS AND AUDIO FREQUENCIES

534.2–14:534.88 2207

Acoustic Intensity Fluctuations and Temperature Microstructure in the Sea—F. H. Sagar. (*J. Acoust. Soc. Am.*, vol. 32, pp. 112–121; January, 1960.) Results of experiments designed to test the applicability of the Mintzer fluctuation formula (see, *e.g.*, 2275 of 1954) are given.

A list of organizations which have available English translations of Russian journals in the electronics and allied fields appears at the end of the Abstracts and References section.

534.21–14 2208

Study of Acoustic Propagation in a Two-Layered Model—R. K. Eby, A. O. Williams, Jr., R. P. Ryan and P. Tamarkin. (*J. Acoust. Soc. Am.*, vol. 32, pp. 88–99; January, 1960.) Measurements have been made in the range 55–600 kc using a small probe in a water layer 0.5–2 cm deep, over a 2-inch slab of rubber of greater density.

534.21–6–14 2209

Guided Waves in a Fluid with Continuously Variable Velocity Overlying an Elastic Solid: Theory and Experiment—I. Tolstoy. (*J. Acoust. Soc. Am.*, vol. 32, pp. 81–87; January, 1960.) Satisfactory agreement between theory and experiment has been secured for the propagation of sound waves in the range 10–20 cps from a point source in a sedimentary layer 607 meters thick, on a solid elastic base.

534.22–14:534.88 2210

Formulas for the Computation of Sound Speed in Sea Water—K. V. Mackenzie. (*J. Acoust. Soc. Am.*, vol. 32, pp. 100–104; January, 1960.) Equations, designed to agree with Kuwahara's tables, are presented for the computation of sound speed.

534.23 2211

Acoustic Investigations on Boundary Surfaces Between Solid and Liquid Substances—R. Krause. (*Z. angew. Phys.*, vol. 11, pp. 149–155; April, 1959.) Existing theory for sound transmission through layers of different media, assuming smooth boundary surfaces and no attenuation, is tested experimentally. Tests are also made for grooved boundary surfaces and a theory covering absorption losses is proposed.

534.24–14 2212

Reflection of Sound from Randomly Rough Surfaces—J. M. Proud, Jr., R. T. Beyer and P. Tamarkin. (*J. Appl. Phys.*, vol. 31, pp. 543–552; March, 1960.) An investigation of reflection of underwater sound from nonperiodic surfaces is reported. The dependence of the specular intensity upon various parameters is discussed and a method of deducing the surface properties from the intensity distribution is illustrated.

534.26–14:534.88 2213

Measurements of the Backscattering of Underwater Sound from the Sea Surface—G. R. Garrison, S. R. Murphy and D. S. Potter. (*J. Acoust. Soc. Am.*, vol. 32, pp. 104–111; January, 1960.) Measurements were made at 60 kc in an effort to relate the strength of the scattered sound to the character of the surface of the sea.

534.75 2214

Interaural Noise Correlations: Examination of Variables—I. Pollack and W. Trittipoe. (*J. Acoust. Soc. Am.*, vol. 31, pp. 1616–1618; December, 1959.) An extension of earlier work (1098 of 1960).

534.782 2215

Toward a Model for Speech Recognition—K. N. Stevens. (*J. Acoust. Soc. Am.*, vol. 32, pp. 47–55; January, 1960.) An approach to the design of a machine for the recognition and synthesis of speech is proposed, with particular emphasis on problems of acoustic analysis.

534.79 2216

Sound Intensity, Relative Loudness and Summation Loudness—R. Feldtkeller, E. Zwicker and E. Port. (*Frequenz*, vol. 13, pp. 108–117; April, 1959.) Various methods of measuring sound intensity and loudness are reviewed and the considerable discrepancies between the results are discussed.

534.8–8 2217

Very-High-Energy Ultrasonics—E. A. Neppiras. (*Brit. J. Appl. Phys.*, vol. 11, pp. 143–150; April, 1960.) Effects and applications are tabulated and the choice of transducer and the design of velocity transformers, are considered.

534.88:534.417 2218

Standard Calibration Hydrophone—C. C. Sims, (*J. Acoust. Soc. Am.*, vol. 31, pp. 1676–1680; December, 1959.) The hydrophone described has a piezoelectric element of lithium sulphate and operates in the range 5 cps–150 kc.

534.88:621.396.677.3 2219

Design of Directional Arrays—J. L. Brown, Jr. and R. O. Rowlands. (*J. Acoust. Soc. Am.*, vol. 31, pp. 1638–1643; December, 1959.) The design of linear and nonlinear arrays for determining the direction of a radiating source is considered theoretically with reference to SNR, directivity and minimization of error.

621.395.61.089.6 2220

Calibration of Microphones at Low Frequencies—G. Berry. (*Engineering (London)*, vol. 188, p. 163; September 11, 1959.) The design of a square-section acoustic duct of total length 35 ft is described. The duct is used for calibrating microphones in the range 30–320 cps. A method of comparing the sensitivities of two microphones is outlined.

621.395.625.3 2221

Contribution on the Stabilization of Tape Tension in Magnetic-Tape Recorders—H. Völz. (*Nachrichtentech. Z.*, vol. 9, pp. 181–186;

April, 1959.) Various methods of maintaining constant tape tension are discussed and compared in terms of the available controlling factors, such as tape weight, diameter of windings, and speed of spool rotation.

621.395.92 2222

On the Influence of the Diffraction of Sound Waves Around the Human Head on the Characteristics of Hearing Aids—C. Wansdronk. (*J. Acoust. Soc. Am.*, vol. 31, pp. 1609–1612; December, 1959.) A method is described for measuring the effect of sound diffraction on the performance of three types of hearing aid worn near the ear. Results show a large difference between hearing aids, probably depending on the sound entry hole and its distance from the skin, but no significant difference between persons.

ANTENNAS AND TRANSMISSION LINES

621.372.21:621.315.212 2223

Synthesis of Strictly Coaxial Networks—O. Maggi and M. Soldi. (*Alta Frequenza*, vol. 28, pp. 155–192; April, 1959.) The theory and procedure discussed are concerned with uniform transmission-line networks of distributed parameters, and are related to the work of Richards (1882 of 1949) and Ozaki and Ishii (IRE TRANS. ON CIRCUIT THEORY, vol. CT-2, pp. 325–336; December, 1955).

621.372.22 2224

A Note on the Optimum Design of Nonuniform Transmission Lines—L. Solymar. (*Proc. IEE*, vol. 107, pp. 100–104; March, 1960.) A design method is given for minimizing the complexity of the characteristic-impedance function. Numerical examples show that it is possible to design tapers even shorter than the Tchebycheff type.

621.372.829 2225

Influence of the Magneto-Dielectric Medium on Electromagnetic Wave Propagation in a Helix Waveguide Located in a Magneto-Dielectric—B. M. Bulgakov and V. P. Shestopalov. (*Izv. Akad. Nauk S.S.S.R., O.T.N.—Energetika i Avtomatika*, pp. 166–176; July/August, 1959.) Mathematical analysis of the dispersion properties of a helix within a stratified isotropic and axially symmetric magnetic dielectric medium. Graphical-analytic solutions of the dispersion equations are obtained and the impedance, wave modes and losses for different permittivity/permeability ratios are calculated.

621.372.852.1 2226

Waveguide Filters with Sharpened Cut-Off Characteristics—F. Künemund. (*Frequenz*, vol. 13, pp. 97–102; April, 1959.) Design examples are given for multistage rectangular-waveguide filters with sharp cut-off for operation at about 4 kc.

621.396.67 2227

Communal Aerials—E. Düniss and K. E. Müller. (*Nachrichtentech. Z.*, vol. 9, pp. 165–173; April, 1959.) Review of design problems concerning the antenna system, the antenna amplifier and the distribution network for communal installations covering long-, medium-, and short-wave, VHF and television reception.

621.396.67:621.397.61 2228

Tunable Diplexers of Constant Input Impedance with Residual-Sideband Filters for Television Picture and Sound Transmitters—Holle. (See 2542 of 1960.)

621.396.67:624.074 2229

Aerial Supporting Structures—P. J. Ward. (*Point to Point Telecommun.*, vol. 4, pp. 24–44; February, 1960.) Basic considerations involved in the design, erection and maintenance of masts and towers are discussed.

621.396.67.062.8 2230

Aerial Exchanges at Admiralty W/T Station—(*Engineer, London*), vol. 209, pp. 258–259; February 12, 1960.) An installation is described for connecting any one of several transmitters to different antennas by means of motor-driven carriages running in a framework of horizontal and vertical guides.

621.396.673 2231

Earthed-Base Medium-Wave Radiators of Extended Bandwidth, and an Appropriate Matching Circuit—H. Scholz. (*Rundfunktech. Mitt.*, vol. 3, pp. 97–101; April, 1959.) The advantages are summarized of earthing the base of an antenna mast which supports television and VHF antennas at the top and also acts as medium-wave radiator. The feed and matching arrangements of a 100-meter mast of this type are described and the characteristics of the medium-wave antenna system given.

621.396.676:629.19 2232

A Broad-Band Spherical Satellite Antenna—H. B. Riblet. (PROC. IRE, vol. 48, pp. 631–635; April, 1960.) Design and performance data are given for an equi-angular spiral slot antenna system surrounding the spherical surface of a TRANSIT satellite for frequencies 54, 108, 162 and 216 mc. Peak-to-peak field intensity variations <10 db are achieved.

621.396.677:621.396.96 2233

Nonreciprocal Radar Antennas—R. L. Mattingly. (PROC. IRE, vol. 48, pp. 795–796; April, 1960.) The side-lobe content of the antenna pattern may be reduced by the use of auxiliary antennas.

621.396.677.3:534.88 2234

Design of Directional Arrays—Brown, Jr. and Rowlands. (See 2219 of 1960.)

621.396.677.3.012:681.142 2235

An Analogue Computer for Investigating the Directivity Characteristics of Complex Arrays of Unit Aerials—G. Mitchell. (*P.O. Elect. Engrg. J.*, vol. 52, pp. 246–250; January, 1960.) Motor-driven and magnetically operated linear and cosine potentiometers form the basis of the computer described, which is designed for calculations on any array with 50–200 unit antennas arranged along diametral rows intersecting at a common point, and with a maximum of fourteen antennas in any one row. The computer can also be used on a semi-automatic basis for dealing with larger or more complex arrays.

621.396.677.32 2236

Recent Developments in Very-Broad-Band Endfire Arrays—A. F. Wickersham, Jr. (PROC. IRE, vol. 48, pp. 794–795; April, 1960.) A tapered Yagi antenna excited via a transmission line can be made to have wide-band matching characteristics.

621.396.677.7 2237

The Radiation of Circular-Cylinder Waveguides Obliquely Mounted in a Metallic Screen—L. Breitenhuber. (*Z. Phys.*, vol. 155, pp. 441–452; July 16, 1959.) Waveguide radiators terminating at an angle α in a metal screen of infinite extent are investigated. Radiation polar diagrams in the E and H planes are given for the TE_{11} mode for $\alpha=0$, 30 and 60° and waveguide diameter $=0.75\lambda$.

621.396.677.73 2238

X-Band Horn Antenna has Broad Beam-Width—R. E. Metter. (*Electronics*, vol. 33, pp. 50–53; March 4, 1960.) Description of an elliptically polarized antenna with 3-db and 6-db beam-widths of 140° for orthogonal planes of polarization.

AUTOMATIC COMPUTERS

681.142 2239

An Eight-Digit Word Generator: Parts 1 & 2—P. L. Owen and T. R. H. Sizer. (*Electronic Engrg.*, vol. 32, pp. 134–139, 212–217; March/April, 1960.) The theory of operation and the design of a binary word generator using surface barrier transistors and a direct coupling technique are given.

681.142 2240

Solving Noise Problems in Digital Computer Memories—A. H. Ashley and E. U. Cohler. (*Electronics*, vol. 33, pp. 72–74; March 25, 1960.) Details are given of a coincident-current storage system in which strobe pulses are obtained directly from driven cores.

681.142:535 2241

Optical Analogue Computers—B. J. Howell. (*J. Opt. Soc. Amer.*, vol. 49, pp. 1012–1021; October, 1959.) A general theory of optical analog computers based on characteristics of the photographic process is developed, and applications of photosensitive materials and photocells for performing mathematical operations are reviewed. 62 references.

681.142:621.318.4 2242

Integrated Magnetic Circuits for Synchronous Sequential Logic Machines—U. F. Gianola (*Bell Syst. Tech. J.*, vol. 39, pp. 295–332; March, 1960.) Two approaches to the problems of storage, gain, and unidirectional transfer of a sequence of bits in a system with a minimum of nonmagnetic components are outlined. Examples are given of experimental synchronous sequential circuits using commercially available multiaperture cores.

681.142:621.318.57:621.382.3 2243

Computer Switching with High-Power Transistors—J. S. Ronne. (*Electronics*, vol. 33, pp. 44–47; March 4, 1960.) A method for selecting the most suitable type of power transistor for particular switching applications is given.

681.142:621.387 2244

The Digitron: A Cold-Cathode Character Display Tube—McLoughlin, Reaney and Turner. (See 2576 of 1960.)

681.142:621.396.677.3.012 2245

An Analogue Computer for Investigating the Directivity Characteristics of Complex Arrays of Unit Aerials—Mitchell. (See 2235 of 1960.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.013.1:621–52 2246

An Approximation to the Harmonic Response of Saturating Devices—R. J. Kavanagh. (*Proc. IEE*, vol. 107, pp. 127–133; March, 1960.) An exponential curve is used to approximate the system characteristics.

621.315.5.011.2 2247

Electrical and Thermal Relaxation Effects in Current-Carrying Metallic Conductors—H. Göddecke. (*Z. angew. Phys.*, vol. 11, pp. 143–147; April, 1959.) The ac admittance of conductors carrying dc is calculated and a relation is found for determining the frequency-dependent thermal conductivity. Results of measurements on lamp filaments confirm this relation. [See also 676 of 1956 (Burgess).]

621.319.42 2248

Some Results on the Cross-Capacitances per Unit Length of Cylindrical Three-Terminal

Capacitors with Thin Dielectric Films on their Electrodes—D. G. Lampard and R. D. Cutkosky. (*Proc. IEE*, vol. 107, pp. 112–119; March, 1960.)

621.372.4:621.3.016.2 2249
The Concept of Power at Complex Frequences—G. Schilling. (*Frequenz*, vol. 13, pp. 70–74; March, 1959.) The significance of the term "negative resistance" is discussed with regard to two different aspects. One is the behavior of two-pole networks with sinusoidal oscillators and a negative-slope V/I characteristic, the other is the case of exponentially decaying oscillations where phase-shifts greater than 90° between current and voltage may arise.

621.372.5 2250
A Method of Synthesis of Linear Passive Networks for Transient Conditions—M. L. D'Atri. (*Note Recensioni Notiz.*, vol. 8, pp. 155–168; March/April, 1959.) A method of synthesis is described for networks with transient response of unit-function or Dirac-function type.

621.372.512.3 2251
Frequency Jumps and Pulling Effects in Coupled Circuits—W. S. Ehrenberg. (*Bull. schweiz. elektrotech. Ver.*, vol. 50, pp. 1009–1016; October 10, 1959.) The admittance equations for two-coupled tuned circuits are formulated and the conditions of oscillation established. The minimum width of the frequency jump is derived as a function of coupling and in relation to the "pulling" properties of the oscillator. The limiting conditions for freedom from frequency jumps are given.

621.372.54:621.391 2252
The Optimization of a Class of Nonlinear Filters—J. K. Lubbock. (*Proc. IEE*, pt. C, vol. 107, no. 11, pp. 60–74.) A mathematical treatment shows that use of the nonlinear filter gives a significant improvement in mean-square error even when signal and noise have the same spectral densities. Optimization is not unduly difficult.

621.372.632:621.382.2 2253
The Influence of the Time-Dependent Series Resistance of a Capacitance Diode in a Mavar Up-Converter—P. Bobisch and C. Sondhauss. (*J. Electronics Control*, vol. 7, pp. 344–366; October, 1959. In German.) An amplifying frequency converter for conversion from 70 to 4000 mc is investigated theoretically. The variable reactance is a semiconductor diode of the type described by Uhlir (2905 of 1958). The effect of the loss resistance of this diode on converter stability and bandwidth is evaluated.

621.373.421.11 2254
Dual-Frequency Oscillator Design—L. Klienberg. (*Electronics*, vol. 33, pp. 182–184; March 11, 1960.) Basic design details are given of circuits in which Hartley and Colpitts oscillators are combined.

621.373.43 2255
Simultaneous Oscillation at Three Natural Frequencies in Nonlinear Feedback Circuits—V. Met. (*Arch. elekt. Übertragung*, vol. 13, pp. 161–170; April, 1959.) The steady states of a nonlinear feedback loop with three degrees of freedom are investigated under the assumption of quasilinearity, with reference to the theory of multimode oscillators given in 3428 of 1957. Experimental results and practical applications of the effect are discussed.

621.373.431.1:621.382.3:621.318.57 2256
Transfluxor Oscillator Gives Drift-Free Output—R. J. Sherin. (*Electronics*, vol. 33, pp. 48–49; March 4, 1960.) A voltage-controlled transistor multivibrator circuit incorporating a transfluxor magnetic core is described; a feature of the arrangement is the stability of the output frequency after removal of the control signal.

621.373.44:621.382.3:621.397 2257
A Simple Pulse Generator and a Pulse Distribution Unit Using Transistors—H. Stierhof. (*Rundfunktech. Mitt.*, vol. 3, pp. 81–90; April, 1959.) Circuit and performance details are given of equipment for use in television studio installations.

621.374.3:621.387.4 2258
Time to Pulse-Height Converter of Wide Range—J. Fischer and A. Lundby. (*Rev. Sci. Instr.*, vol. 31, pp. 10–14; January, 1960.) The converter is designed to plug in to commercial oscilloscopes and can time several events per (linear) time scan relative to one starting event. Counting losses can be reduced by "negative-time scanning." The circuit diagram is given.

621.374.3:621.387.4 2259
Fast Coincidence Circuit for Slow Pulses—J. E. Draper and A. A. Fleischer. (*Rev. Sci. Instr.*, vol. 31, pp. 49–52; January, 1960.) For pulse timing, particularly in coincidence spectroscopy applications.

621.374.4:621.317.7.029.64 2260
Practical Decimetre-Wavelength Harmonic Generator—U. Adelsberger. (*Arch. elekt. Übertragung*, vol. 13, pp. 152–156; April, 1959.) A frequency-stable coaxial-type harmonic generator is described whose design is based on the rod wavemeter (2494 of 1956). Four patterns are available covering different fundamental-wavelength ranges; the highest usable harmonic frequency is about 4500 mc.

621.374.4:621.372.44 2261
Harmonic Generation using Idling Circuits—I. Kaufman and D. Douthett. (PROC. IRE, vol. 48, pp. 790–791; April, 1960.) A theoretical examination of the conditions necessary for efficient harmonic generation using an element with a square-law characteristic.

621.375.12:621.382.23 2262
Tunnel D ode as an Interstage Gain Device—L. A. LoSasso. (PROC. IRE, vol. 48, pp. 793–794; Apr l, 1960.) Application of the tunnel diode as a coupling element, to provide impedance transformation and power gain between each stage of an IF amplifier.

621.375.127.018.78 2263
Distortion in Class-AB Push-Pull Amplifiers—I. S. Docherty and R. E. Aitchison. (*Proc. IRE (Australia)*, vol. 20, pp. 737–741; December, 1959.) Fourier coefficients of the current waveform for a single tube or transistor working under varying bias conditions are evaluated as a function of the angle of flow of the output current. From these results the harmonic distortion is determined for a balanced push-pull system and it is shown that small degrees of unbalance can be responsible for large contributions to the over-all distortion, principally by the introduction of second harmonics.

621.375.222 2264
Compensation of Direct-Coupled Amplifiers Against Drift Caused by Heater Voltage Fluctuations—F. Gutmann. (*Proc. IRE (Australia)*, vol. 20, pp. 692–694; November, 1959.) The output from a thermopile, heated from the source which supplies the tube heaters of an amplifier, is fed back either to the input or to the indicator to compensate for zero drifts. The drift is reduced by two-thirds without loss of amplifier sensitivity.

621.375.227 2265
Self-Balancing Push-Pull Circuits—D. R. Birt. (*Wireless World*, vol. 66, pp. 223–227, 283–285; May/June, 1960.) Theoretical aspects of self-balancing circuits are discussed and a method is described for making a push-pull amplifier self-balancing by applying over-all push-push negative feedback: no close-tolerance components are needed.

621.375.029.3 2266
The New "Isodyne" Phase Splitter—E. F. Worthen. (*Audio*, vol. 42, pp. 26–27; August, 1958.) Details are given of an amplifier in which the phase splitter is directly coupled and the output to the final stage is taken from a cathode follower. Measured intermodulation distortion for the complete amplifier was less than 0.2 per cent for outputs up to 30 w with a frequency response level to within ±0.5 db up to 10 kc.

621.375.232 2267
Unity-Gain Amplifier Offers High Stability—G. M. Davidson and R. F. Brady. (*Electronics*, vol. 33, pp. 66–67; February 26, 1960.) Design equations are given for a double-triode output stage, the gain stability factor of which is the reciprocal of the products of the two gains.

621.375.232.018.78 2268
An Overmodulation Effect in Negative-Feedback Valve Amplifiers—F. Feil and K. Lang. (*Frequenz*, vol. 13, pp. 65–70; March, 1959.) If the final stage of a multistage *RC*-coupled feedback amplifier is driven into grid current through overloading, there is a difference between the input levels at which grid current sets in and that at which it ceases. The distortion caused by this effect is investigated; measures to eliminate it are discussed but are not considered worth undertaking for multichannel carrier systems.

621.375.4 2269
Temperature Stabilization of Transistors in Class-B Amplifiers—K. L. Webber. (*Proc. IRE (Australia)*, vol. 20, pp. 726–733; December, 1959.) Factors affecting stability are discussed and a graphical method is described for estimating the maximum temperature for stable operation. Experimental results obtained using a thermistor to improve stability are in good agreement with theory.

621.375.4.012 2270
Transistor Circuit Design Using Modified Hybrid Parameters—E. R. Aitchison. (*Proc. IRE (Australia)*, vol. 20, pp. 673–679; November, 1959.) Methods are described for calculating modified hybrid parameters and approximate expressions applicable to the design of practical one- and two-stage amplifiers with feedback.

621.375.9:[538.569.4+621.372.44 2271
Masers or Parametric Amplifiers—D. C. Lainé. (*Electronic Technologist*, vol. 37, pp. 174–185; May, 1960.) The main features of the various types of amplifiers in each group are summarized and their relative performance in practical systems is discussed. 30 references.

621.375.9.:538.569.4 2272
On the Theory of Masers—E. Groschwitz. (*Z. Naturforsch.*, vol. 14a, pp. 305–307; March, 1959.) Preliminary note dealing with the effects of phase distribution in the microwave field.

621.375.9:538.569.4 2273
On the Possibility of Maser Action in Nuclear Quadrupole Systems—Donovan and Vuylsteke. (See 2315 of 1960.)

621.375.9:538.569.4 2274
Operation of a Chromium-Doped Titania Maser—H. J. Gerritsen and H. R. Lewis. (*J. Appl. Phys.*, vol. 31, pp. 608–609; March, 1960.) Brief description of a maser which has been operated at several frequencies in the range 8.2–10.6 kmc (pump frequency 35 kmc) and at 22.3 kmc (pump frequency 49.9 kmc).

621.375.9:621.372.44:621.372.2 2275
Parametric Amplification Along Nonlinear Transmission Lines—R. Landauer. (*J. Appl. Phys.*, vol. 31, pp. 479–484; March, 1960.) A pump signal propagating along a dispersionless transmission line with a distributed nonlinear capacitance is subject to deformation, since different parts of the signal move with different velocities. The deformation will affect the parametric amplification process, so that, in general, a sinusoidal signal will not be increased in its fundamental frequency by traveling down the line.

621.375.9:621.372.44:621.385.6 2276
Extension of Longitudinal-Beam Parametric-Amplifier Theory—Sobol. (See 2566 of 1960.)

621.375.9:621.372.44:621.391.822 2277
Note on the Noise Figure of Negative-Conductance Amplifiers—A. van der Ziel and J. Tamiya. (PROC. IRE, vol. 48, p. 796; April, 1960.) The noise figure is calculated for the case when the negative-conductance amplifier stage is connected directly to a receiver and when it is connected via a lossless step-down transformer.

621.375.9:621.382.23 2278
Noise Performance Theory of Esaki (Tunnel) Diode Amplifiers—M. E. Hines and W. W. Anderson. (PROC. IRE, vol. 48, p. 789; April, 1960.) Theory indicates that low noise figures can be achieved only with some sacrifice in gain, stability or bandwidth. [See also 3246 of 1959 (Chang).]

621.376.22:539.23 2279
Thin-Film Balanced Modulator—(*Electronics*, vol. 33, pp. 78, 80; February 26, 1960.) Passive balanced modulator circuits are evolved using single-domain permalloy film which is vacuum-deposited in a magnetic field. Tests have been made using a 4-mc carrier with sinusoidal modulation at 20 cps–20 kc and square-wave modulation up to 100 kc.

621.376.223:621.375.024:681.142 2280
The Use of Silicon Diodes in D.C. Modulators and their Applications to Drift Correctors for Computing Amplifiers—T. Glucharoff and C. P. Gilbert. (*Proc. IEE*, vol. 107, pp. 82–90; March, 1960.) Advantages over the conventional relay modulator include good high-frequency response, low noise level, low switching power requirements, and an operational life limited only by the thermionic tubes. Zero stability is comparable with that of the conventional relay.

621.376.3 2281
Explicit Form of F.M. Distorition Products with White-Noise Modulation—R. G. Medhurst. (*Proc. IEE*, vol. 107, pp. 120–126; March, 1960.) Expressions are derived for frequency and amplitude distortion of the output of a network whose phase or amplitude characteristics vary nonlinearly with frequency. The evaluation of these expressions using a digital computer is illustrated for the case of fourth-order distortion associated with terms in the discriminator amplitude characteristic up to the sixth degree.

621.376.32:621.318.4 2282
Frequency Modulation of a Transmitter by Inductance Variation—W. Moortgat-Pick. (*Frequenz*, vol. 13, pp. 117–120; April, 1959.) The use of a ferrite-cored variable inductor is proposed and its advantages over a reactance-tube circuit are discussed. An experimental FM master oscillator is described.

621.376.4 2283
The Single-Ended Diode Phase-Sensitive Detector—R. Chidambaram and S. Krishnan. (*Electronic Engrg.*, vol. 32, pp. 158–160; March, 1960.) The transfer ratios for the two diodes are found to vary with signal level. The nonlinearity in the output is evaluated and a table is given for assessing the performance of a given detector. (See also 443 of 1960.)

GENERAL PHYSICS

535.2 2284
Theoretical Considerations on the Experimental Determination of Spontaneous Photon Fluctuations—G. A. Spescha and M. J. O. Strutt. (*Helv. Phys. Acta*, vol. 33, pp. 53–68; March 15, 1960. In German.) A method is indicated for measuring fluctuations in the range $h\nu/kT \gg 1$ with a suitable photon detector. (See also 2559 of 1960.)

535.215 2285
Theory and Application of Thermally Stimulated Currents in Photoconductors—R. R. Haering and E. N. Adams. (*Phys. Rev.*, vol. 117, pp. 451–454; January 15, 1960.) Explicit solutions are obta ned in the limits of slow and fast retrapping. A method of obtaining the ionization energy of the relevant traps is discussed.

535.215 2286
Photon Momentum Effects in the Magneto-Optics of Excitons.—J. J. Hopfield and D. G. Thomas. (*Phys. Rev. Lett.*, vol. 4, pp. 357–359; April 1, 1960.) Two magneto-optic effects due to the small but finite wave vector of light are described.

535.317.6:621.397.61:681.4 2287
An Approximation Method for the Treatment of Spherical Aberration of Objectives Using the Methods of Communication Theory—F. Below and H. Grabke. (*Rundfunktech. Mitt.*, vol. 3, pp. 94–96; April, 1959.) The surface of the objective is divided into zones and a simple formula is derived for evaluating the brightness distribution in the image plane.

537.2 2288
Charge Sphere in Cylinder—W. R. Smythe. (*J. Appl. Phys.*, vol. 31, pp. 553–556; March, 1960.) The charge density, potential, and capacitance of the spere are calculated to an accuracy within 1 part in 10^7 using the method described earlier (73 of 1957).

537.291 2289
Wave Functions and Effective Hamiltonian for Bloch Electrons in an Electric Field—G. H. Wannier. (*Phys. Rev.*, vol. 117, pp. 432–439; January 15, 1960.)

537.311.31 2290
Transport Phenomena in Elastically Anisotropic Metals—J. Appel. (*Z. Naturforsch.*, vol. 14a, pp. 379–393; April, 1959.)

537.311.33 2291
Statistical Theory of Kinetic Phenomena: Part 2—M. I. Klinger. (*Fiz. Tverdogo Tela.*, vol. 1, pp. 1225–1238; August, 1959.) General expressions are derived for kinetic coefficients which are used to obtain approximate formulas for the case of weak electron-phonon interaction. The derived formulas are used for investigating the thermoconductivity and thermo-EMF of semiconductors. (For earlier work see *Zh. tekh. Fiz.*, vol. 27, pp. 2780–2783; December, 1957.)

537.311.33 2292
A Dielectric Approach to Impurity Conduction—D. G. H. Frood. (*Proc. Phys. Soc.*, vol. 75, pp. 185–193; February 1, 1960.) The problem of impurity conduction in valence crystals is formulated on a dielectric basis. Calculations are made of 1) the contribution to the permittivity arising from the hydrogenic donor atoms, 2) the donor activation energy, and 3) the effect of screening by conduction electrons.

537.312.62 2293
On the Theory of Superconductivity—Y. Wada and N. Fukuda. (*Prog. Theoret. Phys.*, vol. 22, pp. 775–806; December, 1959.) An exact treatment of the strong-coupling approximation in the theory of Bardeen *et al.* (1386 of 1958) and discussion of the Meissner effect.

537.523.4 2294
Development of Space Change and Growth of Ionization in the Transient Townsend Discharge—Y. Miyoshi. (*Phys. Rev.*, vol. 117, pp. 355–365; January 15, 1960.) An extension of earlier work (730 of 1957).

537.525 2295
Cathode Work Function, Sparking Potentials and Secondary Ionization Coefficients for Oxide-Coated Cathodes in Hydrogen—D. E. Davies and B. J. Hopkins. (*Brit. J. Appl. Phys.*, vol. 10, pp. 498–501; November, 1959.) The contact potential difference between oxide-coated cathodes and a gold reference surface has been measured using the Kelvin technique. (See 2296 of 1960.)

537.525 2296
Influence of the Cathode Work Function on the Sparking Potential in Hydrogen—D. E. Davies and R. K. Fitch. (*Brit. J. Appl. Phys.*, vol. 10, pp. 502–505; November, 1959.) The Kelvin vibrating-electrode method of measuring contact potential differences has been used to follow changes in work function of evaporated metallic films in a parallel-plate electrode system in hydrogen at a pressure of 100 mm Hg.

537.525 2297
Ultimate and Secondary Electron Energies in the Negative Glow of a Cold-Cathode Discharge in Helium—J. M. Anderson. (*J. Appl. Phys.*, vol. 31, pp. 511–515; March, 1960.)

537.56 2298
Equation of State of High-Temperature Plasma—T. Morita. (*Prog. Theoret. Phys.*, vol. 22, pp. 757–774; December, 1959.) Fully ionized plasmas at high temperatures are investigated on the basis of quantum theory and a correction to the Debye-Hückel approximation is evaluated for the free energy and the equation of state.

537.56 2299
Electron and Ion Runaway in a Fully Ionized Gas: Part 2—H. Dreicer. (*Phys. Rev.*, vol. 117, pp. 329–342; January 15, 1960.) An extension of the treatment presented in Part 1 (425 of 1960) to give a more exact estimate of the runaway rate under the action of a weak electric field.

537.56 2300
Electron Velocity Distributions in a Partially Ionized Gas—H. Dreicer. (*Phys. Rev.*, vol. 117, pp. 343–354; January 15, 1960.) Investigation of the transition from the non-Maxwellian distribution characteristic of a poorly ionized gas to the Maxwellian distribution as the degree of ionization rises.

537.56:538.56 2301
Plasma Containment by R.F. and D.C. Field Combinations—D. G. Dow and R. C.

Knechtli. (*J. Electronics Control*, vol. 7, pp. 316–343; October, 1959.) The advantages are discussed of supplementing dc with RF fields, which are sometimes more effective and possibly more stable. The design of "RF plugs" for sealing the ends of a plasma column contained by a dc magnetic field is given in some detail.

537.56:538.56 2302

Propagation of Plasma Waves Across a Density Discontinuity—A. H. Kritz and D. Mintzer. (*Phys. Rev.*, vol. 117, pp. 382–386; January 15, 1960.) When longitudinal waves are incident on a plasma density (or temperature) discontinuity energy can be converted to transverse waves. The inverse process also occurs.

537.56:538.561 2303

Radio Emission by Plasma Oscillations in Nonuniform Plasmas—D. A. Tidman. (*Phys. Rev.*, vol. 117, pp. 366–374; January 15, 1960.) Equations of motion for small-amplitude plasma oscillations interacting with the em field in slowly varying density or temperature gradients are established. The RF noise excited by a wave packet of plasma oscillations traversing such gradients is calculated using the WKB approximation. A similar calculation is made for a density discontinuity.

537.56:538.561:538.6 2304

The Radiation Emitted within Ionized Gas in the Presence of a Magnetic Field—K. Kawabata. *Rep. Ionosphere Res. Japan*, vol. 12, pp. 428–436; 1958.)

537.56:538.6 2305

Electrohydromagnetic Waves in a Fully Ionized Gas: Part 1—K. D. Cole. (*Planet. Space Sci.*, vol. 1, pp. 319–324; September, 1959.) Transport equations are used to determine coefficients which are generalizations for any frequency of electric field of the parallel, Pedersen and Hall conductivities. Applications are discussed.

537.583:537.213 2306

Potential Distribution Between Two Plane Emitting Electrodes—P. A. Lindsay and F. W. Parker. (*J. Electronics Control*, vol. 7, pp. 289–315; October, 1959.) Equations are derived for two plane parallel electrodes made from different materials and kept at different temperatures. Numerical calculations are given for electrodes at the same temperature T with an external potential difference ϕ_{ext} between them. All possible potential distributions are shown to be represented by a family of curves, the parameter of the family being $A = \exp(-e\phi_{ext}/kT)$.

538.11 2307

Theory of Spin-Wave Interactions in Ferro- and Antiferro-Magnetism—T. Oguchi. (*Phys. Rev.*, vol. 117, pp. 117–123; January 1, 1960.) Mathematical treatment of spin wave theory for a ferromagnetic model, giving results in agreements with Dyson's (3696 of 1956).

538.3 2308

Geometrical Representation of the Maxwell Field in Minkowski Space—E. S. Lowry. (*Phys. Rev.*, vol. 117, pp. 616–618; January 15, 1960.) "The electromagnetic field tensor of a classical charged particle is associated with the orientation and density of a family of two-dimensional surfaces radially distributed about the world line of the particle in Minkowski space."

538.311 2309

On the Production of High-Intensity Magnetic Fields of Short Duration—P. Cotti. (*Z. angew. Math. Phys.*, vol. 11, pp. 17–32; January 25, 1960.) The design of capacitor discharge apparatus for pulsed fields is discussed. Calculations are compared with experimental data for coils producing fields of 400,000 oersteds for 1 msec in a volume of 0.5 cm³ and 100,000 oersteds for 100 msec in 10 cm³.

538.311:539.89 2310

Production of Very High Magnetic Fields by Explosion—C. M. Fowler, W. B. Garn and R. S. Caird. (*J. Appl. Phys.*, vol. 31, pp. 588–594; March, 1960.) Fields in the 10–15 megagauss range are obtained by use of high explosives which effectively compress the flux inside a metallic cylinder.

538.566.029.6:537.56 2311

Microwave Whistler-Mode Propagation in a Dense Laboratory Plasma—R. M. Gallet, J. M. Richardson, B. Wieder, G. D. Ward and G. N. Harding. (*Phys. Rev. Lett.*, vol. 4, pp. 347–349; April 1, 1960.) The thermonuclear machine ZETA at Harwell is being used for plasma diagnostics. Measurements of refractive index and attenuation rate at two frequencies will be attempted.

538.569.4:621.375.9 2312

Focusing Molecular Beams of NH_3—J. C. Helmer, F. B. Jacobus and P. A. Sturrock. (*J. Appl. Phys.*, vol. 31, pp. 458–463; March, 1960.) A study of the formation of molecular beams for use in NH_3 masers is described. The design and performance of nonuniform focusers and of effusers is considered. The operation of a system is described in which lower-state molecules produced by maser oscillation may be detected.

538.569.4:621.375.9 2313

Molecular Beam Formation by Long Parallel Tubes—J. A. Giordmaine and T. C. Wang. (*J. Appl. Phys.*, vol. 31, pp. 463–471; March, 1960.) The characteristics of molecular beams formed by sources consisting of long-tube arrays are measured for several sources. The results agree well with theory. Some considerations in source design are also discussed.

538.569.4:621.375.9 2314

Analysis of Focuser for Maser Oscillators—M. Hirono. (*J. Radio Res. Labs. Japan*, vol. 6, pp. 515–532; July, 1959.) A mathematical study of the characteristics of eight-pole, four-pole and square-well focusers. The main conclusions refer to the focusing length and the influence of the velocity distribution in the beam.

538.569.4:621.375.9 2315

On the Possibility of Maser Action in Nuclear Quadrupole Systems—R. E. Donovan and A. A. Vuylsteke. (*J. Appl. Phys.*, vol. 31, pp. 614–615; March, 1960.) Analysis of an apparently promising system shows that the gain in the material is not sufficient to overcome circuit losses.

538.691 2316

Allowed and Forbidden Regions in the Motion of Electrically Charged Particles in Axially Symmetric Magnetic Fields—H. Fisser and R. Kippenhahn. (*Z. Naturforsch.*, vol. 14a, pp. 37–46; January, 1959.) Diagrammatic methods are applied to the case of motion in the field of a magnetic dipole and of a circular line current.

538.691 2317

The Motion of Charged Particles in the Magnetic Field of a Straight Current-Carrying Wire—F. Hertweck. (*Z. Naturforsch.*, vol. 14a, pp. 47–54; January, 1959.) The calculated velocity of a particle in the field along an infinitely long wire is in agreement with Alfvén's approximation.

538.691 2318

Quantum Theory of Transport in a Magnetic Field—P. N. Argyres. (*Phys. Rev.*, vol. 117, pp. 315–328; January 15, 1960.)

539.12 2319

Comparison of the Charges of the Electron, Proton and Neutron—A. M. Hillas and T. E. Cranshaw. (*Nature*, vol. 186, pp. 459–460; May 7, 1960.) Experimental details are classified in reply to criticism by Bondi and Lyttleton. (See 117 of 1960.)

539.2:535.215 2320

Systematic Study of Electron Binding Energies of some Fourth- and Sixth-Period Elements by means of the Photo-Electron Method—E. Sokolowski. (*Ark. Fys.*, vol. 15, pp. 1–30; April 10, 1959.)

539.2:537.29 2321

Theory of Internal Field Emission—J. Kamphusmann. (*Z. Naturforsch.*, vol. 14a, pp. 165–171; February, 1959.) Difficulties in the solution of the Schrödinger equation in the quantum-mechanical treatment of internal field emission are avoided by the use of Bloch functions instead of those based on Houston's theory.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.14:538.69 2322

On the Magnetic Field of the Galaxy—F. Hoyle and J. G. Ireland. (*Monthly Notices Roy. Astron. Soc.*, vol. 120, pp. 173–186; 1960.)

523.152.3:538.12 2323

On the Interplanetary Gas and its Magnetic Field—L. Block. (*Ark. Fys.*, vol. 14, pp. 179–193; October 14, 1958.) Proposed models of the interplanetary magnetic field are critically examined in relation to known physical phenomena.

523.16 2324

Existence of Net Electric Charges on Stars—V. A. Bailey. (*Nature*, vol. 186, pp. 508–510; May 14, 1960.) The hypothesis is proposed that a star, like the sun of mass $M_s g$, carries a net negative charge $-Q_s$ which is given by the formula $Q_s = \beta_s G^{1/2} M_s$ esu where G is the constant of gravitation and β_s is a pure number of the order of 0.03. This hypothesis accounts for the known orders of magnitude and directions relating to a number of phenomena such as the magnetization of the earth and the outer Van Allen belt, and leads to a simple qualitative or semiquantitative explanation for others. Possible sources of the stellar net charge and possible criticisms of the hypothesis are discussed.

523.164 2325

The Simultaneous Observation of Radio Star Scintillations on Different Radio Frequencies—H. J. A. Chivers. (*J. Atmos. Terr. Phys.*, vol. 17, pp. 181–187; February, 1960.) With increasing scintillation activity, the amplitude and rate have distributions which contain discontinuities.

523.164.3:523.42 2326

Apparent Radio Radiation at 11-m Wavelength from Venus—J. D. Kraus. (*Nature*, vol. 186, p. 462; May 7, 1960). From further observation and subsequent analysis, it appears definite that signals previously reported (see 2118 of 1957) were not from Venus but were interference of terrestrial origin.

523.164.32:535.853 2327

The Freiburg Radio Spectrograph (48–165 Mc/s)—K. O. Kiepenheuer and H. H. Rabben.

(*Z. Astrophys.*, vol. 49, pp. 61–67; January 13, 1960.) The design of the solar spectrograph described is based on that of Wild and McCready (1629 of 1951).

523.164.32:551.594.5 2328

The Association of Solar Radio Bursts with Auroral Streams—A. Maxwell, A. R. Thompson and G. Garmire. (*Planet. Space Sci.*, vol. 1, pp. 325–332; September, 1959.) Analysis of two years' observations of Type II radio bursts show a 45 per cent correlation with auroral and magnetic storms, with a mean time delay of 33 *h*.

523.164.32:621.391.812.6 2329

Determination of the Extinction in the Earth's Atmosphere of Radiation at 20- and 3.2-cm Wavelength—F. Fürstenberg. (*Z. Astrophys.*, vol. 49, pp. 42–60; January 13, 1960.) Discussion of the results of microwave absorption measurements on solar radiation made at sunset and sunrise in Berlin and on Rügen in the Baltic.

523.164.34 2330

The Thermal Radiation of the Moon at 1420 Mc/s—P. G. Mezger and H. Strassl. (*Planet. Space Sci.*, vol. 1, pp. 213–226; August, 1959.) A radiation temperature of about 250° has been measured, showing no variation with the phase of the moon.

523.164.4 2331

The Radio Emission from Normal Galaxies: Part 1—Observations of M31 and M33 at 158 Mc/s and 237 Mc/s—R. H. Brown and C. Hazard. (*Monthly Notices Roy. Astr. Soc.*, vol. 119, pp. 297–308; 1959.)

523.165 2332

Changes in the Differential Rigidity Spectrum of Primary Cosmic Rays Associated with Long-Term and Short-Term Intensity Variations—J. R. Storey. (*Phys. Rev.*, vol. 117, pp. 573–577; January 15, 1960.)

523.165 2333

Some Properties of the Van Allen Radiation—A. J. Dessler and R. Karplus. (*Phys. Rev. Lett.*, vol. 4, pp. 271–274; March 15, 1960.) Observations of the electron belt made with Explorer IV and Explorer VI are inconsistent with the solar injection hypothesis; the electrons released in the decay of cosmic-ray neutron albedo may represent a satisfactory source for the outer zone.

523.165 2334

Measurement of Radiation in the Lower Van Allen Belt—F. E. Holly and R. G. Johnson. (*J. Geophys. Res.*, vol. 65, pp. 771–772; Feburary, 1960.) The maximum of the belt occurred at 15°N.

523.165:523.152.3 2335

Cosmic-Ray Intensity in Interplanetary Space—H. Elliot. (*Nature*, vol. 186, pp. 299–300; April 23, 1960.) Cosmic-ray intensity has been calculated as a function of distance from the sun in the region between the orbits of Mars and Venus.

523.165:538.12 2336

Cosmic-Ray Orbits in Interplanetary Magnetic Fields—L. Block. (*Ark. Fys.*, vol. 14, pp. 161–177; October 14, 1958. 43 plates.) Cosmic-ray orbits have been calculated in the interplanetary field suggested by Alfvén (2714 of 1956). The results are briefly discussed and compared with cosmic-ray observations.

523.5 2337

Combined Photographic and Radio Echo Observations of Meteors—J. Davis, J. S. Greenhow and J. E. Hall. (*Proc. Roy. Soc.* (*London*) vol. 253, pp. 121–129; November 17, 1959.) Measurements have been made on a bright Geminid meteor by means of a meniscus Schmidt camera and two pulsed radio transmitters operating at frequencies near 36 mc. The radio echo duration is found to be several orders of magnitude less than would be expected from simple diffusion theory. This behavior is explained in terms of the attachment of electrons to neutral oxygen molecules to form negative ions, and a value for the attachment coefficient is determined.

523.5:621.391.812.5 2338

The Azimuth Distribution of Oblique Reflections from Meteor Trails and its Relation to Meteor Radiant Distributions—W. C. Bain. (*J. Atmos. Terr. Phys.*, vol. 17, pp. 188–204; February, 1960.) The measurements refer to a 1740-km north-south path at frequencies of 37 and 70 mc. The distribution of radiants which best fits the observations is a heliocentric distribution uniform over the celestial sphere.

523.5:621.396.9 2339

The Effect of Attachment on Radio Echo Observations of Meteors—J. Davis, J. S. Greenhow and J. E. Hall. (*Proc. Roy. Soc.* (*London*) vol. 253, pp. 130–139; November 17, 1959.) The observed relation between visual meteor magnitude and echo duration is explained by this mechanism as are the departures from the wavelength-squared variation of echo duration predicted by diffusion theory. Attachment processes also account for the observation that the final heights of enduring meteor echoes all center about 95 km even though bright meteors may show a maximum in light intensity below 80 km. (See 2337 of 1960.)

523.746.5 2340

The Character of the Next Sunspot Maximum—W. Gleissberg. (*Z. Astrophys.*, vol. 49, pp. 25–29; January 13, 1960.) From an application of probability laws previously used (see 92 of 1950) it is concluded that the next sunspot maximum may be weak. During the next 11-year cycle the smoothed monthly averages of the relative sunspot numbers are not expected to exceed 87.7, with a probability of 0.95.

523.746.5 2341

An Estimate of the Peak Sunspot Number in 1968—C. M. Minnis. (*Nature*, vol. 186, p. 462; May 7, 1960.) There is a probability of 0.75 that the next smoothed peak sunspot number will be in the range 110–160.

523.75 2342

On the Origin of Solar Flares—T. Gold and F. Hoyle. (*Monthly Notices Roy. Astr. Soc.*, vol. 120, pp. 89–105; 1960.) The high concentration of energy stored in the chromosphere can be accounted for by a particular class of magnetic field whose lines of forces have the general shape of twisted loops protruding above the photosphere. When loops of opposite sense and twist meet, energy is dissipated in the form of a solar flare.

523.75 2343

Effects of the Solar Flares of 7 July 1958—B. Hultqvist, J. Aarons and J. Ortner. (*Tellus*, vol. 11, pp. 319–331; August, 1959.) An analysis of observations made at Kiruna, Sweden, of ionospheric and geomagnetic disturbances following flares which occurred at 0032 and 0040 U.T.

550.38 2344

Lines of Force of the Geomagnetic Field in Space—E. H. Vestine and W. L. Sibley. (*Planet. Space Sci.*, vol. 1, pp. 285–290; September, 1959.) The lines of force from points in the northern hemisphere are traced to the southern hemisphere using the first nine Gauss coefficients, with particular reference to the auroral zones.

550.385.4 2345

The Ring Current and the Outer Atmosphere—S. Akasofu. (*J. Geophys. Res.*, vol. 65, pp. 535–543; February, 1960.) By hydromagnetic wave propagation through the ionosphere at speeds about 1000 km [see 863 of 1960 (Francis *et al.*)], changes in the geomagnetic field at ground level would be delayed less than 1 minute after ring-current changes at several earth radii. The electric current associated with particle motions in the Van Allen belts is examined. [See also 1535 of 1959 (Hines and Storey).]

550.385.4 2346

Correlated Micropulsations at Magnetic Sudden Commencements—W. K. Berthold, A. K. Harris, and H. J. Hope. (*J. Geophys. Res.*, vol. 65, pp. 613–618; February, 1960.) Using wire loops 52 square miles in area, micropulsations were recorded simultaneously in Arizona and New Jersey during the initial phase of a magnetic storm. Cross-correlation between the two records shows a time difference of two to three seconds which corresponds to that for the sudden commencement.

550.385.4 2347

Diurnal and Annual Variations of the Normalized Amplitudes of Sudden Commencements of Magnetic Storms at Tamanrasset—J. L. Bureau. (*C. R. Acad. Sci.* (*Paris*), vol. 249, pp. 1543–1545; October 19, 1959.) (An extension of earlier work: see 1183 of 1959.)

550.385.4(98) 2348

Three-Dimensional Consideration for Current-System of Geomagnetic Variations: Part 1—Current Flow for Polar Elementary Storm Within a Spherical Conducting Shell—N. Fukushima. (*Rep. Ionosphere Res. Japan*, vol. 12, pp. 437–447; 1958.)

551.507.362.2 2349

Effects of Solar Radiation Pressure on Earth-Satellite Orbits—R. W. Parkinson, H. M. Jones, and I. I. Shapiro. (*Science*, vol. 131, pp. 920–921; March 25, 1960.) "Calculations show that, at a mean altitude of 1000 miles, radiation pressure can displace the orbit of the 100-foot Echo balloon at rates up to 3.7 miles per day, the orbit of the 12-foot Beacon satellite at 0.7 mile per day. For certain resonant conditions this effect accumulates, drastically affecting the satellite's lifetime."

551.507.362.2 2350

Perturbations in Perigee Height of Vanguard I—P. Musen, R. Bryant, and A. Bailie. (*Science*, vol. 131, pp. 935–936; March 25, 1960.) The effect of solar radiation pressure on the perigee height of satellite 1958 $\beta 2$ is considered. The inclusion of this effect leads to closer agreement between observed orbital data and theoretical results.

551.507.362.2 2351

A Doppler-Cancellation Technique for Determining the Altitude Dependence of Gravitational Red Shift in an Earth Satellite—R. S. Badessa, R. L. Kent, J. C. Nowell, and C. L. Searle. (PROC. IRE, vol. 48, pp. 758–764; April, 1960.) Description of a technique by which the frequency of a moving source is corrected for first-order Doppler shifts.

551.507.362.2 2352

Profile of Upper-Atmosphere Air Density at the Height 180–212 km derived from the Orbit of Sputnik III—Z. V. Bochníček. (*Nature*, vol. 186, pp. 460–461; May 7, 1960.)

551.507.362.2 2353
Satellite Orbits and Atmospheric Densities at Altitudes up to 750 km Obtained from the Vanguard Orbit Determination Program—J. W. Siry. (*Planet. Space Sci.*, vol. 1, pp. 184–192; August, 1959.) The program is outlined. Results of Vanguard I measurements show that the Minitrack and computing systems can be operated with an over-all accuracy within 1 milliradian. Atmospheric density results are discussed.

551.507.362.2 2354
On an Aspect of the Relations Between Mechanics and Artificial Earth Satellites—M. Roy. (*C. R. Acad. Sci. (Paris)*, vol. 249, pp. 1424–1425; October 19, 1959.) It is suggested that satellites which orbit equatorially may offer an opportunity to study fundamental laws of mechanics.

551.507.362.2:061.3 2355
A Report on the Symposium on the Use of Space Vehicles at the Fall 1958 U.R.S.I. Meeting—W. C. Hoffman. (*Planet. Space Sci.*, vol. 1, pp. 238–248; August, 1959.) A discussion of possible experiments in the following classes: 1) VLF satellite measurements, 2) ion probes, 3) Faraday rotation measurements, 4) topside ionospheric sounders, 5) studies of anomalous propagation.

551.507.362.2:523.165 2356
Ionizing Radiation Detected by Pioneer II—A. Rosen, P. J. Coleman, Jr. and C. P. Sonett (*Planet. Space Sci.*, vol. 1, pp. 343–346; September, 1959.) The total ionizing component of cosmic radiation was measured to 1550 km, and shows a steady decrease from 24°N to 30°N near this height.

551.507.362.2:523.165 2357
Altitude Dependence and Time Variation of the Radiation Intensity observed by U. S. Satellite 1958α—Y. Miyazaki and H. Takeuchi. (*Rep. Ionosphere Res. Japan*, vol. 12, pp. 448–458; 1958.) [See also 448 of 1959 (Aono and Kawakami.)]

551.507.362.2:523.75 2358
Solar Activity and Deceleration of Earth Satellites—W. Priester. (*Naturwiss.*, vol. 46, pp. 197–198; March, 1959.) A correlation between the irregularities of the orbital period of satellite 1957β and solar radio emission at 20 cm λ during the period December, 1957–January, 1958 is discussed.

551.507.362.2:537.56 2359
The Satellite Ionization Phenomenon—J. D. Kraus, R. C. Higgy and W. R. Crone. (PROC. IRE, vol. 48, pp. 672–678; April, 1960.) Evidence is given of ionization effects due to the passage of satellites; they are most noticeable a day or so after large solar flares.

551.507.362.2:621.3–71 2360
Radiative Cooling of Satellite-Borne Electronic Components—J. R. Jenness. (PROC. IRE, vol. 48, pp. 641–643; April, 1960.) The basic principles of radiative cooling in a satellite are discussed. It is considered feasible that temperatures of 250°K or less could be maintained provided the proper skin coating is used.

551.507.362.2:621.391.812.63 2361
Ionospheric Scintillations of Satellite Signals—H. P. Hutchinson and P. R. Arendt. (PROC. IRE, vol. 48, pp. 670–671; April, 1960.) Results showing variations from a smooth Doppler-shift curve for various frequencies and orbits give a measure of frequency scintillation occurring and hence the roughness of the intervening ionospheric path.

551.51 2362
Thermal and Gravitational Atmospheric Oscillations—Ionospheric Dynamo Effects Included—M. L. White. (*J. Atmos. Terr. Phys.*, vol. 17, pp. 220–245; February, 1960.) The resonance theory of gravitational and thermal oscillations in a rotating atmosphere is extended to include an electron and positive-ion gas with a permanent magnetic field superposed. Expressions are obtained for the variation with height of the electric field, current density, and ion-drift velocities. Some comparisons are made with observed ionospheric data.

551.510.5:551.507.362 2363
Atmospheric Densities from Satellites and Rocket Observation—K. W. Champion and R. A. Minzner. (*Planet. Space Sci.*, vol. 1, pp. 259–264; September, 1959.) Current data suggest revision of the 1956 A.R.D.C. model atmosphere to lower densities near 100-km altitude and to higher densities above 160 km. A revised model is presented.

551.510.535 2364
Artificial Electron Clouds: Part 1—F. F. Marmo, L. M. Aschenbrand and J. Pressman. (*Planet. Space Sci.*, vol. 1, pp. 227–237; August, 1959.) A summary report of the methods, techniques and operating procedures used in a series of eight rocket experiments in which artificial electron clouds were produced in the upper atmosphere at heights between 70 and 130 km.

551.510.535 2365
Artificial Electron Clouds: Part 2—F. F. Marmo, J. Pressman and L. M. Aschenbrand. (*Planet. Space Sci.*, vol. 1, pp. 291–305; September, 1959.) Theoretical and practical considerations in the creation of artificial clouds are presented. The altitude limits for the generation of long-lived dense clouds are between 70 and 200 km. (See pt. 1: 2364 of 1960.)

551.510.535 2366
Artificial Electron Clouds: Part 3—F. F. Marmo, L. M. Aschenbrand and J. Pressman. (*Planet. Space Sci.*, vol. 1, pp. 306–318; September, 1959.) A cloud was formed at 121 km by the release of atomic potassium from an Aerobee rocket. An analysis of results gave approximate values for wind velocity, diffusion constant, thermal ionization efficiency, and initial half-width of the cloud. (See pt. 2: 2365 of 1960).

551.510.535 2367
The Mesopause Region of the Ionosphere—J. D. Whitehead. (*Nature*, vol. 186, p. 461; May 7, 1960.) A proposition by Gregory (2728 of 1958) that the layer at a height of 80–90 km is formed by photo-ionization of dust particles and is considered to be untenable.

551.510.535 2368
Strongly Absorbing Layers below 50 km—B. Hultqvist and J. Ortner. (*Planet. Space Sci.*, vol. 1, pp. 193–204; August, 1959.) Measurements at Kiruna, Sweden, during the summer of 1958 showed the height of the ionized layer causing long-duration absorption to be about 50 km. This supports the hypothesis that the ionization is produced by protons. (See 2374 of 1960.)

551.510.535 2369
On Some Characteristics of the E_s Layer—M. Bossolasco and A. Elena. (*Planet. Space Sci.*, vol. 1, pp. 205–212; August, 1959.) Some aspects of the dependence of E_s ionization on the geomagnetic field are discussed. An attempt is made to deduce the mean world-wide drift direction of E_s in summer by using the time displacement that occurs for the daily maxima of the hourly median values of f_oE_s.

551.510.535 2370
An Attempt to Measure the Co lision Frequency of Electrons in the F Region of the Ionosphere—D. M. Schlapp. (*J. Atmos. Terr. Phys.*, vol. 17, pp. 246–253; February, 1960.) Measurements have been made of the change in the F_2-layer reflection coefficient with change in group path. Results indicate that values of collision frequency obtained from this type of measurement are unreliable, but an upper limit of about 5×10^3 sec^{-1} is deduced.

551.510.535 2371
Abnormal Features of the F_2 Region of the Ionosphere at some Southern High Latitude Stations—R. G. Rastogi. (*J. Geophys. Res.*, vol. 65, pp. 585–592; February, 1960.) The diurnal variation of f_0F_2 at Port Lockroy shows a maximum near midnight and a minimum near noon, during the summer months. This abnormal variation is characteristic of the southern west zone only and is not observed at other high-latitude stations in either the south or the north.

551.510.535 2372
Generalization of the Appleton-Hartree Magneto-ionic Formula—H. K. Sen and A. A. Wyller. (*Phys. Rev. Lett.*, vol. 4, pp. 355–357; April 1, 1960.) Recent laboratory results [151 of 1960 (Phelps and Pack)] show that it is not valid to neglect the effect of the electron velocity distribution function. An applicable generalized expression for the complex refractive index is obtained.

551.510.535 2373
The Ion Distribution above the F_2 Maximum—F. S. Johnson. (*J. Geophys. Res.*, vol. 65, pp. 577–584; February, 1960.) Recombination is important up to 550 km but above this height the ion distribution is controlled by diffusion. Protons produced near 550 km by charge exchange between hydrogen atoms and oxygen ions move upward along the magnetic-field lines with an equal number of electrons and produce a medium for the propagation of whistlers.

551.510.535:523.75 2374
On the Interpretation of Ionization in the Lower Ionosphere occurring on both Day and Night Side of the Earth within a Few Hours after some Solar Flares—B. Hultqvist. (*Tellus*, vol. 11, pp. 332–343; August, 1959.) The absorption effect following solar flares may by interpreted as being caused by a high-energy ion beam of very small density emitted from the sun, the ions moving in Störmer orbits.

551.510.535:551.55 2375
Large-Scale Irregularities in High-Altitude Winds—J. S. Greenhow and E. L. Neufeld. (*Proc. Phys. Soc.*, vol. 75, pp. 228–234; February 1, 1960.) A discussion of irregularities at heights of 80 to 100 km, having a depth of about 10 km and a horizontal extent of over 100 km. The lifetime is found to be about 1.5. (See also 4058 of 1959.)

551.510.535(98) 2376
Enhanced Ionization in the Polar Ionosphere caused by Solar Corpuscular Emissions—T. Obayashi and Y. Hakura. (*J. Radio Res. Labs., Japan*, vol. 7, pp. 27–66; January, 1960.) Enhanced ionization, which follows several hours after a large solar flare, is considered to be caused by high-energy protons of 10–100 mev which follow essentially Störmer orbits.

551.510.535(98):621.391.812.631 2377
Polar Blackouts associated with Severe Geomagnetic Storms on September 13, 1957,

and February 11, 1958—Y. Hakura, Y. Takenoshita and T. Otsuki. (*Rep. Ionosphere Res. Japan*, vol. 12, pp. 459–468; 1958.)

551.594.5 2378

On the Shape of the Auroral Zones—L. Block. (*Ark. Fys.*, vol. 14, pp. 153–160; October 14, 1958.) Model experiments on Alfvén's theory of auroras performed earlier (2621 of 1955) have been extended to study auroral-zone eccentricity and include measurements of space charges and the influence of the terella potential. Results show that the theory is applicable to the model experiments.

551.594.5 2379

Geometry of the Southern Auroral Zone and the Evidence for the Existence of an Inner Zone—T. Hatherton. (*Nature*, vol. 186, pp. 288–299; April 23, 1960.)

551.594.5: 550.385 2380

Studies of the Upper Atmosphere from Invercargill, New Zealand: Part 1—Characteristics of Auroral Radar Echoes at 55 Mc/s—R. S. Unwin. (*Ann. géophys.*, vol. 15, pp. 377–394; July/September, 1959. In English.) Full report of work briefly described earlier (2593 of 1959).

551.594.5: 550.385 2381

Studies of the Upper Atmosphere from Invercargill, New Zealand: Parts 2 & 3—M. Gadsden. (*Ann. géophys.*, vol. 15, pp. 395–402, 403–411; July/September, 1959. In English.) Analysis of data shows 1) that the probability of occurrence of radar echoes increases with an increase in local magnetic K-index; 2) a radar auroral zone whose size is related to the K-index; 3) that echoes are observed in the daytime to less extent than suggested by the diurnal variation of local K-index; 4) that there is no direct relation between radar echo regions observed at 55 mc and visual auroras. (See pt. 1: 2380 of 1960.)

551.594.6 2382

A Four-Year Summary of Whistler Activity at Washington, D. C.—H. E. Dinger. (*J. Geophys. Res.*, vol. 65, pp. 571–575; February, 1960.) Results are analyzed and presented in the form of curves of diurnal and month-to-month variations of whistlers, hiss and dawn chorus.

551.594.6 2383

The Relation between Lightning Discharges and Whistlers—H. Norinder and E. Knudsen. (*Planet. Space Sci.*, vol. 1, pp. 173–183; August, 1959.) Observations made near Uppsala show that whistler-producing atmospherics contain strong components around 5 kc and often show multiple discharges. Atmospherics not producing whistlers often show irregular constitution and a single discharge, but those giving rise to high field strengths are always followed by whistlers.

551.594.6: 550.385.4 2384

Low-Frequency Electromagnetic Radiation Associated with Magnetic Disturbances—G. R. A. Ellis. (*Planet. Space Sci.*, vol. 1, pp. 253–258; September, 1959.) Continuous observation of naturally occurring radiation in the band 2–40 kc at Sydney, Australia, show major bursts to be associated with strong auroras and magnetic activity.

LOCATION AND AIDS TO NAVIGATION

534.88 2385

Determining Sonar System Capability—G. Rand. (*Electronics*, vol. 33, pp. 41–45; February 19, 1960.) Charts are given relating to pulse length, noise generated by surface disturbances, temperature changes etc., from which a figure of merit can be obtained to compare different systems under various operating conditions.

621.396.9: 523.164 2386

A Study of Natural Electromagnetic Phenomena for Space Navigation—R. G. Franklin and D. L. Birx. (Proc. IRE, vol. 48, pp. 532–541; April, 1960.) It is doubtful whether radiation from natural celestial bodies, except for the sun, is strong enough for a practical navigation system.

621.396.9: 535–15 2387

Background Noise Measurements at the Sea Horizon—L. J. Free. (*J. Opt. Soc. Amer.*, vol. 49, pp. 1007–1011; October, 1959.) Description of an infrared background-noise analyzer and discussion of measurements made in 1956 and 1957 of the horizontal wave-number distribution of noise from the sky and the sea at the horizon.

621.396.9: 551.507.362.2 2388

Navigation using Signals from High-Altitude Satellites—A. B. Moody. (Proc. IRE, vol. 48, pp. 500–506; April, 1960.) A system using a small number of satellites at an optimum height between 1000 and 12,000 miles in conjunction with a radio sextant, is considered feasible.

621.396.9: 551.507.362.2 2389

A Satellite Doppler Navigation System—W. H. Guier and G. C. Weiffenbach. (Proc. IRE, vol. 48, pp. 507–516; April, 1960.) A positional accuracy of 0.5 nautical mile is theoretically possible by making full use of the Doppler curve obtained from a single satellite transit.

621.396.9: 551.507.362.2 2390

Measurement of the Doppler Shift of Radio Transmissions from Satellites—G. C. Weiffenbach. (Proc. IRE, vol. 48, pp. 750–754; April, 1960.) The technique, intended for use as a navigational aid, has yielded promising results when used experimentally at 108 mc, but requires further development.

621.396.933 2391

Flight Safety—O. Heer. (*VDI Z.*, vol. 102, pp. 498–502; April 21, 1960.) Annual review covering changes in flight regulations, with particular reference to the Frankfurt (Main) area, and mentioning recent developments in the field of radio and radar navigational aids. 24 references.

621.396.96: 621.3.087.4 2392

Storage Processes for the Improvement of Signal/Noise Ratio of Almost Periodic Signals, Particularly in Radar Applications—H. Meinke and K. Rihaczek. (*Nachrichtentech. Z.*, vol. 12, pp. 176–180; April, 1959.) In the storage of measurements of random processes a minimum NSR is obtained depending on storage time and fluctuation of the measured signal with time. Formulas and curves are given for evaluating the optimum storage period for a given bandwidth to ensure minimum errors of measurement. Reference is made to the tracking of weather balloons by radar.

621.396.963.3: 551.507.362.2 2393

Tracking and Display of Earth Satellites—F. F. Slack and A. A. Sandberg. (Proc. IRE, vol. 48, pp. 655–663; April, 1960.) Predicted satellite paths can be displayed on an orthographic or Mercator projection overlay on a cathode-ray tube.

621.396.969.3: 551.507.362.2 2394

The Navy Space Surveillance System—R. L. Easton and J. J. Fleming. (Proc. IRE, vol. 48, pp. 663–669; April, 1960.) Describes a system of satellite tracking using a CW transmitter coplanar with two interferometer receivers.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215: 546.47'221: 539.23 2395

Noise and Signal Response in Lead Sulphide Photoconductive Films—H. E. Spencer. (*J. Appl. Phys.*, vol. 31, pp. 505–510; March, 1960.) Results of experiments over the temperature range 25°C to −173°C agreed well with theory.

535.215: 546.48'221 2396

Field Effect in CdS Single Crystals—W. Thielemann. (*Z. Naturforsch.*, vol. 14a, pp. 92–93; January, 1959.) Preliminary report on observations of momentary photocurrent changes resulting from jerky displacements of a shadow on the crystal surface.

535.215: 546.48'221 2397

Luminescence and Conductivity of CdS Crystals with Excitation by Medium-Energy Electrons—G. Eichhoff, G. O. Müller and G. Schubert. (*Naturwiss.*, vol. 46, p. 201; March, 1959.) The luminescence spectrum was obtained and the cathodoconductivity measured on nonactivated crystals excited by electron bombardment at 0.6–6 kev and by X-rays.

535.215: 546.48'221 2398

Optical Absorption of CdS Single Crystals at the Fundamental Lattice Absorption Edge—K. W. Böer and H. Gutjahr. (*Z. Phys.*, vol. 155, pp. 328–331; June 10, 1959.) Measurements show that optical absorption is governed by indirect band-to-band transitions involving phonons.

535.215: 546.48'221: 534.286–8 2399

Photosensitive Ultrasonic Attenuation in CdS—H. D. Nine. (*Phys. Rev. Lett.*, vol. 4, pp. 359–361; April 1, 1960.) The effect of light intensity on the ultrasonic attenuation proved to be linear but of opposite direction in two crystals examined. Two types of mechanism are being considered as possible explanations.

535.37: [546.47'221 + 546.48'221 2400

The Incorporation and Effects of Oxygen in ZnS and CdS Phosphors—N. Riehl and R. Sizmann. (*Z. Naturforsch.*, vol. 14a, pp. 394–403; April, 1959.) Experimental investigations show the direct effect of oxygen in the sulphide lattice on luminescence properties of the phosphors.

535.37: 546.681'17 2401

The Edge Emission and other Emissions of GaN—H. G. Grimmeiss and H. Koelmans. (*Z. Naturforsch.*, vol. 14a, pp. 264–271; March, 1959.)

535.37.092 2402

Effect of Pressure on Phosphor Decay—D. W. Gregg and H. G. Drickamer. (*J. Appl. Phys.*, vol. 31, pp. 494–496; March, 1960.) An apparatus has been developed to measure the effect of pressure to over 50,000 atm on the decay rate of phosphors. Results for certain phosphors are given, with a tentative explanation of the effects.

535.377: 546.47'221 2403

Thermoluminescence of ZnS Single Crystals—H. Arbell and A. Halperin. (*Phys. Rev.*, vol. 117, pp. 45–52; January 1, 1960.)

537.226 + 537.228.1]: 546.714–31 2404

The Dielectric and Piezoelectric Behaviour of Pyrolusite (Polycrystalline Ore of MnO_2)—

J. N. Das. (*Z. Phys.*, vol. 155, pp. 465–471; July 16, 1959. In English.) (See also 1266 of 1959.)

537.226:621.319.2 2405

Field Measurements on Electrets and Investigations of the Origin of Homocharge—J. van Calker and L. van der Linde. (*Z. Phys.*, vol. 155, pp. 413–421; July 16, 1959.) The results of field mapping in the immediate vicinity of carnauba-wax electrets are in agreement with the field distribution calculated by Swann (608 of 1951). The development of the homocharge is shown to be due to charge-carrier injection.

537.227 2406

Dielectric Properties of a Single Crystal of Partially Deuterated Glycine Sulphate—J. Chapelle and L. Taurel. (*C. R. Acad. Sci.*, (*Paris*), vol. 249, pp. 1332–1333; Ocrober 12, 1959.)

537.227 2407

Effect of Hydrostatic Pressure on the Ferroelectric Properties of Triglycine Sulphate and Selenate—F. Jona and G. Shirane. (*Phys. Rev.*, vol. 117, pp. 139–142; January 1, 1960.) The variation of transition temperature with pressure up to 2700 atm is linear.

537.227 2408

Wall Velocity in Ferroelectrics—J. C. Burfoot. (*Proc. Phys. Soc.*, vol. 75, pp. 312–314; February 1, 1960.) A discussion of existing evidence regarding the reversal of polarization in ferroelectric crystals under an applied field, by the advance of the reversed phase behind 180° walls.

537.227:546.431'824–31 2409

Symmetry of the Low-Temperature Phase of Barium Titanate—E. Sawaguchi and M. L. Charters. (*Phys. Rev.*, vol. 117, pp. 465–469; January 15, 1960.) Domain patterns and optical properties were examined using thin (111) plates. Crystal symmetry below −80°C was shown to be rhombohedral.

537.227:621.318.57 2410

The Mechanism of the Reversal of the Spontaneous Polarization in $LiH_3(SeO_3)_2$ Single Crystals—E. Fatuzzo. (*Helv. Phys. Acta*, vol. 33, pp. 21–26; March 15, 1960.) The model of a switching mechanism proposed by Fatuzzo and Merz (1267 of 1960) is applied to the ferroelectric material $LiH_3(SeO_3)_2$. The domain-nucleus interaction is found to be very high.

537.228.1.082 2411

Determination of Piezoelectric Properties as a Function of Pressure and Temperature—J. E. McKinney and C. S. Bowyer. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 56–61; January, 1960.) Data on a mixed titanate system composed of 82 per cent $BaTiO_3$, 9.1 per cent $CaTiO_3$, 3.6 per cent $PbTiO_3$ and 4.4 per cent TiO_2 have been obtained with an apparatus designed to measure the dynamic compressibility of materials.

537.311.31:546.621 2412

Measurements of the High-Temperature Electrical Resistance of Aluminum: Resistivity of Lattice Vacancies—R. O. Simmons and R. W. Balluffi. (*Phys. Rev.*, vol. 117, pp. 62–68; January 1, 1960.)

537.311.33 2413

The Question of the Existence of Tetrahedral Phases—O. G. Folberth. (*Z. Naturforsch.*, vol. 14a, pp. 94–96; January, 1959.) The possibility of obtaining over 100 compounds of tetrahedral structure by the process of "cross-substitution" [see 847 of 1959 (Goodman)] is discussed; reasons for some exceptions to the rules governing the possibility of existence of such compounds are given.

537.311.33 2414

The Current/Voltage Characteristic of an n-p Junction Considering Generation and Recombination of Current Carriers in the Space-Charge Layer—A. D. Chevychelov. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1205–1212; 1959.) Rigorous calculation of I/V characteristics for forward and reverse current directions with reference to the theory developed by Sah, *et al.* (3899 of 1957). For values of $d/L \approx 1$, where d is the half-width of the transition layer and L the diffusion length, the carrier densities are not appreciably decreased by recombination in the transition layer.

537.311.33 2415

Impurity Band Theory—V. L. Bonch-Bruevich. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1213–1220; 1959.) Theoretical treatment of impurity states in a doped homopolar semiconductor such as Ge doped with elements of Group III or V. Taking account of the continuous spectrum, which is equivalent to allowing for virtual transitions to the conduction band of the base material, leads to impurity band formation at concentrations as low as 10^{14}–10^{15} cm^{-3}.

537.311.33 2416

The Theory of Electron Plasma in Semiconductors—V. L. Bonch-Bruevich and Sh. M. Kogan. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1221–1224; 1959.) Note describing an approximation method for the calculation of the boson Green's function from which the plasma frequencies can be determined.

537.311.33 2417

Determination of the Capture Cross-Section in the Case of Recombination on Multiply Charged Centres—Yu. A. Kontsevoĭ. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1289–1293; 1959.) Description of a method based on measurements of the dependence of the lifetime of non-equilibrium carriers on their concentration.

537.311.33 2418

Entrainment of Ions by Electrons in Semiconductors—V. B. Fiks. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1321–1323; 1959.) An estimate is made of the concentration of electrons at which the entrainment effect due to electron scattering becomes appreciable. Investigation of this effect based on the effective mobility of ions and neutral atoms is suggested as a method of studying the mechanism of scattering of electrons by impurity centers, particularly at high temperatures.

537.311.33:535.215 2419

Determination of Free-Carrier Lifetimes in Semiconductors from the Relaxation Time of Photo-Excited Infrared Absorption—L. Huldt. (*Ark. Fys.*, vol. 15, pp. 229–236; May 8, 1959.)

537.311.33:535.215:538.63 2420

Elimination of Edge Effect in the Measurement of Photomagnetic E.M.F. in Semiconductors—B. Ya. Moĭzhes. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1239–1242; 1959.) Formulas of one-dimensional theory are applicable for any width of the illuminated region provided this is a sufficient distance from the contacts and the ends of the specimen.

537.311.33:537.323.08 2421

Several New Methods to Measure the Thermal Diffusivity of Semiconductors—J. H. Becker. (*J. Appl. Phys.*, vol. 31, pp. 612–613; March, 1960.)

537.311.33:538.615 2422

Line Shapes of I.M.O. Absorption in the Semiconductors—T. Ohta and T. Miyakawa. (*Prog. Theoret. Phys.*, vol. 22, pp. 893–895; December, 1959.) A note on the development of the theory of interband magneto-optical absorption, with reference to strong-field and weak-field approximations. [See 2285 of 1959 (Burnstein *et al.*).]

537.311.33:546.23 2423

Electric Properties of Selenium with Gold Impurity—V. G. Sidyakin. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1172–1175; 1959.)

537.311.33:546.273'171 2424

Crystal Potential and Energy Bands of Semiconductors: Part 2—Self-Consistent Calculations for Cubic Boron Nitride—L. Kleinman and J. C. Phillips. (*Phys. Rev.*, vol. 117, pp. 460–464; January 15, 1960.) Results are given and compared critically with previous calculations (2052 of 1960).

537.311.33:546.27.18 2425

Semiconducting Properties of Cubic Boron Phosphide—B. Stone and D. Hill. (*Phys. Rev. Lett.*, vol. 4, pp. 282–284; March 15, 1960.)

537.311.33:[546.28+546.289 2426

On the Possibility of Detection of Excitons in Germanium and Silicon—B. Ya. Moĭzhes and Yu. I. Ravich. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1243–1245; 1959.) Estimates of probable exciton densities indicate that excitons should be detectable in the photoconduction of Ge at temperatures of the order of 1.5°K.

537.311.33:546.28 2427

Electron Capture by a Lattice Vacancy in Si—A. Morita. (*Phys. Rev.*, vol. 117, pp. 84–89; January 1, 1960.) Calculation of electron-capture cross-section in Si.

537.311.33:546.28 2428

Experimental Proof of the Existence of a New Electronic Complex in Silicon—J. R. Haynes. (*Phys. Rev. Lett.*, vol. 4, pp. 361–363; April 1, 1960.) Recombination radiation from doped Si crystals shows extremely sharp lines at low temperatures. The radiation is produced by recombination of an electron and a hole both of which are bound in an immobile four-particle complex consisting of an impurity ion and three electronic particles.

537.311.33:546.28 2429

The Effect of Annealing on the Carrier Lifetime of Silicon—G. Ziegler and M. Zerbst. (*Z. Naturforsch.* vol. 14a, pp. 93–94; January, 1959.) The quenching effect on carrier lifetime was investigated in high-purity Si, and results obtained agree in general with those of Bemski (161 of 1957). An explanation for the observed recovery in lifetime, after its pronounced reduction by quenching, is given on the basis of tests on sand-blasted specimens.

537.311.33:546.28:538.569.4 2430

Electron Spin-Lattice Relaxation in Phosphorus-Doped Silicon—A. Honig and E. Stupp. (*Phys. Rev.*, vol. 117, pp. 69–83; January 1, 1960.) The various spin-lattice relaxation mechanisms of paramagnetic electrons associated with P impurities in Si are identified; their individual properties are determined and their origins discussed.

537.311.33:546.28:538.569.4 2431

Spin Resonance of Transition Metals in Silicon—H. H. Woodbury and G. W. Ludwig. (*Phys. Rev.*, vol. 117, pp. 102–108; January 1, 1960.)

537.311.33:546.28–31:539.23 2432

Electron-Microscope Investigations of Electrolytically Formed Silicon Oxide Films—A. Polityeki and E. Fuchs. (*Z. Naturforsch.*, vol. 14a, pp. 271–275; March, 1959.) The porous nature of anodically formed oxide films is investigated. Films formed on polished Si during exposure to air appear to be nonporous. [See also 3530 of 1957 (Schmidt and Michel).]

537.311.33:546.289 2433

Theory of the Lattice Vibrations of Germanium—W. Cochran. (*Proc. Roy. Soc.* (*London*), vol. 253, pp. 260–276; November 24, 1959.) An extension of the Born-von Kármán theory of lattice dynamics to apply to a simple model of the Ge crystal in which each atom is regarded as a charged core coupled to an oppositely charged cell.

537.311.33:546.289 2434

The Anisotropy of the Conductivity of Hot Electrons and their Temperature in Germanium—E. G. S. Paige. (*Proc. Phys. Soc.* (*London*), vol. 75, pp. 174–184; February 1, 1960.) The analysis makes use of the data given by Koenig on n-type Ge (3345 of 1959) to calculate the effective temperature of electrons in different valleys. The results are compared with those obtained from an energy-balance equation, and the comparison indicates strong interaction between electrons in different valleys.

537.311.33:546.289 2435

Grain Boundaries in Germanium—R. S. Wagner and B. Chalmers. (*J. Appl. Phys.*, vol. 31, pp. 581–587; March, 1960.) Investigation of the relative energies and the boundary orientations of simple tilt boundaries.

537.311.33:546.289 2436

The Effect of Bismuth on the Density of Dislocations in Germanium Single Crystals—V. G. Alekseeva and P. G. Eliseev. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1304–1307; 1959.)

537.311.33:546.289 2437

The Temperature Dependence of the Coefficient of Electron Capture at the Central Level of Copper in Germanium—S. G. Kalashnikov and A. I. Morozov. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1294–1296; 1959.) Note of measurements of electron lifetime in Ga- and B-doped Ge with diffused Cu.

537.311.33:546.289 2438

Calculation of the Influence of Anisotropy and of a Superimposed Magnetic Field on Electrical Breakdown, with Application to Germanium—W. Frie. (*Z. Naturforsch.*, vol. 14a, pp. 54–62; January, 1959.) Equations are derived for the increase in breakdown field strength by the magnetic field. The theoretical conclusions are compared with results obtained by Finke. (See 2439 of 1960.)

537.311.33:546.289 2439

The Behaviour of n-Type Ge Single Crystals in the Temperature Region of Liquid Helium—G. Finke and G. Lautz. (*Z. Naturforsch.*, vol. 14a, pp. 62–74; January, 1959.) The influence of p-n compensation in the transition from normal to impurity-band conduction in semiconductors shown theoretically [*e.g.*, 155 of 1957 (Conwell)] is confirmed experimentally by measurements on Ge specimens with differing impurity concentrations in the temperature range 4.2°–300°K. The effect on breakdown strength of an external magnetic field is also discussed and used in the interpretation of the phenomena.

537.311.33:546.289 2440

Conductivity of Grown Germanium Bicrystals—H. F. Mataré, B. Reed and O. A. Weinreich. (*Z. Naturforsch.*, vol. 14a, pp. 281–284; March, 1959. In English). Discussion of measurements of grain-boundary conduction. [See also 2657 of 1959 (Reed, *et al.*).]

537.311.33:546.289:537.226.2 2441

On the Dielectric Constant of Germanium at Microwave Frequencies—A. C. Baynham, A. F. Gibson and J. W. Granville. (*Proc. Phys. Soc.*, vol. 75, pp. 306–309; February 1, 1960.) The results show a dispersion region with a resonance frequency of 32.2 kmc. The resulting absorption, in addition to the free-carrier absorption, is about 3 db/cm at 34.75 kmc.

537.311.33:546.289:548.73 2442

X-Ray Detection of Dislocations in Germanium—V. Gerold and F. Meier. (*Z. Phys.*, vol. 155, pp. 387–394; July 16, 1959.) The method of Barth and Hosemann (*Z. Naturforsch.*, vol. 13a, pp. 792–794; September, 1958) is used; the majority of dislocations lie in the (111) planes.

537.311.33:546.47–31 2443

Investigations of the Electrical Conductivity of Zinc Oxide—J. Derén, J. Haber and T. Wilkowa. (*Z. Phys.*, vol. 155, pp. 453–464; July 16, 1959.) Polycrystalline ZnO was investigated in the temperature range 100°–700° C at normal atmospheric pressure and in vacuum. The differences in the electrical properties of ZnO sintered above 1000°C and material which is unsintered, or sintered below 1000°C, are discussed and an interpretation is given of the irreversible conductivity changes observed between 20° and 450°C.

537.311.33:546.623'681'86 2444

Investigation of Conductivity and Hall Effect in Solid Solutions of the AlSb-GaSb System—I. I. Burdiyan and B. T. Kolomiets. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1165–1171; 1959.)

537.311.33:546.681'19 2445

Growth of Gallium Arsenide by Horizontal Zone Melting—J. L. Richards. (*J. Appl. Phys.*, vol. 31, pp. 600–603; March, 1960.)

537.311.33:546.681'19 2446

Etch Pits in Gallium Arsenide—J. L. Richards and A. J. Crocker. (*J. Appl. Phys.*, vol. 31, pp. 611–612; March, 1960.)

537.311.33:546.681'19 2447

Magnetoresistance in Gallium Arsenide—R. K. Willardson and J. J. Duga. (*Proc. Phys. Soc.* (*London*), vol. 75, pp. 280–290; February 1, 1960.) A study of the transverse magnetoresistance in n-type GaAs as a function of impurity concentration and density of defects introduced by fast-neutron irradiation. The experimental results agree with theoretical predictions.

537.311.33:546.681'86 2448

Experimental Investigation of Conduction Band of GaSb—A. Sagar. (*Phys. Rev.*, vol. 117, pp. 93–100; January 1, 1960.) Measurements have been made of 1) Hall effect and conductivity between 1.5° K and 370° K, 2) change of resistance and Hall effect with hydrostatic pressure, and 3) resistance change due to uniaxial stress between 77° K and 370° K. An explanation of the data is given on the basis of a double conduction band.

537.311.33:546.682'18 2449

Piezoresistance in n-Type InP—A. Sagar. (*Phys. Rev.*, vol. 117, p. 101; January 1, 1960.) Results of piezoresistance measurements indicate a spherical energy band.

537.311.33:546.682'19 2450

Mass-Spectrometer Investigation of Phenomena During Vaporization of Indium Arsenide—H. B. Gutbier. (*Z. Naturforsch.*, vol. 14a, pp. 32–36; January, 1959.)

537.311.33:546.682'19'18 2451

Optical Investigations on the Semiconducting Mixed-Crystal Series $In(As_yP_{1-y})$—F. Oswald. (*Z. Naturforsch.*, vol. 14a, pp. 374–379; April, 1959.) Infrared transmissivity and reflection measurements on n-type specimens of differing conductivity were made at 1–35 $\mu\lambda$. [see also 1683 of 1955 (Oswald and Schade).] The width of the forbidden band as a function of As content and temperature is determined from an optical determination of the energy gap (814 of 1957). The results of other authors are discussed with reference to these measurements.

537.311.33:546.817'241 2452

Dislocation Etch Pits on p-Type Lead Telluride—B. B. Houston and M. K. Norr. (*J. Appl. Phys.*, vol. 31, pp. 615–616; March, 1960.)

537.311.33:[546.873'231+546.47'86 2453

On the Diffusion of Certain Impurities in Bi_2Se_3 and ZnSb—A. A. Kuliev. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1176–1178; 1959.)

537.533.8 2454

An Investigation of Electron Reflection from some Metals—I. M. Bronshtein and R. B. Segal. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1246–1249; 1959.) Experimental investigation of the secondary-emission coefficient and the coefficient of inelastically scattered primary electrons for angles of incidence up to 60°.

537.533.8:537.226 2455

An Investigation of the Secondary Electron Emission of certain Dielectrics at Low Primary Electron Energies—S. A. Fridrikhov and A. R. Shul'man. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1259–1267; 1959.) The first critical potential of various materials including glass, mica, and stibnite have been measured. The coefficient of elastic reflection at very low electron energies is quite high (0.6–0.7).

538.22:548.5 2456

The Growth of Oxide Single Crystals Containing Transition Metal Ions—F. W. Harrison. (*Research* (*London*), vol. 12, pp. 395–403; October/November, 1959.) Different methods are reviewed, including the flame fusion process, the Bridgman-Stockbarger method and lead oxide solution methods. 38 references.

538.221 2457

Anisotropy Induced by a Magnetic Field in Thin Vapour-Deposited Permalloy Films—W. Andrä, Z. Málek, W. Schüppell and O. Stemme. (*Naturwiss.*, vol. 46, pp. 257–258; April, 1959.) Preliminary note on magnetic anisotropy measurements by the torque method on permalloy films of 700–1600Å thickness, in a field of 6000 oersteds.

538.221 2458

The Passage through Zero of the Crystal Energy in Magnetite Containing Manganese—A. V. Kienlin. (*Naturwiss.*, vol. 46, p. 258; April, 1959.) The variation of permeability with temperature was measured on specimens of Fe_3O_4 with substituted Mn in the range −195° to +50°C. The permeability maximum is displaced towards lower temperatures with increasing Mn content.

538.221 2459

Investigations of the Temperature Dependence of the Procopiu Effect—F. A. Koch. (*Z. Phys.*, vol. 155, pp. 475–478; July 16, 1959.) The effect described by Hofbauer and Koch (1272 of 1952) is the occurrence of a longitudinal magnetic component when a ferromagnetic wire is circularly magnetized. Oscillograms and curves are given resulting from tests made on Fe wire in the temperature range 100°–800°C.

538.221:539.23 2460

The Problem of the Magnetization of Very Thin Iron Films—H. Hoffmann and C. Schwink. (*Naturwiss.*, vol. 46, pp. 198–199; March, 1959.) Preliminary report on magnetization measurements on Fe films of thickness 60–80Å and 150Å, which show that magnetic properties of these films change even under high-vacuum conditions.

538.221:621.318.134 2461

Cooperative Magnetism in Oxide Structures—L. C. F. Blackman. (*Research* (*London*), vol. 12, pp. 164–171, 218–225; May/June, 1959.) Antiferromagnetic, ferromagnetic and ferrimagnetic properties of various oxide systems are discussed in terms of indirect magnetic exchange interations. 43 references.

538.221:621.318.134 2462

Some Aspects of the Stability of Permeability of Ferrites—R. Smith. (*Proc. IRE* (*Australia*), vol. 20, pp. 733–736; December, 1959.) Measurements on a sample of Mn-Zn ferrite which has been magnetically saturated and subsequently demagnetized show that the permeability increases initially and then decays to a value 3–4 per cent below the maximum value. Effects of heating on permeability are also discussed.

538.221:621.318.134 2463

A Comparison of Static and Microwave Measurements of Magnetocrystalline Anisotropy in Cobalt Manganese Ferrite—R. F. Pearson and R. W. Teale. (*Proc. Phys. Soc.* (*London*), vol. 75, pp. 314–316; February 1, 1960.)

538.221:621.318.134:621.318.57 2464

Reversal Time of Ferrites as a Storage System—A. Jaecklin. (*Tech. Mitt. PTT*, vol. 37, pp. 140–144; April 1, 1959.) A method is described in which the time taken for magnetic reversal of the square-loop core is measured and used to represent a stored digit.

538.221:621.318.134:621.372.85 2465

Ferrite Bodies with Temperature-Independent Gyromagnetic Properties—W. Haken and C. von Haza-Radlitz. (*Arch. elekt. Übertragung*, vol. 13, pp. 157–160; April, 1959.) For certain dimensions of magnetically biased ferrite bodies the temperature dependence of the resonance frequency, and of the nonreciprocal phase shift or Faraday rotation, becomes negligible. These dimensions are calculated for ferrite strips in rectangular waveguide, and are confirmed by measurements.

538.221:621.395.625.3 2466

Experimental and Theoretical Investigation of the Magnetic Properties of Iron Oxide Recording Tape—E. D. Daniel and I. Levine. (*J. Acoust. Soc. Amer.*, vol. 32, pp. 1–15; January, 1960.) Remanence tests show that the major anhysteretic properties of a tape may be expressed in terms of three easily measured constants. The results are interpreted theoretically in conjunction with the Preisach diagram.

538.652:539.37 2467

The Influence of Magnetostriction on the Energy and Thickness of Bloch Walls—G. Rieder. (*Z. Naturforsch.*, vol. 14a, pp. 96–98; January, 1959.) Calculation of internal stresses and plastic deformation effects as a function of Bloch-wall parameters. (See also 2167 of 1958, and for a general treatment of mechanical work in plastic processes, *Z. angew. Phys.*, vol. 10, pp. 140–150; March, 1959.)

621.315.6:538.569.3 2468

Electromagnetic Properties of Insulators: Part 1—V. Ambegaokar and W. Kohn. (*Phys. Rev.*, vol. 117, pp. 423–431; January 15, 1960.) The response of a perfect insulator to weak external EM fields of long wavelengty is discussed on a many-particle basis. It is completely described by a single frequency-dependent dielectric constant.

MATHEMATICS

512.9 2469

Some Tests for the Number of Positive Zeros and for the Numbers of Real and Complex Zeros of a Real Polynomial—O. P. D. Cutteridge. (*Proc. IEE*, vol. 107, pp. 105–110; March, 1960.)

MEASUREMENTS AND TEST GEAR

621.3.011.32(083.74) 2470

Correction for Size of Cross-Section of the Secondary Windings of Mutual-Inductance Standards of the Campbell Type—P. Vigoureux. (*Brit. J. Appl. Phys.*, vol. 10, pp. 481–483; November, 1959.) The mutual inductance can be calculated with negligible error by replacing each turn of the secondary winding by a circle at the center of the cross section of the wire.

621.3.018.41(083.74):621.317.373 2471

Continuous Phase Comparison of Frequency Standards—G. Zito. (*Alta Frequenza*, vol. 28, pp. 100–118; April, 1959.) A phase comparator is described for operation at 100 kc. It has a resolution of 1° and can accumulate the total phase shift over any number of beat cycles.

621.317.33.029.3 2472

The Use of Differential Transformers for Resistance Measurements at High Frequencies—B. Lavagnino and B. Alby. (*Alta Frequenza*, vol. 28, pp. 119–132; April, 1959.) A method of measuring the resistance of conductors at audio and ultrasonic frequencies is described. The wires under test are arranged as two parallel lines of 10-m length 25 cm apart and short-circuited at the ends; this forms the single-turn secondary of a special toroidal transformer. Results of measurements at frequencies up to 200 kc are given for various wire materials.

621.317.361:551.507.362 2473

Applications of Doppler Measurements to Problems in Relativity, Space-Probe Tracking and Geodesy—R. R. Newton. (PROC. IRE, vol. 48, pp. 754–758; April, 1960.) Various systems are proposed for improving the accuracy of Doppler measurements.

621.317.411.029.62:621.376.432 2474

Measurement of Complex Permeability by Discriminator—J. C. Anderson. (*J. Brit. IRE*, vol. 20, pp. 219–223; March, 1960.) The theory and design of a circuit for measuring the real and imaginary parts of VHF permeability in a ferromagnetic sample are given.

621.317.44 2475

Field Measurements by the Method of Harmonics: Theoretical Conscierations—J. Greiner. (*Nachrichtentech.*, vol. 9, pp. 173–180; April, 1959.) The operating principles of three fundamental types of magnetometer head are described. The theory underlying the magnetometer principle adopted, *e.g.*, Palmer (1142 of 1954) is particularly considered; practical considerations will be covered in two subsequent papers.

621.317.61.029.6 2476

Microwave Gain Measurement—L. G. Sebestyen. (*Electronic Technologist*, vol. 37, pp. 195–196; May, 1960.) When a gas-discharge diode is used as a signal source for measuring the gain of a traveling-wave tube, the measured gain may be different from the true gain. The reason for this is explained and a digaram is given from which the required correction factor may be obtained.

621.317.7:621.374.32 2477

An Events-per-Unit-Time Meter (E.P.U.T. Meter)—J. D. Storer. (*Electronic Engrg.*, vol. 32, pp. 160–162; March, 1960.) The circuit is given of an inexpensive instrument counting up to 9999 impulses per second with a read-out period of 0.75 second.

621.317.7:621.391.821 2478

Atmospheric Noise Structure—Clarke. (See 2504 of 1960.)

621.317.72:537.311.31 2479

Measurement of Contact-Potential Changes for Gas Adsorption on Metal Surfaces with Constant Capacitance of the Measuring Capacitor—A. Eberhagen, R. Jaeckel and F. Strier. (*Z. angew. Phys.*, vol. 11, pp. 131–134; April, 1959.)

621.317.723:551.507.362 2480

High-Speed Electrometers for Rocket and Satellite Experiments—J. Praglin and W. A. Nichols. (PROC. IRE, vol. 48, pp. 771–779; April, 1960.) Instruments for measuring currents down to 5×10^{-14} Å with a response up to about 30 cps are described.

621.317.742.029.64 2481

Simple Standing-Wave Measurements at Low Input Powers—A. Stainforth and J. H. Craven. (*J. Brit. IRE*, vol. 20, pp. 243–245; March, 1960.) The voltage SWR of microwave components containing a crystal detector can be measured by the method described.

621.317.755 2482

Sampling Oscilloscope for Millimicrosecond Pulses at a 30-mc Repetition Rate—A. S. Farber. (*Rev. Sci. Instr.*, vol. 31, pp. 15–17; January, 1960.) The samples synchronous signals at a repetition rate of 31.25 mc, or multiples. The sweep frequency is 1 kc. When used with the waveguide coincidence circuit described for examining a signal of carrier frequency 10 kmc, the maximum sensitivity is approximately 1 mv of output per μw of signal, with a noise level of 2 μw.

621.317.761 2483

Construction of a Frequency Meter for High Frequencies—H. Hahn. (*C.R. Acad. Sci.* (*Paris*), vol. 249, pp. 1199–1201; October 5, 1959.) Brief description of a frequency meter and associated discriminator system suitable for the absolute measurement of small frequency fluctuations in the range 29.625–29.635 mc.

621.317.772.029.6 2484

Precision Phasemeter for C.W. or Pulsed U.H.F.—R. T. Stevens. (*Electronics*, vol. 33, pp. 54–57; March 4, 1960.) An instrument is described for measuring the phase difference between either CW or pulsed RF signals in the frequency range 100–520 mc to within 0.2 and 0.5 degrees respectively. The range may be extended up to X-band frequencies or down to 20 mc.

621.317.789.029.63:621.316.8 2485

Designing and Construction of 100-Watt Water Load for Power Standard in U.H.F. Band—T. Takahashi and N. Nakahashi. (*J. Radio Res. Labs., Japan*, vol. 7, pp. 9–18; January, 1960.) A water-cooled coaxial resistor based on a design proposed by Kohn (2019 of 1954) is described. The voltage SWR is less than 1.1 for frequencies up to 700 mc

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

535.376:681.6 2486

An Electroluminescent Digital Indicator with a Silicon Carbide Coding Matrix—D. H. Mash. (*J. Sci. Instr.*, vol. 37, pp. 47–50; February, 1960.) The coding matrix is made by a printed-circuit technique and operates in conjunction with a single-pole ten-way switch to select the required elements on an electroluminescent panel.

621.3.029.64:621.179.15 2487

The Use of Linearly Polarized Microwaves for the Nondestructive Testing of Materials—A. Dietzel, E. Deeg, and E. M. Amrhein. (*Elektron. Rundschau*, vol. 13, pp. 122–123; April, 1959.) Equipment is described for the recording of the anisotropy distribution in ceramic objects.

621.316.72.078:621.382.2/.3:629.11 2488

Solid-State Generator Regulator for Autos—L. D. Clements. (*Electronics*, vol. 33, pp. 52–54; February 19, 1960.) Two transistors and a Zener diode regulate the voltage; reverse-current and overload protection are provided by a diode and transistor.

†621.383.292:536.581.1 2489

Temperature-Regulated Ignition Source—J. Preston and L. E. Ashman. (*Rev. Sci. Instr.*, vol. 31, pp. 53–58; January, 1960.) A photomultiplier monitors radiation from a wire heated to between 700°C and melting point; changes in radiation result in compensating changes in current. The temperature can be kept constant to within a few degrees.

621.385.833 2490

Reflexion Electron Microscopy Using Diffracted Electrons—J. S. Halliday and R. C. Newman. (*Brit. J. Appl. Phys.*, vol. 11, pp. 158–162; April, 1960.) Better resolution is obtained than with normal reflection techniques.

621.385.833 2491

The Visualization of Dislocations for Reproduction by Electron Microscopy—H. Bethge and V. Schmidt. (*Z. Naturforsch.*, vol. 14a, pp. 307–309; March, 1959.) A method based on crystal growth along dislocations is described.

621.385.833:539.23 2492

Application of Electron Microscopy—A. Saulnier and P. Mirand. (*Metal Treat.*, vol. 27, pp. 91–100, 102; March, 1960.) The preparation of thin metal and alloy films suitable for direct examination under the electron microscope in metallurgical investigations is described.

621.385.833:539.23 2493

Interference Microscopy Using Electron Waves—R. Buhl. (*Z. Phys.*, vol. 155, pp. 395–412; July 16, 1959.) The design of a biprism electron-interferometer microscope is described; micrograms are reproduced and discussed.

621.396.9:526.2:061.3 2494

International Symposium on Electronic Distance-Measuring Techniques—(*J. Geophys. Res.*, vol. 65, pp. 385–528; February, 1960.) The text is given of some of the papers presented at a symposium held in Washington, D. C., May 5–12, 1959; others are summarized.

621.398:681.142 2495

An Analogue and Digital Airborne Data Acquisition System—D. H. Ellis and J. M. Walter, Jr. (PROC. IRE, vol. 48, pp. 713–724; April, 1960.) An account is given of the principles, construction and operation of a flexible, fully transistorized system for recording data on 14 or 23 tracks of a 1-inch magnetic tape.

PROPAGATION OF WAVES

621.391.812.5:523.5 2496

The Azimuth Distribution of Oblique Reflections from Meteor Trails and its Relation to Meteor Radiant Distributions—Bain. (See 2338 of 1960.)

621.391.812.6 2497

A Possibility of the Long-Distance H.F. Propagation Along the Exospheric Field-Aligned Ionizations—T. Obayashi. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 603–612; July, 1959.) Propagation along exospheric field-aligned ionization is suggested as the explanation of HF back-scatter echoes at ranges of 7000 km. Ray-path calculations give echo distances which agree well with observed results. The average exospheric electron density deduced from echo spreads is about 10^4–10^6 electrons/cm^3.

621.391.812.62 2498

Vertical Distribution of the Scattering Parameter for Radio Wave—K. Hirao, K. Akita and I. Shiro. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 595–601; July, 1959.) Observations in Japan are described which were carried out in the first 1000 m above ground level using a captive, balloon-borne refractive-index variometer. The results suggest that the scattering parameter in this region is inversely proportional to the 2/3 power of the height and varies with season.

621.391.812.63.029.45 2499

On the Propagation of E.L.F. Radio Waves and the Influence of a Nonhomogeneous Ionosphere—J. R. Wait. (*J. Geophys. Res.*, vol. 65, pp. 597–600; February, 1960.) A model of a spherical ionosphere whose electron density increases exponentially with height is used to explain observed attenuation as a function of frequency in the range 100 cps–1 kc.

621.391.812.63.029.62 2500

Night-Time Equatorial Propagation at 50 Mc/s: First Results from an I.G.Y. Amateur Observing Program—M. P. Southworth. (*J. Geophys. Res.*, vol. 65, pp. 601–607; February, 1960.) Transequatorial propagation is an evening phenomenon occurring most frequently at 2100–2200 L.M.T. for the mid-point of the circuit and being most pronounced around the equinoxes. It is unlikely to be due to layer tilts but may be due either to the high electron densities at the last sunspot maximum or to equatorial spread-F.

621.391.812.631:523.75 2501

Short-Wave Fade-Outs without Reported Flares—H. DeMastus and M. Wood. (*J. Geophys. Res.*, vol. 65, pp. 609–611; February, 1960.) Of 15 such events during the I.G.Y. period only in one case was there no *Hα* activity. Over the same period the number of fade-outs which did occur with flares was about 280.

621.391.812.8 2502

On a Method of Short-Term Prediction of f_0F_2—H. Shibata and S. Watanabe. (*J. Radio Res. Labs. Japan*, vol. 7, pp. 19–25; January, 1960.) Analysis of a proposed time series indicates that a short-term prediction of f_0F_2 could be made, provided that seasons could be chosen so that the series is stationary during each season.

621.391.812.8 2503

World Maps of F_2 Critical Frequencies and Maximum-Usable-Frequency Factors for Use in Making Ionospheric Radio Predictions—D. H. Zacharisen and V. Agy. (*J. Geophys. Res.*, vol. 65, pp. 593–595; February, 1960.) Note on a National Bureau of Standards publication giving values of f_0F_2 and the M4000 factor at a sunspot number of 50, and the rates of change of these parameters with sunspot activity.

RECEPTION

621.391.821:621.317.7 2504

Atmospheric Noise Structure—C. Clarke. (*Electronic Technologist*, vol. 37, pp. 197–204; May, 1960.) A detailed description of equipment for omnidirectional measurements in the range 15 kc–20 mc of the amplitude probability distribution and other characteristics of atmospheric radio noise.

STATIONS AND COMMUNICATION SYSTEMS

621.391 2505

The Limit Behaviour of Extremely Selective Communication Channels in Information Theory—H. Wolter. (*Arch. elekt. Übertragung*, vol. 13, pp. 171–174; April, 1959.) A Gaussian error function cannot be a real band function irrespective of the phase response, and there is no Küpfmüller limit for these functions. (See also 2161 of 1960.)

621.391 2506

The Logical Connection as a Unit in Information Processing—K. Steinbuch. (*Nachrichtentech. Z.*, vol. 12, pp. 169–175; April, 1959.) Various terms based on the concept of "logical connection" are defined. These terms facilitate the evaluation of the efficiency of information-handling systems.

621.391 2507

Theoretical Considerations for the Element Error Rate in Binary Code Transmission—H. Akima. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 543–572; July, 1959.) Relations between element error rate and SNR are calculated for fading-free signals in the presence of random noise. Amplitude and frequency-shift systems of modulation are considered and the results are tabulated in numerical form.

621.391:621.376.3 2508

Intermittent Communication of Frequency-Modulated Binary Code—K. Ikushima. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 573–593; July, 1959.) Systems of intermittent transmission controlled by a signal from the receiver terminal are considered. Theoretical evaluation indicates a power gain of 20–30 db over conventional continous-transmission systems.

621.391:621.395.665.1 2509

On the Recovery of a Band-Limited Signal, After Instantaneous Companding and Subsequent Band Limiting—H. J. Landau. (*Bell Syst. Tech. J.*, vol. 39, pp. 351–364; March, 1960.) The stability and convergence properties of a proposed iterative formula for signal recovery are investigated and a description is given of an analog simulator giving a first approximation to a practical realization of the recovery scheme.

621.391:629.19 2510

Communication Efficiency Comparison of Several Communication Systems—R. W. Sanders. (PROC. IRE, vol. 48, pp. 575–588; April, 1960.) Orthogonal matched filter systems require less energy per bit than others.

621.396.41:621.391.827 2511

Radio-Frequency Interference Considerations in the TD-2 Radio Relay System—H. E. Curtis. (*Bell Syst. Tech. J.*, vol. 39, pp. 369–387; March, 1960.) The relation between the ratio of a desired RF carrier to an interfering co-channel RF carrier and the resulting telephone-channel interference is developed, and interference in a hypothetical long system is expressed in terms of an equivalent permissible noise level. The extent to which theoretical objectives can be achieved in practice is discussed.

621.396.43:629.19 2512
Space Communications Requirements of the Department of Defence—J. O. Spriggs. (Proc. IRE, vol. 48, pp. 600–602; April, 1960.) Mainly a discussion of the use of space vehicles as passive or active relays.

621.396.43:629.19 2513
Interference and Channel Allocation Problems Associated with Orbiting-Satellite Communication Relays—F. E. Bond, C. R. Cahn and H. F. Meyer. (Proc. IRE, vol. 48, pp. 608–612; April, 1960.) Calculations of the extent of mutual interference show that careful coordination will be necessary. The advantages of passive systems are noted.

621.396.43:629.19 2514
Passive Satellite Communication—J. L. Ryerson. (Proc. IRE, vol. 48, pp. 613–619; April, 1960.) Using power per channel per unit range as a criterion, it is considered that such systems are likely to be one order of magnitude better than those using tropospheric scatter.

621.396.43:629.19 2515
The Use of a Passive Spherical Satellite for Communication and Propagation Experiments—T. H. Vea, J. B. Day and R. T. Smith. (Proc. IRE, vol. 48, pp. 620–624; April, 1960.) The theoretical scattering properties of an aluminum-coated sphere are considered.

621.396.43:629.19 2516
Project SCORE—S. P. Brown and G. F. Senn. (Proc. IRE, vol. 48, pp. 624–630; April, 1960.) Details of the development, construction and operation of "Signal Communication by Orbiting Relay Experiment."

621.396.65:621.398 2517
Automatic Switching and Remote Signalling in Radio Link Systems—A. Vighi. (*Note Recensioni Notiz.*, vol. 8, pp. 179–185; March /April, 1959.) Problems of switching-on standby equipment and signalling in case of failure are discussed with reference to CCIR recommendations. Details are given of the automatic equipment used in the radio link Rome–Pescara.

621.396.946 2518
A Pragmatic Approach to Space Communication—G. E. Mueller. (Proc. IRE, vol. 48, pp. 557–566; April, 1960.) Communication with large information rates is possible throughout the solar system, but practical interstellar communication is not yet feasible.

621.396.946 2519
Propagation and Communications Problems in Space—J. H. Vogelman. (Proc. IRE, vol. 48 pp. 567–569; April 1960.) Discussion of the problems of Doppler shift, Faraday rotation, tracking and stabilization of antennas.

621.396.946 2520
Propagation Doppler Effects in Space Communications—F. J. Tischer. (Proc. IRE, vol. 48, pp. 570–574; April, 1960.) A generalized derivation of the Doppler equation taking inhomogeneities into account.

621.396.946 2521
Extraterrestrial Noise as a Factor in Space Communications—A. G. Smith (Proc. IRE, vol. 48, pp. 593–599; April, 1960.) An examination of existing data, with bibliography.

621.396.946 2522
Extraterrestrial Radio Tracking and Communication—M. H. Brockman, H. R. Buchanan, R. L. Choate and L. R. Malling. (Proc. IRE, vol. 48, pp. 643–654; April, 1960.) Comprehensive details are given of the probe transmitter and ground receiver of the TRAC(E) system for tracking and telemetry at lunar distances. A phase locking technique is used with receiver bandwidth 20 cps at an operating frequency of 960.05 mc. A range of over 4×10^5 miles was achieved with Pioneer IV, and ranges up to 4×10^9 miles are predicted.

621.396.946:621.398 2523
Maximum Utilization of Narrow-Band Data Links for Interplanetary Communications—W. F. Sampson. (Proc. IRE, vol. 48, pp. 589–593; April, 1960.) A general discussion defining fields requiring investigation.

621.396.946:621.398 2524
Signal-to-Noise Considerations for a Space Telemetry System—R. W. Rochelle. (Proc. IRE, vol. 48, pp. 691–693; April, 1960.)

621.398:621.396.934 2525
Telemetry Bandwidth Compression using Airborne Spectrum Analyzers—A. G. Ratz. (Proc. IRE, vol. 48, pp. 694–702; April, 1960.)

621.398+621.396]:629.19 2526
The Telemetry and Communication Problem of Re-entrant Space Vehicles—E. F. Dirsa. (Proc. IRE, vol. 48, pp. 703–713; April, 1960.) Methods for overcoming the thermal ionization (plasma) problem are discussed.

SUBSIDIARY APPARATUS

621.311.69:537.311.33:535.215 2527
Effect of Temperature on Photovoltaic Solar Energy Conversion—J. J. Wysocki and P. Rappaport. (*J. Appl. Phys.*, vol. 31, pp. 571–578; March, 1950.) Theoretical investigation of the behavior of materials with band gaps varying from 0.7 to 2.4 ev, over a temperature range of 0–400°C, show that the best conversion performance is obtained for the ideal junction current; it is degraded by the presence of recombination current. Experimental data are given for Si, GaAs and CdS cells, and are compared with theoretical expectations.

621.314.634 2528
Investigation of Selenium Rectifiers Under Pulse Conditions—D. N. Nasledov and I. M. Yashukova. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1188–1192; 1959.) To investigate the voltage dependence of the resistance in the reverse direction, pulse measurements have been made of back resistance, capacitance, barrier-layer thickness and electric-field intensity, as functions of the applied voltage.

621.316.721.078.3:538.569.4 2529
Stabilization of the Magnetic Field of an Electromagnet—R. Becherer and R. Reimann. (*C.R. Acad. Sci.* (*Paris*), vol. 249, pp. 1340–1342; October 12, 1959.) A system is described for controlling both rapid and slow variations of the magnetic field by means of a transistorized control stage and feedback winding. Slow variations are detected by the magnetic resonance of protons in a ferric nitrate solution using a technique similar to that described earlier [1402 of 1960 (Reimann)]. The stability over a 1-h period is within 1 part in 5×10^6.

621.316.79:536.581 2530
Precision Thermostat—Y. Hiruta, T. Kawana and T. Shirai. (*J. Radio Res. Labs., Japan*, vol. 6, pp. 533–542; July, 1959.) A temperature-sensitive resistance bridge controls a crystal oven to give a short-period temperature stability within 1×10^{-4}°C.

621.316.79:621.382.08 2531
A Stable Thermostat for the Measurement of the Noise and Drift of Semiconductor Circuit Elements—E. Baldinger and A. Maier. (*Z. angew. Math. Phys.*, vol. 11, pp. 68–73; January 25, 1960.) The rectified output of a bridge circuit containing a 100-Ω Pt resistance thermometer is used to control the heating of a stirred oil bath. The noise voltage of transistors in a power-supply circuit and of a Zener reference diode have been measured at very low and audio frequencies using the apparatus described.

621.316.925.451:621.382 2532
An Experimental Impedance Relay using the Hall Effect in a Semiconductor—H. E. M. Barlow and J. C. Beal. (*Proc. IEE*, vol. 107, pp. 48–50; February, 1960.) Description of a "definite" impedance relay for the protection of power transmission systems. The Hall output voltage from an InSb crystal is balanced by a voltage from a rectifier unit; when a fault occurs in the system an out-of-balance voltage is produced which operates an electromagnetic relay.

TELEVISION AND PHOTOTELEGRAPHY

621.397:621.391.83 2533
Determination of a Noise Evaluation Filter for Television—J. Müller and E. Demus. (*Nachrichtentech. Z.*, vol. 12, pp. 181–186; April, 1959.) A noise-weighting curve for the 625-line system is derived from subjective tests and is compared with various international proposals embodied in CCIR and CCITT documents. A practical filter network is given whose attenuation characteristic closely conforms to the weighting curve.

621.397.132:535.88 2534
The Eidophor Colour Projection System—(*J. Telev. Soc.*, vol. 9, pp. 127–131; October /December, 1959.) A description is given of the monochrome projector. Color pictures are projected using in addition two synchronously rotating filter disks mounted in front of the camera and projector, producing a line-sequential system. [See also 620 of 1959 (Gretener).]

621.397.132:621.395.625.3 2535
The State of Development and Possibilities of Application of the Ampex Method of Magnetic Recording of Television Signals—H. J. v. Braunmühl. (*Rundfunktech. Mitt.*, vol. 3, pp. 61–65; April, 1959.) Performance and operating costs of the Ampex tape-recording method, modified for the 625-line CCIR system, and 16-mm film recording are compared.

621.397.132:621.396.62 2536
An Experimental Colour TV Receiver: Setting-Up and Adjustment—E. Ribchester. (*J. Telev. Soc.*, vol. 9, pp. 137–145; October /December, 1959.) A discussion of the experience gained in developing and using several experimental receivers operating on the NTSC system.

621.397.331.22 2537
The Problem of the Reduction in Resolution of Image Orthicons due to Crosstalk caused by the Scanning Fields in the Image Converter Section—H. Fix and W. Habermann. (*Rundfunktech. Mitt.*, vol. 3, pp. 76–80; April, 1959.) Tests were made to assess the magnitude of the crosstalk effect. An improvement in resolution is achieved by applying to the tube a field opposing that causing the interference.

621.397.331.222:535–15 2538
Doped Silicon and Germanium Photoconductors as Targets for Infrared Television Camera Tubes—R. W. Redington and P. J. van Heerden. (*J. Opt. Soc. Amer.*, vol. 49, pp.

997–1001; October, 1959.) The properties of Au-doped Si as a photoconductive target in a vidicon-type infrared camera tube are superior to those of commercial vidicon targets in the visible range in every respect except resolution. Cu-, Au-, Ag- and Te-doped Ge all showed a nonimaging state which developed with use.

621.397.331.24:778.53 2539

Measurements on a New Tube for Image Recording—W. Dillenburger. (*Elektron. Rundschau*, vol. 13, pp. 115–118; April, 1959.) Details are given of a television picture tube for use in film recording. Measurements were made of image sharpness in terms of modulation depth, and of reduction of line structure and interference bands by an adjustment of spot wobble.

621.397.6:778.53 2540

The Equipment of the B.B.C. Television Film Studios at Ealing—N. F. Chapman. (*B.B.C. Engrg. Div. Mono.*, No. 27, 31 pp; January, 1960).

621.397.61:621–52 2541

Automation of Television Programme Switching—G. E. Partington. (*J. Brit. IRE*, vol. 20, pp. 181–195; March, 1960. Discussion pp. 195–196.) Operations are controlled by punched paper tape prepared with the program schedule.

621.397.61:621.396.67 2542

Tunable Diplexers of Constant Input Impedance with Residual-Sideband Filters for Television Picture and Sound Transmitters—J. Holle. (*Frequenz*, vol. 13, pp. 102–107; April, 1959.) The design of picture/sound dividing networks for bands I, III and IV conforming to CCIR transmission specifications is discussed with reference to performance measurements on existing equipment.

621.397.62:621.376.33 2543

A TV Sound Section Using the Locked-Oscillator Quadrature-Grid Detector—R. A. Darnell. (*Proc. IRE* (*Australia*), vol. 20, pp. 680–687; November, 1959.) Details are given of a detector using a Type-6DT6 tube which is simpler than the ratio detector and more sensitive than the quadrature-grid detector. It operates satisfactorily with an input of 5–10 mv rms to the driver stage.

621.397.712.3 2544

A Contribution to the Planning of Television Studios—G. Stump aad U. Steppputat. (*Rundfunktech. Mitt.*, vol. 3, pp. 91–93; April, 1959.) Description of a regional studio with adjoining control rooms where monitoring and recording equipment is installed.

621.397.712.3 2545

The Re-equipment of the Austrian Television Studios—F. Brunner. (*Rundfunktech. Mitt.*, vol. 3, pp. 66–75; April, 1959.) Details are given of the structural modifications to the studio buildings, and the mixer and control equipment installed is described.

621.397.74:621.372.55 2546

A Transversal Equalizer for Television Circuits—R. V. Sperry and D. Surenian. (*Bell Syst. Tech. J.*, vol. 39, pp. 405–422; March, 1960.) The design and operation of an adjustable residual equalizer that produces harmonically related cosine shapes of gain and delay is explained on the basis of paired-echo theory.

621.397.74.001.4 2547

A Particular Class of Pulses Used For Evaluating the Response of Television Networks—R. Codelupi. (*Note Recensioni Notiz.*, vol. 8, pp. 169–178; March/April, 1959.) Pulse waveforms for use in transient-response analysis are proposed consisting of parabolic arcs fitted together, which approximate to a Gaussian waveform. This avoids the use of convolution integrals or series expansion.

621.397.9:621.039 2548

The Application of Closed-Circuit Television in the Nuclear Industry—P. Barratt and I. M. Waters. (*J. Brit. IRE*, vol. 20, pp. 225–241; March, 1960.) Possible ambient radiation and temperature conditions and their influence on television camera design are discussed.

TUBES AND THERMIONICS

621.38:621.391.822.3 2549

Amplitude Distribution of Shot Noise—E. N. Gilbert and H. O. Pollak. (*Bell Syst. Tech. J.*, vol. 39, pp. 333–350; March, 1960.) Regarding $I(t)$ as the superposition of similarly shaped impulses occurring at random Poisson-distributed times, an integral equation is derived for $Q(I)$, the cumulative amplitude probability distribution. Explicit expressions for $Q(I)$ are given for a number of impulse shapes.

621.382.2/.3 2550

Diffused-Junction Depletion-Layer Calculations—H. Lawrence and R. M. Warner, Jr. (*Bell Syst. Tech. J.*, vol. 39, pp. 389–403; March, 1960.) Results derived from an analytical integration of Poisson's equation for diffusion as a function of reverse voltage, assuming either a Gaussian or complementary error function distribution, indicate the transition between linearly graded behavior at low voltages and step behavior at high voltages. Charts are presented for Si and Ge junctions showing the extent of the transition region.

621.382.2:621.372.632 2551

The Influence of the Time-Dependent Series Resistance of a Capacitance Diode in a Mavar Up-Converter—Bobisch and Sondhauss. (See 2253 of 1960.)

621.382.23 2552

Esaki Diode in InSb—R. L. Batdorf, G. C. Dacey, R. L. Wallace and D. J. Walsh. (*J. Appl. Phys.*, vol. 31, pp. 613–614; March, 1960.) A note on the fabrication of InSb diodes and some of their experimental properties, illustrating the extremely low time constants achieved.

621.382.23:621.375.9 2553

Noise Performance Theory of Esaki (Tunnel) Diode Amplifiers—Hines and Anderson. (See 2278 of 1960.)

621.382.3:621.317.3 2554

Determination of Transistor Parameters $\bar{h}_{12}$ and $\bar{h}_{22}$—R. Dessoulavy and B. Secrétan. (*Z. angew. Math. Phys.*, vol. 11, pp. 75–80; January 25, 1960. In French.) To determine transistor h parameters at frequencies below about 100 cps, measurements are made at 1 kc using a neutralization method. Measurement circuits and results are shown.

621.382.33:621.391.822 2555

Transistor Noise-Factor Calculations—F. Hibberd. (*Electronic Engrg.*, vol. 32, pp. 163–165; March, 1960.) A general expression for the minimum noise factor is derived in terms of the transistor equivalent-circuit parameters.

621.382.333 2556

The Static Current Amplification Factor as a Function of Emitter Current in Transistors with Diffused and Homogeneous Base Regions—P. Kaufmann. (*Arch. elekt. Übertragung*, vol. 13, pp. 141–151; April, 1959.) The current amplification factor is calculated for a diffused-base transistor and results are compared with those obtained for homogeneous-base types. A table is included which gives the product of emitter conductivity and minority-carrier diffusion length for p-n-p and n-p-n Ge and Si transistors with different doping elements.

621.382.333.33 2557

The Threshold Frequency of a Drift Transistor Taking into Account the Variation of the Drawing Field and the Mobility of Carriers in the Base—B. Ya. Moizhes. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1308–1311; 1959.) An expression is derived for the collector current, and the variation of the threshold frequency with impurity concentration is discussed.

621.383.5 2558

Preparation of Photovoltaic Cells from InSb with an Alloyed n-p Junction—V. V. Galavanov and N. A. Erokhina. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1198–1200; 1959.) A note on the construction and performance of cells with Cd on In as the alloying element. Photo-response is linear for low values of illumination.

621.383.5:535.2 2559

Experimental Determination of Spontaneous Photon Fluctuations—G. A. Spesch and M. J. O. Strutt. (*Helv. Phys. Acta*, vol. 33, pp. 69–88; March 15, 1960. In German.) Methods and results of measurements of the efficiency and noise characteristics of a Ge p-n photodiode are discussed. Over a certain range the characteristics fit the requirements of a photon detector (2284 of 1960.)

621.383.5:621.311.69 2560

The Reflection Coefficient of Antireflection Coated Surfaces of Silicon Photocells—V. M. Malovetskaya, V. S. Vavilov and G. N. Galkin. (*Fiz. Tverdogo Tela*, vol. 1, pp. 1201–1204; 1959.) The reflection coefficient of Si surfaces is lowered by a factor of 3–4 in the range of maximum sensitivity of a solar battery by coating it with a durable layer of SiO_2 produced by oxidation. The short-circuit current is 20–25 per cent higher than for photoelements with an etched surface.

621.385.032.213.13 2561

The Conductivity of Oxide Cathodes: Part 7—Solid Semiconduction—G. H. Metson and E. Macartney. (*Proc. IEE*, vol. 107, pp. 91–97; March, 1960.) The observed vulnerability of the solid-conductivity phase to low-pressure oxygen attack, even at 300°K, indicates that this is largely a surface phenomenon. [Part 6: 2428 of 1959 (Metson).]

621.385.1:534.29 2562

Noise Acceleration—a New Test Method for the Assessment of Microphony of Thermionic Valves—H. Hellmann. (*Frequenz*, vol. 13, pp. 83–89; March, 1959.) A review of conventional microphony tests is followed by a description of a white-noise vibration method; its advantages over sinusoidal or pulse test methods are discussed. [See also 2194 of 1960 (Köhler and Uhlenbrok).]

621.385.1.029.6 2563

Microwave Tubes—an Introductory Review with Bibliography—A. F. Harvey. (*Proc. IEE*, vol. 107, pp. 29–59; March, 1960. Discussion.) The review covers grid-controlled tubes, crossed-field tubes and noise. Underlying principles are discussed, and typical construction and performance data are given. 698 references.

621.385.3.029.62/.63 2564

The Cause of Deviations from the Theoretical Value of Shot Noise in V.H.F. Triodes with High Mutual Conductance—R. Thiekert. (*Nachrichtentech. Z.*, vol. 12, pp. 201–204; April, 1959.) The results of experimental investigations on tubes exhibiting deviations from the theoretical noise value show that the noise consists of a frequency-dependent and a frequency-independent component. These components are identified with reference to theory.

621.385.6:537.533:621.391.822 2565

On the Nonconservation of Noise Parameters in Multivelocity Beams—J. Berghammer and S. Bloom. (*J. Appl. Phys.*, vol. 31, pp. 454–458; March, 1960.) Microscopic equations for conservation of charge, momentum, and energy, as derived from the Boltzmann equation, are applied to an electron stream having a velocity spread. The minimum noise figure becomes less with position along the beam, as has been demonstrated elsewhere by a numerical solution of the Boltzmann equations [749 of 1958 (Siegman *et al.*)].

621.385.6:621.375.9: 621.372.44 2566

Extension of Longitudinal-Beam Parametric-Amplifier Theory—H. Sobol. (Proc. IRE, vol. 48, pp. 792–793; April, 1960.) Improved agreement with experimental data is obtained by using a model in which both the upper and lower first sidebands of the pump frequency are coupled to the signal.

621.385.63 2567

Certain Mean Values in the Theory of the Travelling-Wave Amplifier—L. A. MacColl. (*Bell Syst. Tech. J.*, vol. 39, pp. 365–367; March, 1960.) Assuming the instantaneous local values of conductor current, potential, linear charge density of electron stream, and velocity of electrons to be periodic functions of time, expressions are derived for their mean time integrals.

621.385.631:621.391.822 2568

The Minimum Noise Figure of the Backward-Wave Amplifier—A. J. Fox. (*J. Electronics Control*, vol. 7, pp. 270–271; September, 1959.) An assessment is made of some variations in noise performance of backward- and forward-wave amplifiers arising from the different couplings of noise power to the circuits.

621.385.633.1 2569

Remarks on the Theory of Backward-Wave Valves—K. H. Löcherer. (*Nachrichtentech. Z.*, vol. 12, pp. 187–192; April, 1959.) Gundlach's theory of delay-line tubes (642 of 1958) is applied to the O-type carcinotron.

621.385.633.1 2570

Construction and Properties of Backward-Wave Valves—G. Bolz. (*Nachrichtentech. Z.*, vol. 12, pp. 120–127; March, 1959.) The design and characteristics of the O-type carcinotron are described. [See 1904 of 1956 (Palluel and Goldberger).]

621.385.633.1 2571

Frequency Stabilization of a High-Power Carcinotron—J. Hervé, J. Pescia and M. Sauzade. (*C.R. Acad. Sci. (Paris)*, vol. 249, pp. 1486–1488; October 19, 1959.) In the system described the carcinotron and reference frequencies are compared in a hybrid-T junction and the difference frequency is used to derive a voltage which controls the sole potential. Stability is maintained over a frequency range of ±150 mc.

621.385.83 2572

Electron-Beam Devices—W. Taeger. (*Frequenz*, vol. 13, pp. 75–82; March, 1959.) A summary of the principles of operation of electron-optical systems with details of recent applications including the resistron camera tube [2451 of 1955 (Heimann)].

621.385.832.032.269.1 2573

High Transconductance Wide-Band Cathode-Ray Gun—E. Atti. (*J. Brit. IRE*, vol. 20, pp. 171–180; March, 1960.) A screen-grid amplifier gun with a low shunt capacitance is described. The operation is analyzed for various types of gun drive and the construction is illustrated.

621.387:621.316.722 2574

Impedance/Frequency Characteristics of Glow-Discharge Reference Tubes—F. A. Benson and P. M. Chalmers. (*Proc. IEE*, vol. 107, pp. 199–208; March, 1960.) Results of measurements over the frequency range 300 cps–5 mc agree with an extended form of an early theory by van Geel. Possible origins of phenomena affecting the LF impedance characteristics are discussed.

621.387:621.374.32 2575

Miniature Gas-Filled Tubes for High-Speed Counting—K. Apel and P. Berweger. (*Electronics*, vol. 33, pp. 46–47; February 19, 1960.) Counting rates up to 1 mc with 25,000 h expected life are achieved.

621.387:681.142 2576

The Digitron: a Cold-Cathode Character Display Tube—N. McLoughlin, D. Reaney and A. W. Turner. (*Electronic Engrg.*, vol. 32, pp. 140–143; March, 1960.) These tubes can have any desired shape of cathode with a single anode. Side- and end-viewing tubes are briefly compared. Their I/V characteristics and associated switching circuits are shown.

MISCELLANEOUS

621.37/.39:061.6 2577

C.C.I.R. Study Group Reviews—(*Point to Point Telecommun.*, vol. 4, pp. 17–109; October, 1959.) Recommendations of the 1959 Los Angeles conference, formal questions and future study programs are given. [See also 1461 of 1960 (Herbstreit).]

621.38:061.3 2578

Recent Progress in Solid-State Technology—M. M. Perugini and N. A. Lindgren. (*Electronics*, vol. 33, pp. 39–43; March 4, 1960.) A note of recent developments discussed at the 1960 Solid-State Circuits Conference, Philadelphia, Pa., February 10–12, 1960, including applications of the tunnel diode and cryotron.

Translations of Russian Technical Literature

Listed below is information on Russian technical literature in electronics and allied fields which is available in the U. S. in the English language. Further inquiries should be directed to the sources listed. In addition, general information on translation programs in the U. S. may be obtained from the Office of Science Information Service, National Science Foundation, Washington 25, D. C., and from the Office of Technical Services, U. S. Department of Commerce, Washington 25, D. C.

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	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Journal of Abstracts, Electrical Engineering (Reseratvnyy Zhurnal: Electronika)	Monthly	Abstracts of Russian and non-Russian literature		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Journal of Experimental and Theoretical Physics (Zhurnal Eksperimentalnoi i Teoreticheskoi Fiziki)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
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	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Radio Engineering and Electronics (Radiotekhnika i Elektronika)	Monthly	Complete journal	National Science Foundation—MIT	Pergamon Institute 122 E. 55 St., New York 22, N. Y.
	Monthly	Abstracts only		Office of Technical Services U. S. Dept. of Commerce Washington 25, D. C.
Solid State Physics (Fizika Tverdogo Tela)	Monthly	Complete journal	National Science Foundation—AIP	American Institute of Physics 335 E. 45 St., New York 17, N. Y.
Telecommunications (Elekprosviaz')	Monthly	Complete journal	National Science Foundation—MIT	Pergamon Institute 122 E. 55 St., New York 22, N. Y.
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Electronics Express	10/year	A digest: abstracts, summaries, annotations of various journals		International Physical Index, Inc. 1909 Park Ave., New York 35, N. Y.
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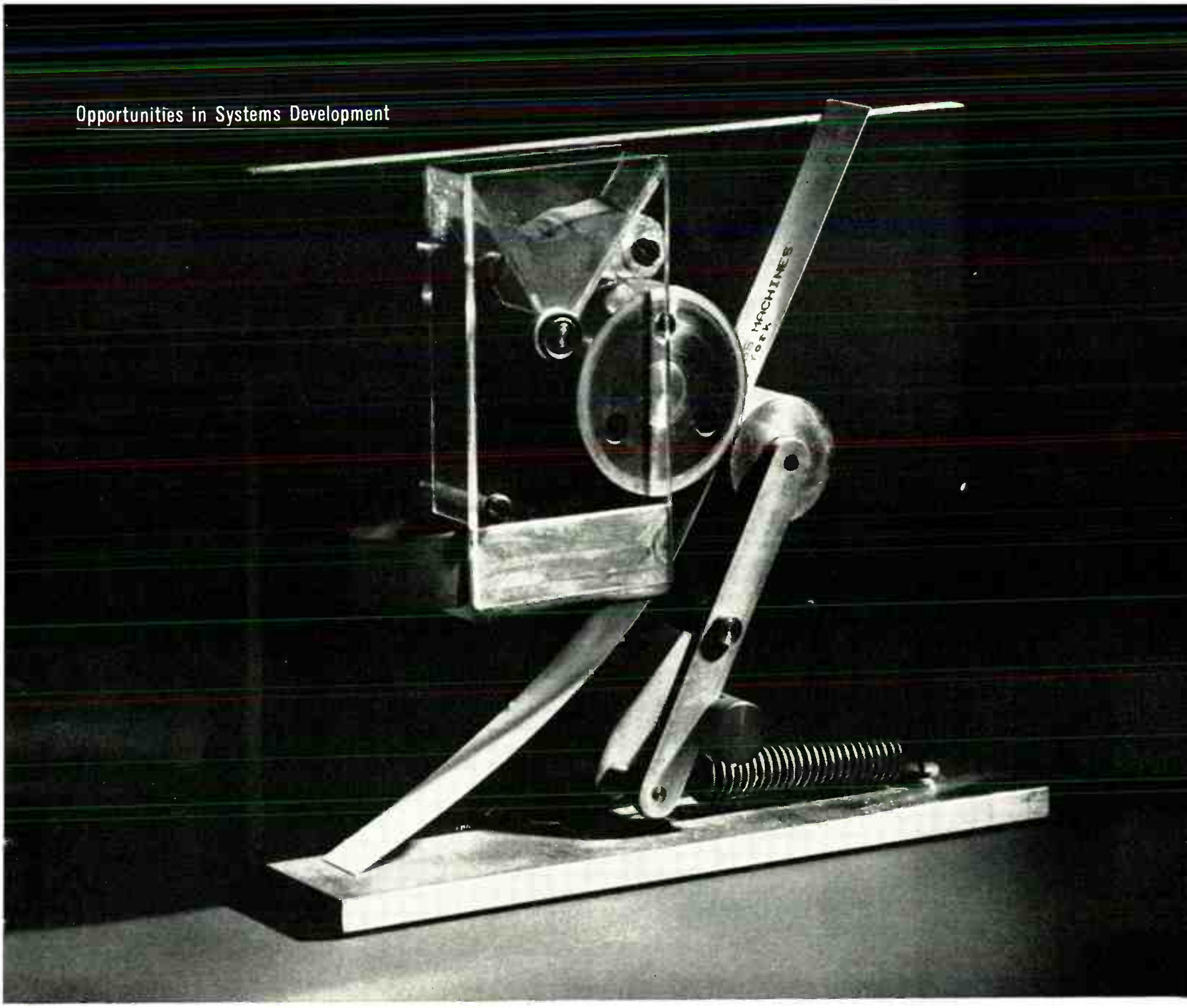
Express Contents of Soviet Journals Currently being Translated into English	Monthly	Advance tables of contents of translated journals		Consultants Bureau, Inc. 227 W. 17 St., New York 22, N. Y.
Technical Translations	Twice a month	Central directory in the U. S. of translations available from all major sources in the U. S.	OTS and Special Libraries Assoc.	Superintendent of Documents U. S. Gov't Printing Office Washington 25, D. C.



Bringing to life the principles of magnetic printing

This little hand-cranked model makes practical use of the principles of magnetic printing—principles that have been known in theory for more than a hundred years.

The model represents a big jump from theory to practice. It was developed and built by a group of IBM engineers and scientists brought together for the sole purpose of proving or disproving the feasibility of the magnetic printing process.

These IBM researchers discovered a way to "write" characters on a rotating nickel-cobalt plated drum by casting a magnetic dot image on the drum with time-sequenced electronic pulses. The latent magnetic image attracts and holds powdered iron-resin ink. The ink is transferred to paper and permanently bonded by passage through a heat and pressure fix station.

The machine at left has a working speed of 125 inches per second—equivalent to 1,000 cards a minute. It can be operated for sustained periods at 160 inches per second. A full parallel printer based on these principles is capable of producing many times the output of the fastest IBM impact-type printer.

Engineering Achievement

The magnetic printer project is typical of the many opportunities for achievement awaiting engineers and scientists in advanced systems development at IBM. Perhaps you might be more interested in what IBM people are doing in such areas as: semiconductors, Teleprocessing*, computer development, optics or others. If you have an engineering degree and would like to learn more about these assignments, write, outlining briefly your experience and field of interest, to:

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NEWS New Products

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(Continued from page 112A)

Varipak printed-circuit card. This is a nine-bit register, 150kc maximum frequency, for 12V±20% dc power supply, with 6 volt input/output pulse. All ultraflex facilities are supplied.

Typical applications are tape-to-card, card-to-tape, analog-digital buffering and other short-term storage applications. Commercial and military models are available, for temperatures up to 100°C.

Circuit Breakers

A new concept in circuit breaker design—a hydraulic-magnetic tripping element—is being used by the **Westinghouse Electric Corp.,** Box 2278, Pittsburgh, Pa., for a new line of circuit interrupting devices called Hynetic* breakers. Developed for aircraft, missile and ground-support applications, the breakers can be supplied in 360 standard current ratings from 0.020 to 50 amperes, or in any fractional rating within these limits.

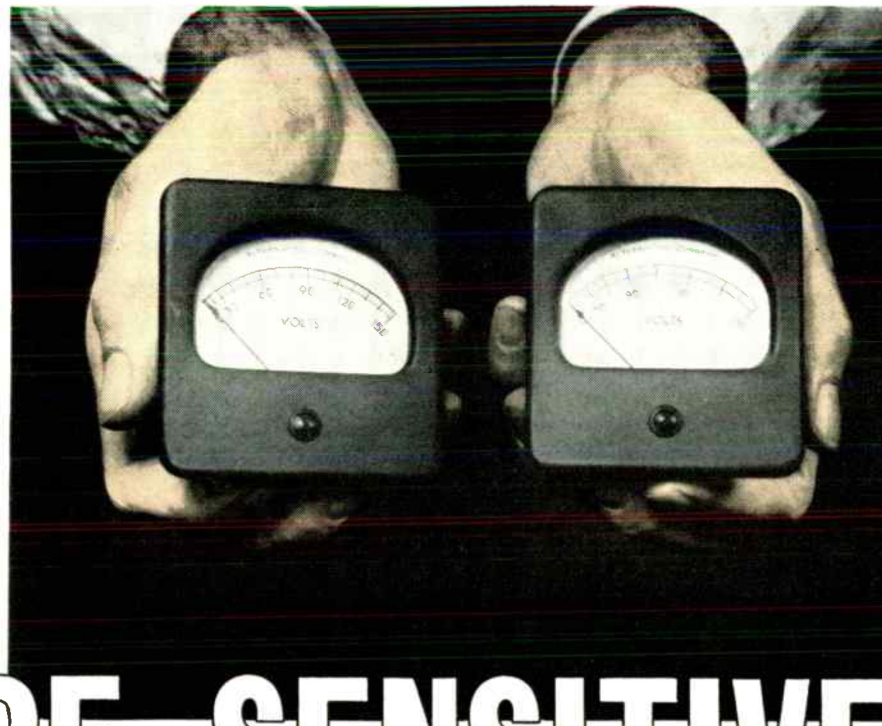
The breakers are available in three trip ranges—standard, fast, and instantaneous. The standard, or general purpose time delay, is most applicable for the protection of motors (because of their relatively long inrush time) and for wiring. The fast time delay has been designed to protect electronic circuits, including components such as vacuum tubes and transformers. For special applications that require an immediate response without an intentional time delay, the breakers can be supplied without the time delay element.

Four voltage ratings can be supplied. They are 32- or 50-volts dc, and 125 or 250 volts at 60 cps. Interrupting capacity of the breakers ranges from 1000 to 3000 amperes, depending on the voltage rating.

Available as an optional feature in a single-pole double-throw auxiliary switch, that is actuated by the breaker mechanism, but is electrically insulated from the breaker circuit. Auxiliary switches are rated five amperes at 125-volts ac, or one amp at 50-volts dc. Also optional is a handle tie to provide manual, simultaneous operation of two or more adjacent breakers.

(Continued on page 118A)

WHICH METER WOULD YOU CHOOSE TO GET MORE SENSITIVE LINE VOLTAGE READINGS?



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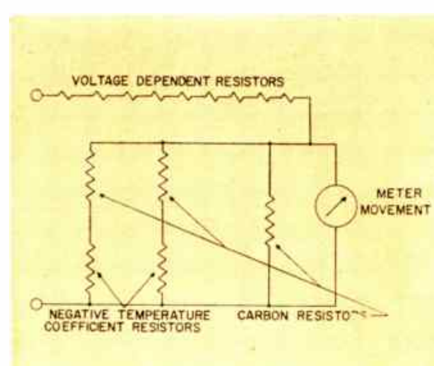
Easy to see our suggestion is well-founded. This meter is far more sensitive because it employs non-linear elements that precisely control voltage and compensate for temperature: FXC's Varistors (VDR's) and Thermistors (NTC's).

Easy to understand how these components make for more simplified circuitry . . . greater stability . . . and less susceptibility to overloads. Take a look at the circuit shown below . . .

The meter is but one of many products made better by the use of FXC's VDR's and NTC's.* Write for complete technical information — or, better still, obtain the VDR's and NTC's you'll need for making initial investigations by ordering our FERROXKIT NO. VT-1. The kit, priced at $10.00, contains the 9 Varistors and 2 Thermistors used in the circuit illustrated here.



*These same units also provide improved sensitivity in automatic feedback control circuits.

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(Continued from page 116A)

Pulse Meter—Power Supply

Airtronics International Corp., Fort Lauderdale, Fla., makes available—as an extension of its line of telephone testing equipment—a Pulse Meter and Adjustable Power Supply, Model PMVPS-1. Designed as a compact, handily portable, and versatile instrument for use in the telephone equipment room, laboratory, factory, etc., this new dual purpose unit can be used for checking the speed of square wave pulses, in the speed range of 0 to 15 pulses per second, and the width of the active component of the square wave. The meter can accept loop, battery (48–50) volts, polar duplex (alternate resistance battery and ground), or carrier (resistance battery) pulsing.

In addition, the instrument can be used as a light-duty power supply source which is continuously variable from 0 to 300 volts. A feature of the power supply is the smoothness of voltage control, particularly in the low-voltage ranges.

Random Noise Generator

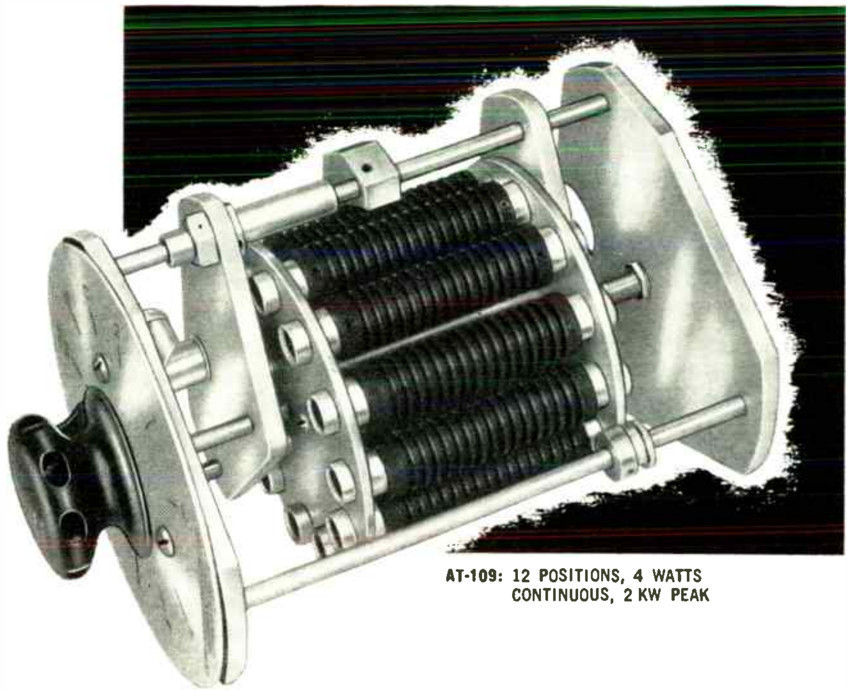
H. H. Scott, Inc., 111 Powder Mill Rd., Maynard, Mass., announces a new Random Noise Generator, Model 811-B. It features higher, more uniform output and a new "pseudo-RMS" metering circuit which reads identically on sine waves and white noise.

The 811-B has output of at least 2.5 volts RMS on all ranges, and response from 2 cps to 1.5 mc. A meter input jack enables its meter to be used for measuring other signals.

Available either as a chassis unit, in a

(Continued on page 120A)

Accuracy
Reliability
Versatility



AT-109: 12 POSITIONS, 4 WATTS CONTINUOUS, 2 KW PEAK

EMPIRE COAXIAL ATTENUATORS

AT-50: 1 W CONT., 1 KW PEAK
AT-60: 2 W CONT., 2 KW PEAK

AT-75: 4 W CONT., 2 KW PEAK

AT-106: ATTENUATOR PANEL FOR VALUES UP TO 120 DB USING COMBINATIONS OF STEP ATTENUATORS.

AT-103: 6 POSITIONS, 1 WATT CONTINUOUS, 1 KW PEAK

AT-108: 6 POSITIONS, 4 WATT CONTINUOUS, 2 KW PEAK

AT-104: 12 POSITIONS, 1 WATT CONTINUOUS, 1 KW PEAK

Rated up to 4 Watts

Empire attenuators have been designed and constructed for the reliable performance so essential to modern, complex systems. Their rugged construction . . . conservative power ratings for CW and pulse operation . . . and exclusive deposited carbon precision resistors . . . enable them to satisfy your microwave attenuation requirements.

Resistive coaxial networks shown here are intended for operation up to 4000 MC, a complete line of coaxial attenuators to 10,000 MC also available, and is described in our Catalog No. 604. Low VSWR and high accuracy are inherent features. Attenuation values up to 60 DB are obtained in individual pads, rated up to 4 watts continuous and 2 KW peak (AT-75) or as six and twelve position step attenuators (AT-108, AT-109). With two or three attenuators connected in series, values up to 120 DB can be obtained.

In many cases our engineering files contain a ready solution to your unusual coaxial attenuation problems.

For complete technical information, write for free Catalog No. 604.

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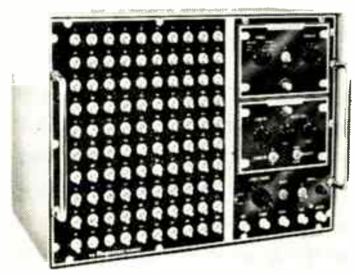
(Continued from page 118A)

cabinet, or on a 19″×5¼ rack panel, the 811-B Random Noise Generator is priced from $195 to $205.

For complete technical information on the 811-B, write to H. H. Scott, Inc., Investment Division.

Digital Pattern Generator

A new development in digital test equipment is announced by the **Magnavox Co.**, Government and Industrial Div., Fort Wayne, Ind. It is called the Digital Pattern Generator and is designed for wide application in the computer field.



The generator is a transistorized piece of test equipment which provides a simulated and flexible time division pattern output. It is designed to supply a repeating bit pattern for the testing and development of a wide range of computers and communications systems. Modular techniques have been used to allow for future expansion of present capabilities. Rugged construction, combined with conservatively designed solid state electronics, assure maximum life with a minimum of maintenance.

The unit features an adjustable pattern length from 1 to 100 bits with the state of each bit independently controlled as to "1" or "0" from the front panel. The output impedance is a nominal 500 ohms and remains constant as output level is adjusted. Positive and negative levels of 0 to ±15 volts are independently adjustable. Two clock generators are available; one variable from 10 to 100,000 bits per second, the other from 20 to 200,000 bits per second. A clock pulse is provided at the bit rate and a cycle pulse is supplied at the beginning of each pattern. For more information contact the firm.

Recorder/Reproducer

An all-solid-state magnetic tape recorder/reproducer offering features of larger instruments in a small package that can be easily transported or mounted in standard racks has been developed by the **DataTape Div., Consolidated Electrodynamics Corp.**, 360 Sierra Madre Villa, Pasadena, Calif., a subsidiary of Bell & Howell Co.

The Type PR-2300 Recorder/Reproducer is modularized, measures 22¾ inches high, 17¾ inches wide, 15 inches deep and weighs less than 150 pounds. The tape transport system is mounted in an aluminum carrying case. For vertical mounting in a 19-inch rack, the transport may be mounted either in or out of its case. For horizontal mounting, the transport is removed from the case.

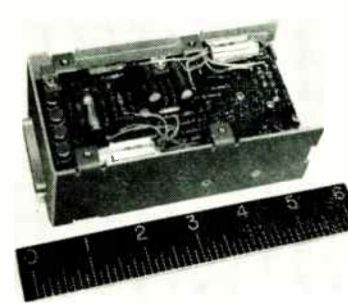
Electronics are mounted in a matching case of the same depth and width, but 3⅝ inches in height. Each case holds up to 14 plug-in record or reproduce amplifiers. One or two of these can be fastened to the transport case to build one integrated system. An electronic module case complete with 14 amplifiers weighs about 25 pounds.

Up to 14 channels of direct-record or wideband FM analog data can be recorded and played back bi-directionally. Head assemblies for two record and two reproduce heads of seven channels each can be plugged in. An accessory erase head is also available.

The Type PR-2300 records and reproduces at speeds of 60, 30, 15, 7½, 3¾, and 1⅞ inches per second. Speeds are changed in pairs by changing drive belts. Choice between the two speeds in each pair is made from push-button controls. Tape speed is held constant to within 0.25 per cent.

Input range for the direct record system is 0.25 to 25 volts rms for normal record level. Frequency response is 100 to 100,000 cps ±3 db at 60 ips. High-frequency response is proportionately lower for lower tape speeds. Signal-to-noise ratio is a minimum of 38 db from 300 cps to the highest frequency response specified for each tape speed.

DC Amplifier



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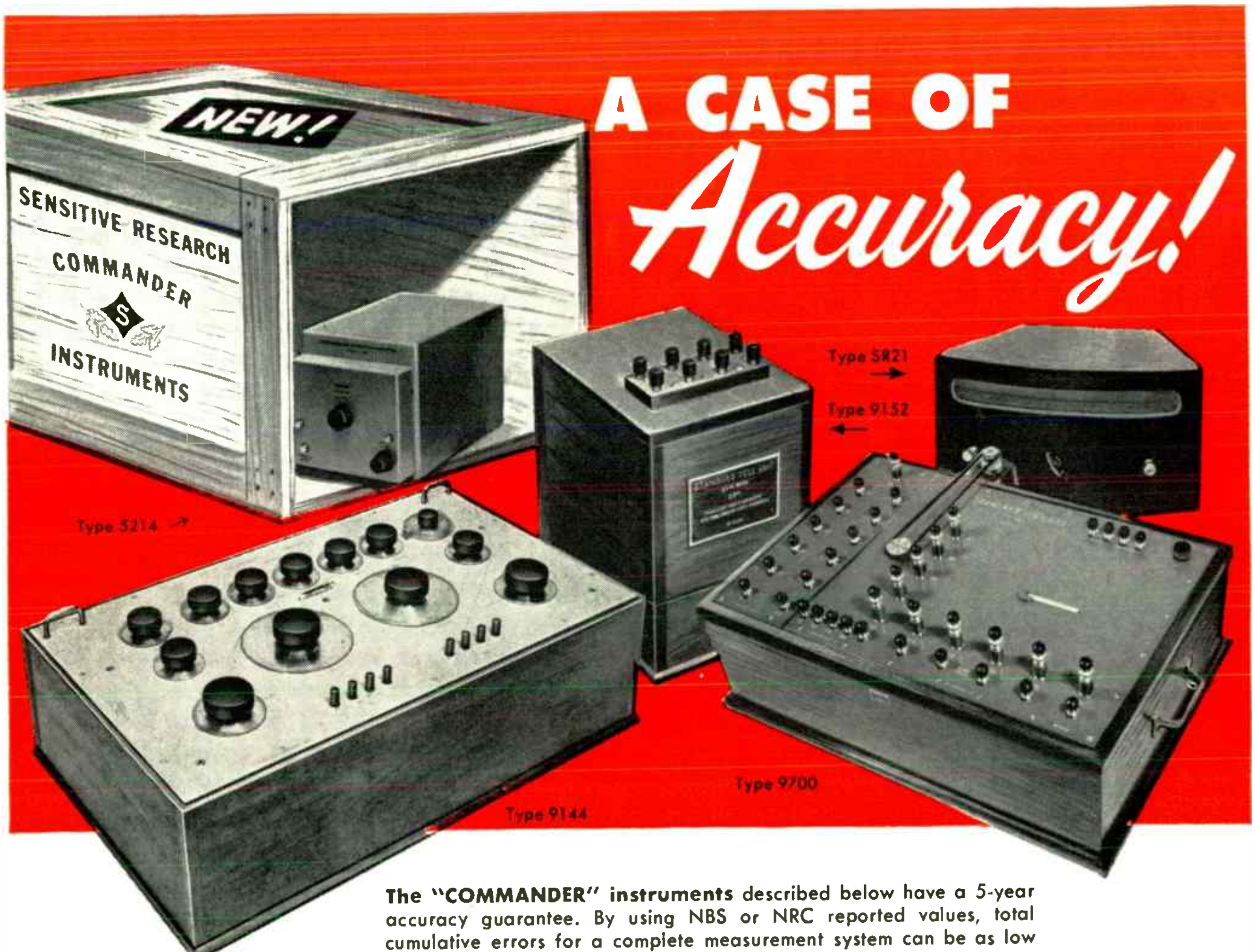

The "COMMANDER" instruments described below have a 5-year accuracy guarantee. By using NBS or NRC reported values, total cumulative errors for a complete measurement system can be as low as ±.002%.

DAUPHINEE POTENTIOMETER TYPE 9144

ACCURACY ±.001%

A 7-figure (above 1 V.) DC vernier potentiometer with a total measuring range of 2.101010V. *Accuracy is at least 10 x that of similar commercially available equipment.* Direct readout on 4 dials in increments of .1 μv (no slidewire). Thermal emf's less than .5 μv on x 1 range and .1 μv on x .1 range. Contains 2 saturated standard cells in an internally thermostated enclosure. Completely "Self-Checking." *May be used with equal facility and accuracy as a Saturated Standard Cell or Resistance Comparator — and for the calibration of .01% accurate potentiometers.*

The Type 9144 is an original design by Dr. T. M. Dauphinee of the National Research Council in Canada. All rights are protected by a United States patent application.

VOLT RATIO BOX TYPE 9700

ACCURACY ±.005%

A volt ratio box similar to that used by the National Bureau of Standards as described by NBS Research Paper RP 1419. Self-heating and surface leakage negligible. Ranges: .15/.3/.45/.6/.75/1.5/3/4.5/6/7.5/15/30/45/60/75/150/300/450/600/750 V. (Type 9700A includes 1500 V. range). Furnished in a thermostated oil bath with a motor-driven impeller.

CONSTANT TEMPERATURE STANDARD CELL ENCLOSURE TYPE 9152 AND SATURATED STANDARD CELL TYPE 4305

ACCURACY ±.001%

Enclosure accommodates up to 4 cells and is air thermostated at 28°C ± .025°C. Transistorized circuit. Operates on 110V/60 cps (Battery standby).

THE TYPE 5214 GALVANOMETER AMPLIFIER AND TYPE SR21 LIGHT SPOT GALVANOMETER.

The amplifier operates on the differential photocell principle in conjunction with a liquid-filled primary galvanometer. The secondary galvanometer has a scale length of 120-0-120 mm and is stable and free from the effects of external vibration. Over-all sensitivity is approximately 350,000 mm/μa and 35,000 mm/μv. It is ideally suited for use with the type 9144 potentiometer.

Sensitive Research "COMMANDER" Instruments are manufactured by Guildline Instruments, Ltd. (formerly Tinsley Instruments (Canada) Ltd.)

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SENSITIVE RESEARCH INSTRUMENT CORPORATION

NEW ROCHELLE, N. Y.

ELECTRICAL INSTRUMENTS OF PRECISION SINCE 1927

Symbol of Quality

NEWS New Products

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

(Continued from page 120A)

Temco Electronics, a division of **Temco Aircraft Corp.,** P.O. Box 6191, Dallas 22, Texas, announces a new dc amplifier, Model 100A, designed to convert low level dc input to ± 2.5 dc output. The unit has continuously adjustable gain between steps of 10–100 and 100–500. Features include: 0.5 per cent best straight line linearity, gain stability of 0.5 per cent full scale of 25°C value, less than 5K output impedance, and 20 mv peak-to-peak ripple (carrier). Frequency response filters are interchangeable, and the unit will meet applicable portions of MIL-E-5272. The unit measures $5 \times 2\frac{1}{2} \times 1\frac{3}{4}$ inches.

Designed for application in strain gages and thermocouples, the unit can be delivered in sixty days.

Belden Appoints Clough to Board

Herbert W. Clough, Vice President, Marketing, of **Belden Manufacturing Co.,** P.O. Box 5070-A, Chicago 80, Ill., producer of electronic and electrical wire and cable, has been elected to the company's Board of Directors.

Clough's service with Belden began in the Magnet Wire Sales Service Department in 1922. He has since held the positions of salesman, Canadian Sales Manager, Supervisor of Government Orders, organizer and first Manager of the Merchandise Sales Division and General Sales Manager. He became Vice President in Charge of Sales and Advertising in 1940 and Vice President in Charge of Marketing in 1959.

He has served as a director and Vice President of the Electronics Industries Association. He also served the Electronic Parts and Equipment Manufacturers and the Electronic Industry Show Corporation in similar capacities.

Currently he is Vice Chairman of the Wire and Cable Division of the National Electrical Manufacturers Association, a director of the National Automotive Parts Association, and a member of the Advisory Committee of the Underwriter's Laboratories.

S Band Amplifier

Menlo Park Engineering, 711 Hamilton Ave., Menlo Park, Calif., has developed an S band amplifier, Model TA-568. Its uses are in the fields of antenna pattern measurements, countermeasure equipment design, microwave power amplifiers and simulator equipment.

(Continued on page 124A)

NEWS New Products

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

(Continued from page 122A)

Specifications are: frequency range—2.0–4.0 kmc, power output—10 watts, saturated power gain—30 db, small signal gain—30 db minimum, input impedance—50 ohms, output impedance—50 ohms, spurious modulation—30 db below, signal, power supply—115 vac, 1 phase, 60 or 400 cps, 800 watts, panel space—21×14¼×24 inches, and weight—180 lbs.

3-Axis Accelerometer

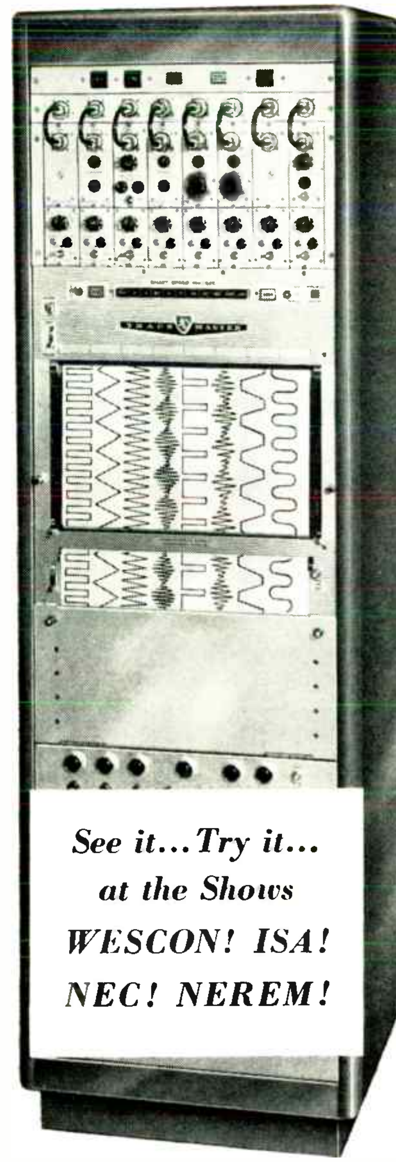
A small 3-axis accelerometer with potentiometer output is being produced for missile applications by **Humphrey, Inc.,** 2805 Canon St., San Diego 6, Calif., manufacturer of guidance instruments. This single accelerometer provides information on linear acceleration along three different axes, replacing multiple instruments at savings in size, weight and cost. Total weight of the unit is about one-half pound and size is approximately 2½" diameter and 2" long.

This firm also manufactures a 2-axis linear accelerometer that is smaller and lighter. Total weight of the 2-axis unit is four-tenths of a pound and dimensions are about 2¼" diameter and 1½" length.

Both accelerometers can be furnished in hermetically sealed cases, with an assortment of mounting provisions and a choice of connectors. The instruments have been qualified for use under severe environmental conditions in missile applications. They will withstand −65°F to 180°F while operating, relative humidity 100 per

(Continued on page 126A)

Here's why the NEW AO TRACE-MASTER is the world's finest 8-channel direct writing recorder!



American Optical Company, famous for precision instrumentation for 138 years, introduces an electronic direct-writing recorder of unique design, in which ultra-precise electromechanics has been combined with advanced electronics to achieve *truly superior performance.*

Finest Writing Method Ever

Unique direct-carbon-transfer writing method. Trace is uniformly black and up to four times thinner than that made by any other recorder. Minute variations in phenomena measured are more faithful, meaningful. Carbon trace cannot fade... may be easily reproduced.

Finest Frequency-Amplitude Performance

TRACE-MASTER'S multiple-feedback wide-range Driver circuitry, combined with the advanced pen-motor design, produces wider frequency response *at larger amplitudes* than any other recorder. TRACE-MASTER response is flat—*within 1%*—from dc to 110 cps at 40 mm!

Band Amplitude Product (i.e. Bandwidth times Amplitude) is 5600...140 cps (3 db point) x 40mm!

Finest Chart-Drive Facilities

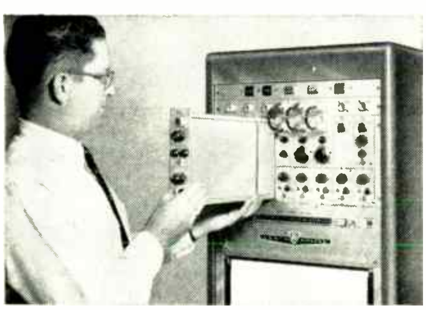
TRACE-MASTER provides widest chart-speed range...0.1 to 500 mm/sec...of any direct-writing recorder! Convenient push-button selection. Take-up reel automatically stores full 1000 ft. record. Writing table tilts for easy chart annotations. Guide rails permit quick, easy paper-roll changes. Low cost chart paper makes practical protracted recording at high speeds.

Finest Resolution, Linearity, Stability

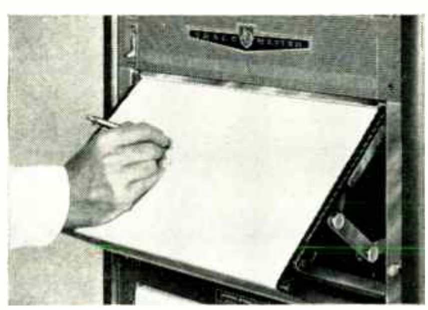
Thin carbon trace (thinner by 4 to 1 over most recorders) and high Band Amplitude Product (higher by 6 to 1 over other recorders) provide up to 24 times the resolving power or ability to detect short, sharp variations in the record. The superior linearity (± 1%) and stability in rectilinear presentation permit full use of this unexcelled resolution.

Finest Systems Oriented Compatability

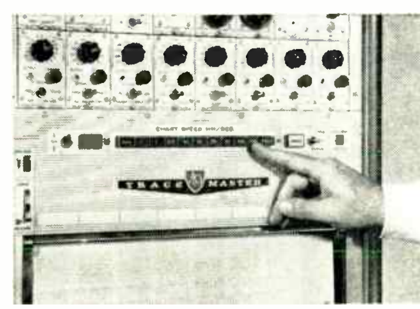
Fully transistorized circuitry...application of combined dc level and signal multiple feedback...complete interchangeability of modular signal-conditioning elements... are some of the features that make the AO TRACE-MASTER the world's finest 8-channel direct writing recorder.



Entire channel easily accessible and completely interchangeable as single unit.



Platen tilts to convenient writing angle.



Widest range chart speed... push-button selection through 0.1 mm/sec to 500 mm/sec.

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Complete Engineering Bulletins available. Field Sales Engineers at your service everywhere.

American Optical COMPANY

INSTRUMENT DIVISION, BUFFALO 15, NEW YORK

Dept. H286

Please send Engineering Bulletins on the complete AO TRACE-MASTER line of recording modules and systems.

Name__________

Business Address__________

City__________ Zone____ State________

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

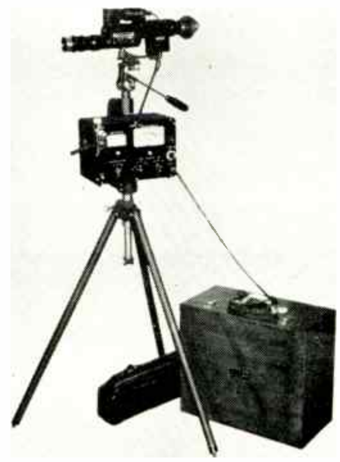
(Continued from page 124A)

cent, unlimited altitude, shock of 75 G for 6 milliseconds on any axis and acceleration of 75 G on any axis without damage.

For further technical information, write to the firm.

Infrared Optical Pyrometer

Nippon Electric Co., Ltd., 2 Shiba Mita Shikoku-Machi, Minato-Ku, Tokyo, Japan, has developed an optical pyrometer which is to be used for measuring the temperature of a distant body heated at or below 600°C. The NEC Type XV-8B, makes it possible to measure any temperature values as low as 350°C by observing the image of the heated body through a vacuum tube called the image converter tube which is highly sensitive to infrared radiation.



TP-2637C

The new pyrometer has many important applications in highly diversified fields of industry where measurement of relatively low temperature is desired. For example, the temperature of oxide cathode of vacuum tubes, temperature distribution of anode in the process of exhausting, especially of large transmitting tubes.

Specifications are as follows. Power Supply: Supply Voltage: 100 to 115 volts, Frequency: 50/60 cps, Power Required: 25 watts approximately. Performance: Range of measurement: 350° to 1000°C. (The range can be expanded to as high as 2,000°C by the use of a special optical filter.) Accuracy of measurement is better than ±1%, Effective Range is more than 20 centimeters, Image Resolution is 400 lines or more. Types and numbers of vacuum tubes and incandescent lamp used: Image converter tube Type 1P25 1, rectifier Type 12XB 1, incandescent lamp of standard brightness 1. Weight and dimensions: telescope section: 6.69″ high, 3.15″ wide, 17.32″ long. Weight 7.72 lbs. Power Supply and Indicator Section: 7.68″ high, 11.02″ wide, 5.90″ long. Weight 17.64 lbs.

(Continued on page 142A)

Timothy M. Cunningham, electron tube engineer, at the RCA Library, Harrison, N.J.

PROCEEDINGS *really takes a beating . . .* *FROM ITS READERS!*

If you've been in many company libraries, we're sure you've seen "dog-eared" copies of *Proceedings*. It's not a case of poor paper and printing—we use the best quality—it's just an example of pass-along readership taken to extremes!

Of course, 57,334 (ABC) professionally qualified men receive individual copies of *Proceedings* at home each month, as well as 13,976 students in engineering colleges. What's the reason for this important following?

Proceedings of the IRE enlisted the aid of the John Fosdick Organization to take a survey of its many readers to find out what they thought. Here are some of their reactions. "We use *Proceedings* as a reference. It's really a text. Has the largest amount in proportion of text, the highest quality text, and the largest amount of informational advertising of any book in the industry." And, of special interest to advertisers, one chief engineer said, "I've saved half-a-million dollars by buying from ads in *Proceedings*, and I've bought a quarter of a million dollars worth of equipment from the ads."

Fact is, 100% of those interviewed said they have some purchasing responsibility! And, as a further indication of the effectiveness of using *Proceedings* to reach buying factors of electronic equipment, components and supplies, these readers expressed a 2 to 1 preference for *Proceedings* when compared with mentions of all the electronic books. (Survey available upon request—ask for it!)

Are *you* now reaching this important, select audience for your product? You can, you know, reach them 12 times a year in *Proceedings* for just $8,640. (1960 rates). A similar schedule in a semi-monthly would cost $29,087.50 . . . and in a weekly, $46,280! Call or write today for all the facts. Find out what selling power *Proceedings* really has! Reach 61,957 (net paid circulation) top-level radio-electronics men each and every month throughout the year!

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IRE

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.

ELECTRICAL ENGINEER

BSEE. 1957. 1/Lt. Sig. C. 1½ years as Project Officer at WSMR, New Mexico. Experience in the field of guidance and countermeasures of missiles. Interested in obtaining an LL.B. Desires position which requires an engineering and legal combination with management opportunities. Box 2072 W.

ELECTRONICS ENGINEER

European assignment desired in western or southern Europe in management, technical or liaison area. 15 years experience R&D, both civilian and military, in communications, electronic counter-measures, television and missile electronics. Considerable supervisory experience. BEE. and MEE. degrees. Box 2073 W.

ELECTRONICS ENGINEER

B.S. in physics. MSEE. 5 years in transistorized computer circuit design and radar computer systems. Single, age 28. Desires position in Western Europe. Box 2074 W.

ENGINEERING MANAGER

Communications systems engineering of telephone, radio relay, microwave, HF and SSB Systems. 20 years military experience all phases applied engineering, siting, installation, operation and maintenance. Performed Staff Engineering assignments at Pentagon and major command levels. Currently Deputy Chief, Maintenance Engineering & Test Branch, multibillion dollar air defense project office. Supervisory experience with over 500 personnel. 21 months military electronics schooling plus BBA. and MBA. Member IRE. Desires managerial position with growth potential. Box 2075 W.

SENIOR CONSULTING ENGINEER

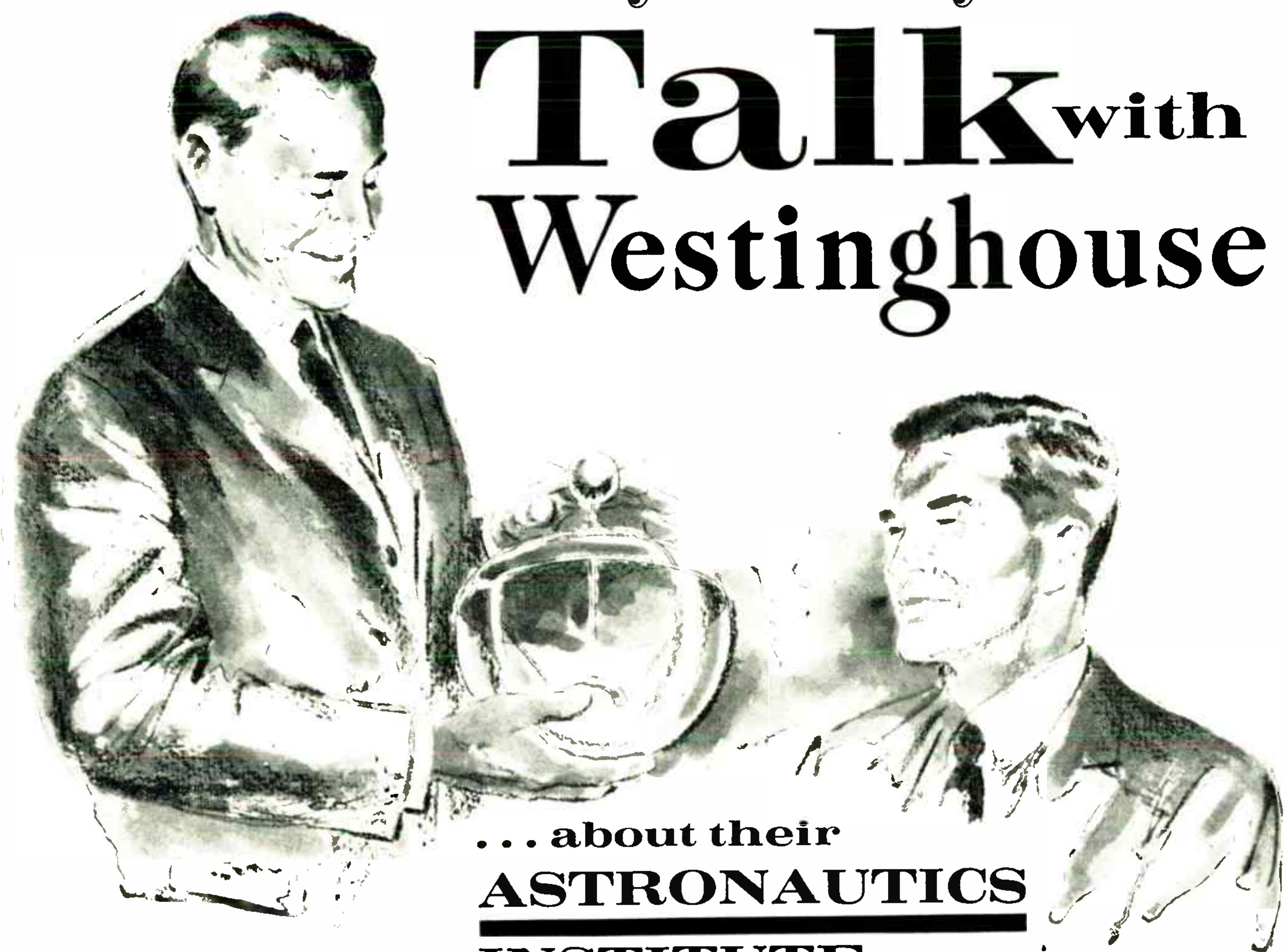
Associate Professor of Nuclear Engineering desires summer employment leading to a consulting position. BSEE, 1949; MSEE, 1951; 7 years industrial and 3 years teaching experience in electronic and nucleonic instrumentation, microwaves and radioisotope techniques; active security clearance, AEC licenses, located in southwest. Box 2077 W.

ENGINEERING MANAGER

MSEE. degree. Background includes all levels in R&D to 3 years as manager of development groups. 12 years experience in analog and digital

(*Continued on page 132A*)

Why don't you Talk with Westinghouse



...about their ASTRONAUTICS INSTITUTE

Westinghouse engineers are engaged in the study of the entire range of space operations from satellites to inter-planetary operations, including the design of space vehicles. Talk with Westinghouse about these and other advanced projects such as Typhon, and the facilities open to engineers and scientists whose careers are linked with Westinghouse . . . pioneer and leader in the electronics field.

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Underwater Acoustics
Microwave Antennas & Components
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Molecular Electronics
Systems Planning
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Send resume to: Mr. A. M. Johnston, Dept. 300

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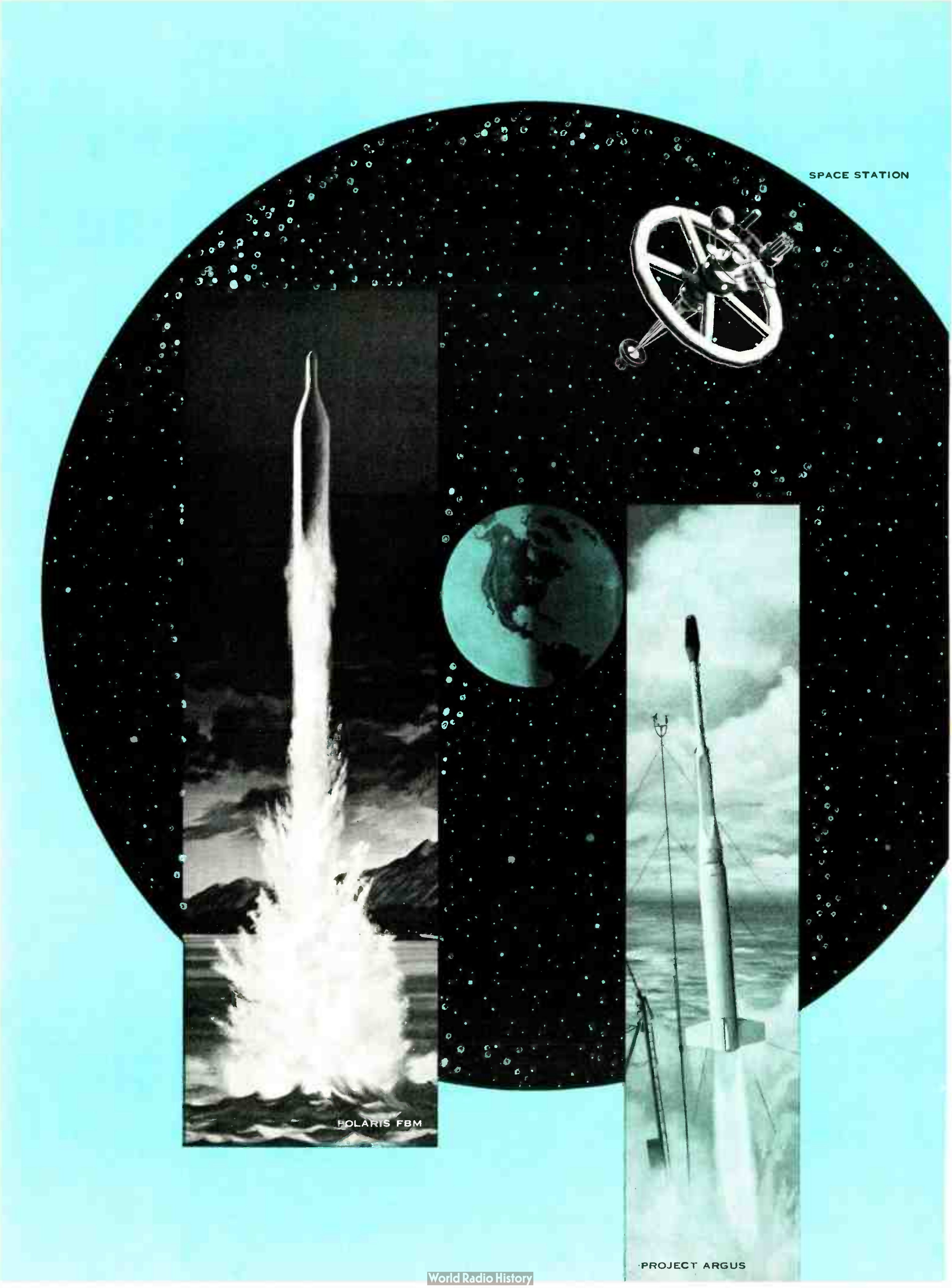
SPACE STATION
POLARIS FBM
PROJECT ARGUS

DISCOVERER SATELLITE

EXPANDING THE FRONTIERS OF SPACE TECHNOLOGY

ADVANCED PROJECTS AT LOCKHEED

POLARIS FBM Now in its advanced development status, the Navy-Lockheed POLARIS Fleet Ballistic Missile is scheduled to be fully operational and aboard its specially designed submarines late this year. Full-scale test vehicles have been successfully flown on a regular schedule of firings for months with only two failures, a remarkable achievement in view of the totally different environmental problems involved in its underwater launch. With nearly three-quarters of the earth's surface being water, practically no target in the world is outside the POLARIS' range of over 1200 nautical miles. The Division is systems manager for the POLARIS under the direction of the Special Projects Office of the Navy.

SATELLITE PROGRAMS—The Air Force-Lockheed AGENA satellite is a versatile space vehicle capable of numerous assignments. In its present DISCOVERER program configuration, it is 19 feet long, 5 feet in diameter with an orbital weight of approximately 1700 pounds. Payload of several hundred pounds includes telemetry, instrumentation, guidance and attitude control systems, reentry vehicle and recovery capsule. The AGENA has accomplished several significant space "firsts." It was first to be placed on the difficult polar orbit; first to be placed on a precise, predicted, and nearly circular orbit; first to change its attitude on orbit, with a turn of 180 degrees and a downward tilt of 60 degrees; first to eject a capsule; and first to prove advanced space systems such as ground-space communications, instrumentation, attitude and guidance and life-sustaining devices. The AGENA can be modified for a variety of space missions such as navigation; geophysical investigations; lunar probes; long-range communications; and space probes.

In addition to the DISCOVERER program, the Division is developing advanced AGENA satellites for the MIDAS program (Missile Defense Alarm System) and the SAMOS strategic warning system. Lockheed is system manager and prime contractor for these projects under the direction of the Air Force Ballistic Missile Division (ARDC).

SPACE STATION—An orbiting research facility to serve as an advanced base for space exploration, has been proposed in practical detail by Lockheed's research and development staff. The station would carry a 10-man crew. Prefabricated compartments for the rim of the wheel, the spokes, and the three hubs would be launched separately by ballistic missiles and assembled in space by means of the specially-designed, Lockheed Astrotug.

X-17, KINGFISHER, X-7—The Air Force-Lockheed X-17 solid-propellant ballistic missile has pioneered many new techniques, and the valuable experience gained from this program facilitated development of other, inter-service projects, including the Navy POLARIS FBM. The Navy's Project Argus radiation explosion featured the X-17 as the vehicle. Developed for the Air Force, the Lockheed KINGFISHER is designed to simulate enemy attacks to test our nation's anti-bomber and anti-guided-missile defenses. The Air Force X-7 is a unique, recoverable ramjet-engine test vehicle designed to test new developments in advanced components for other missiles.

The successful completion of projects such as these requires a bold and imaginative approach to entirely new environments. Lockheed's programs reach far into the future. It is a rewarding future which scientists and engineers of outstanding talent and inquiring mind are invited to share. Write: Research and Development Staff, Dept. H-33, 962 W. El Camino Real, Sunnyvale, California. U.S. citizenship or existing Department of Defense industrial security clearance required.

Lockheed / MISSILES AND SPACE DIVISION

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CAPE CANAVERAL, FLORIDA • ALAMOGORDO, NEW MEXICO • HAWAII

Positions Wanted

By Armed Forces Veterans

(Continued from page 128A)

instrumentation, data processing, and automation with an even division between missile and industrial fields. Will relocate to either coast. Box 2078 W.

ELECTRONICS ENGINEER

Mature, very active electronics engineer anticipating retirement from technical military duties at 50 interested in challenging engineering management position; 7 years complex electric power service problems; 19 years broad Navy engineering management duties in diverse U.S. and overseas communications and anti-submarine electronics systems projects and construction, scheduling work. Trained to relocate periodically on short notice and quickly assume specific responsibility for important technical projects requiring smooth coordination with other staff specialists. Outstanding references and accomplishments. Recent technical refresher training. Particularly interested in Phila., Pa. or Calif. areas, but will consider any long range opportunity. Box 2079 W.

ENGINEER

Supervisory engineer of systems or sales on systems management or evaluation of electro-mechanical weapons system equipment in radar, computers, etc. and microwave components. Have excellent market and sales background with industrial and Government agencies on prime hardware or components. Home base New York City or Long Island. Will travel extensively if necessary. Matured. Formerly Lt. Colonel in the USAF. Box 2080 W.

TEACHING

British, age 35; Member IRE, IEE. Telecommunications engineering diploma. 10 years lecturing, 4 years as head of department. Seeks U.S. teaching appointment in E.E. Dept. where facilities exist for working towards MS. degree. Experienced administrator. Visited U.S. 1959. Familiar with E.E. curricula. Specialties: circuitry and measurements. Frustrated by bureaucracy. Box 2081 W.

ENGINEER

Communications equipment, design circuitry and hardware. BSEE, MSEE. Seeks assignment Continental Europe. Top Drawer company only. Age 34; 9 years experience. Box 2082 W.

ELECTRICAL ENGINEER

5 yr. E.E. plus good Liberal Arts background. Polyglot; age 31; family; lifelong U.S. citizen. Diverse experience ranging from power during military, 3½ years design and manufacture electric, electronic instrumentation and control. 2 years solid state R and D, 1 year specifications and estimating for contracts and projects, 4½ years supervisory, presently project level problem solver and trouble shooter for large company on communication, navigation, ultrasonic and control. Box 2084 W.

ADMINISTRATIVE ENGINEER

Desires position of responsibility as Technical and Administrative Assistant to Director of Engineering or General Manager in U.S. or overseas. 10 years varied experience in Military and Comm. Elec. Systems. Shared responsibility for 5 million European Plant. Resume sent to management. Box 2091 W.

(Continued on page 134A)

CALIFORNIA

AGAIN THE LAND OF PROMISE

Science has brought us to a unique vantage point from which we can foresee vast technological change. As the curve of scientific progress approaches the vertical, today's pioneers who wish to contribute significantly in their chosen fields will do well to consider the unequalled challenge offered by association with one of the nation's foremost scientific laboratories. Positions are currently available in physics, electronics, mathematics, and in mechanical and aeronautical engineering, as well as in other technical disciplines.*

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1030 East Green Street
Pasadena, California

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- **U. S. Naval Ordnance Test Station at China Lake and Pasadena:** Developers of guided missiles, rockets, advanced propulsion systems, torpedoes, and other undersea weapons.
- **U. S. Naval Ordnance Laboratory at Corona:** Developers of guidance and telemetry systems and other missile system components. Researchers in such fields as IR spectroscopy, magnetism and semiconductors.
- **U. S. Naval Radiological Defense Laboratory at San Francisco:** The nation's only major center for research on nuclear effects and development of counter measures.
- **U. S. Navy Electronics Laboratory at San Diego:** One of the Navy's largest organizations engaged in the research and development of radar, sonar, radio and acoustics.

*These positions are in the Career Civil Service.

Positions Wanted

By Armed Forces Veterans

(*Continued from page 132A*)

ELECTRICAL ENGINEER

Four years active duty as Naval Officer. MSEE. University of Illinois 1958. Last 2 years spent in development of high power klystron transmission systems and systems controls. Desires position in system management or administration in medium sized concern. Box 2092 W.

ELECTRONICS ENGINEER

Position desired in commercial product or industrial automation fields. 15 years experience in data processing and display fields. M.S. in E.E. Preferred location—Southwest. Box 2093 W.

ENGINEER

Age 27; veteran; married, family. BS.: Electronic Physics 1958; military electronic technician experience. 2½ years experience in Solid State Research. Measurements, Metallurgy. Desires position not wholly technical with managerial future. Will relocate. Box 2099 W.

DIRECTOR SALES

13 years in communications, engineering and sales. Interested in a much greater opportunity to promote and sell HF-SSB equipments and systems. Have contacts in Marine, Government and some of the largest commercial companies on the Atlantic Seaboard. Primary interest and objective, Sales. Box 3000 W.

Positions Open

The following positions of interest to IRE members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

Proceedings of the IRE
1 East 79th St., New York 21, N.Y.

ENGINEERS—PHYSICISTS

Company engaged in research, production and development of camera tubes, infrared devices and image amplifiers seeks graduate engineers and physicists to work in new laboratories at beautiful Fort Washington, Pa. Reply to Box 2020.

PROFESSOR

Rapidly expanding engineering school located in snow-free, smog-free area with ideal climate needs electronic engineer to teach undergraduate courses and to help develop graduate program. MS. minimum requirement. Salary dependent upon industrial experience and graduate degree. Modern facilities and equipment with new building under construction. Apply to Prof. M. P. Capp, Engineering Div., San Diego State College, San Diego 15, Calif.

TECHNICAL MANAGER

Advanced Data Processing Systems. Direct planning requirements for organization of com-

(*Continued on page 136A*)

Positions Open

(Continued from page 134A)

prehensive business information processing systems. Requires a Ph.D. in E.E., physics or math, plus several years experience in digital data processing systems. Managerial experience desirable. Attractive salary. Forward resume to Box 2021.

ASSISTANT PROFESSOR

Assistant Professor of Electrical Engineering. Ph.D. required with interest in circuit theory, physical electronics or solid state. Active research and graduate program. Contact W. P. Smith, Chairman, E.E. Dept., University of Kansas, Lawrence, Kansas.

ELECTRICAL OR ELECTRONICS ENGINEER

Large medical institution is interested in employing an engineer to direct the maintenance and trouble-shooting of a large variety of complex electrical instrumentation, and to make modifications in existing equipment. Position requires a person with a degree in E.E. and a minimum of 4 years experience in one or more of the following fields: electrical instrumentation; field engineering or electrical instruments; computers; f.m. tape equipment; telemetering equipment; radiation equipment and counters. Write Personnel Section, Mayo Clinic, Rochester, Minnesota.

PHYSICIST

M.S. or Ph.D. Solid background in both optics and electronics. Research on techniques and phenomena for application to space navigation. Send resume to Personnel Director, The Franklin Institute, Phila. 3, Pa.

RADIO ENGINEERS

Design and development positions for Circuit Designers, Development Engineers, Antenna Engineers in connection with direction finding equipment, ECM devices, advanced communication systems, air traffic control, multi-channel receiver development, advanced filter design, R.F. systems analysis and simulation. Apply Servo Corporation of America, 111 New South Road, Hicksville, L.I., New York.

ENGINEERS

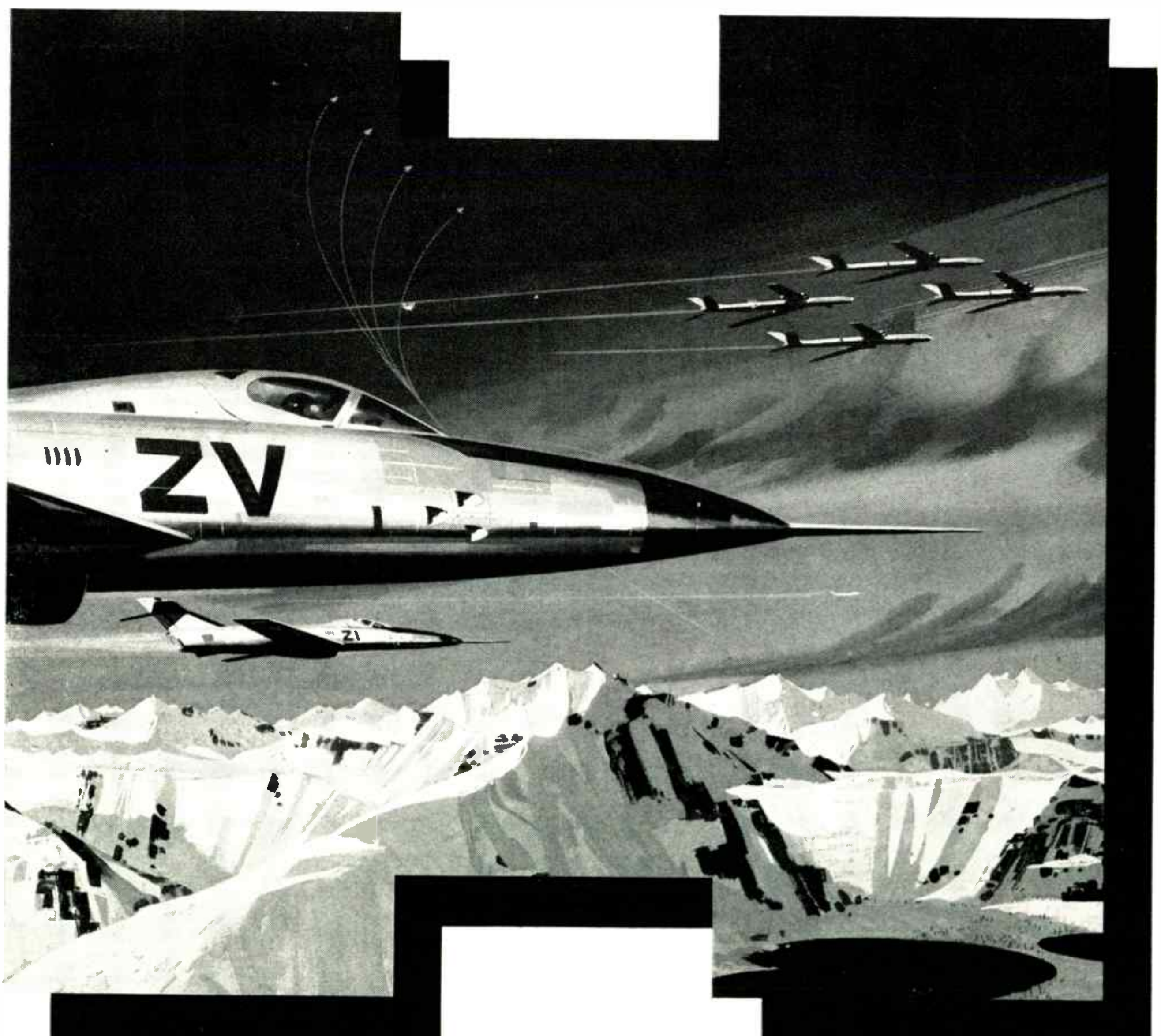
Electrical engineering staff openings in an expanding undergraduate program. (1) Instructor: to teach undergraduate courses to sophomore and junior engineering students. Candidate must have B.S. degree and be willing to take graduate work at nearby university. (2) Assistant, associate or full professor: to teach electronics and senior E.E. electives. Candidate should possess a graduate degree (M.S. or Ph.D.) specializing in electronics and/or communications. Encouragement given to private research and consultation. Rank and salary commensurate with academic attainment and experience. Apply: L. M. Gonsalves, chairman, Dept. of E.E., New Bedford Institute of Technology, New Bedford, Mass.

ASSISTANT OR ASSOCIATE PROFESSOR

Assistant or Associate Professor for E.E. Dept. M.S. or Ph.D. degree required. Opening in electronics. Salary commensurate with qualifications. Position available Sept. 1960. Write, Head, E.E. Dept., South Dakota School of Mines and Technology, Rapid City, South Dakota.

(Continued on page 140A)

How will this picture look



Positions Open

(Continued from page 136A)

SENIOR ANALOG COMPUTER ENGINEER

Milgo Electronic Corporation of Miami, Florida has an opening for a senior level engineer, with a minimum of 5 years experience in the field of analog circuitry design. This individual would become the nucleus of a new products design group, with unlimited opportunity for advancement. Interested candidates forward resumes to Mr. R. H. Mattox, Milgo Electronic Corp., 7620 N.W. 36th Ave., Miami 47, Florida.

ELECTRONIC ENGINEER—PHYSICISTS

The Cambridge Electron Accelerator, a joint undertaking of Massachusetts Institute of Technology and Harvard University, has openings for electronic engineers for research and development in instrumentation and circuitry from d.c. to millimicrosecond; physicists interested in instrumentation and nucleonics. 2 years minimum experience. Citizenship not required. Applications from abroad also invited. Write to Director, Cambridge Electron Accelerator, Harvard University, 42 Oxford St., Cambridge 38, Mass.

PROJECT ENGINEER

Project Engineer with a diversified background in electromechanical devices and with the capability of learning new fields rapidly. Will be responsible for taking a basic design idea and translating it into a production prototype. Must coordinate activities between the research and production groups and handle customer contacts. Block Associates, Inc. 395 Putnam Ave., Cambridge 39, Mass.

ELECTRONIC ENGINEER

Electronic Engineer with experience in fabrication of miniature transistorized subassemblies. Must supervise the fabrication, troubleshooting and inspection of electronic assemblies. Also should be capable of making design changes to improve reliability and reduce costs. Apply Block Associates, Inc., 395 Putnam Ave., Cambridge 39, Mass.

ENGINEER

ACADEMIC: Ph.D. in electrical engineering or physics with a strong background in instrumentation and/or process control who will teach graduate courses only leading to MS. degree in Instrumental Science. Expected to develop own research program. May start any time up to Sept. Rank and salary dependent on qualifications. Graduate Assistantships available in the same area. Send resume and references to Head, Electronics & Instrumentation Graduate Institute of Technology, University of Arkansas, Little Rock, Arkansas.

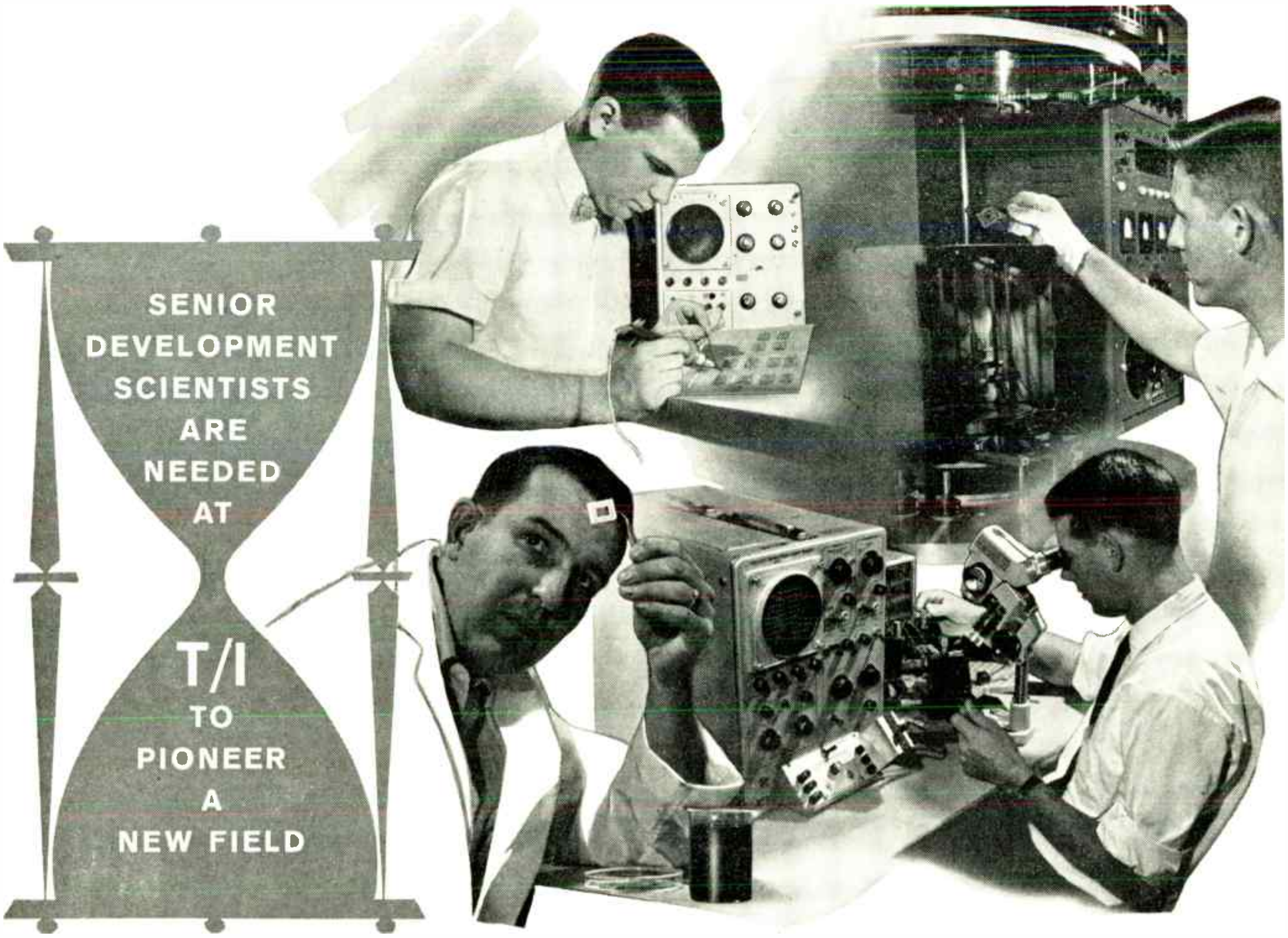
TRANSLATORS

Proven ability to translate technical material into fluent English essential. Attractive full-time or free-lance arrangement. All languages of interest, particularly Russian and Japanese. Send resume to A.T.S., Inc., Drawer 271, East Orange, New Jersey.

ELECTRICAL ENGINEERS—PHYSICISTS (PH.D)

Announcing the creation of a new position as Director of University Relations with progressive, long established electronics firm located in

(Continued on page 142A)



THE GOAL is microelectronics. **THE PROBLEM** is to provide microminiature circuits which are more reliable...smaller in size...lighter in weight...lower in cost...than conventional components. **THE SOLUTION** is *Solid Circuit** semiconductor networks—a new and rapidly expanding field at Texas Instruments.

This wide-ranging project — the opening of a true frontier — requires continuing and new investigations. Explorations involve such techniques as solid state diffusion, alloying of metals and semiconductors, vacuum deposition of metals, semiconductor surface chemistry, solid state physical measurements. Immediate creative application of skills in various sciences is required: solid state physics, physical chemistry, inorganic chemistry, metallurgy, electronics, and mechanical engineering.

The need is for the scientist or engineer sufficiently experienced that he can explore this project from his own viewpoint and make immediate, significant contribution. Depending entirely on his own qualifications, he may either join a semiconductor network team — or he may take charge of a group. The opportunity for leadership — whether immediate or in the future — is here.

The desire to see the full semiconductor technology . . . curiosity concerning both circuits and devices . . . the ability to direct and inspire . . . these drives will advance the scientist at TI.

*Trademark of Texas Instruments Incorporated.

This — in actual size — is an interconnected stack of 11 semiconductor networks. *Solid Circuit* semiconductor networks are complete electronic circuits synthesized within a semiconductor material. By selectively diffusing and shaping conductance paths in this material, semiconductor networks have been designed to perform such circuit functions as amplification, switching, counting, pulse generation, etc. In addition to effecting a significant advance in microelectronics, semiconductor networks provide improved reliability and performance. This TI development is now producing devices which are being evaluated for satellite, missile and airborne applications.

INTERVIEWS are scheduled for your area. If the challenge and opportunity of the semiconductor network field at TI intrigues you, please send a confidential resume immediately to C. A. BESIO, Dept. 127.

SUCCEED IN THE '60s WITH TI

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Please send me the booklet *TIPS* containing details on career openings in my field at TI S-C and information on living in Dallas.

NAME__________

ADDRESS__________

CITY__________ STATE__________

My professional field is__________

My specialty is__________

Positions Open

(*Continued from page 140A*)

the East. The Director will represent the office of Vice Pres. of Engineering and Research in establishing and maintaining strong scientific relations with university and professional societies. Will develop policies and plan aspects of educational fellowships, cooperative and graduate study programs. This is a prestige position for a physicist or an electrical engineer at the Ph.D. level and will carry a sufficiently attractive salary to interest a discriminating scientist. Inquires will be treated in the utmost confidence. Box 2025.

NEWS New Products

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

(*Continued from page 126A*)

Solid State Transducers

A number of industrial organizations, including **Statham Instruments, Inc.**, 12401 W. Olympic Blvd., Los Angeles 64, Calif., have been engaged in the design of transducers for the measurements of acceleration, pressure and displacement in which the latest solid state sensors may be incorporated. However, the peculiar thermal characteristics of semiconductor materials have delayed the appearance of a solid state transducer on the market.

Evaluations of the solid state transducer prototypes indicate that the thermal coefficient of sensitivity is in the order of 0.015%/°F and that the thermal zero shift is less than 0.01 per cent of full scale per degree Fahrenheit. The temperature curve for the Model PG313-5M-70, serial number 12, illustrates the typical thermal behavior of the new solid state transducer. Furthermore, a transfer function of 0.2 volt per volt excitation is entirely feasible. Therefore, with an excitation of 25 volts an output of 5 volts can be realized. This output level makes the instrument fully

(*Continued on page 144A*)

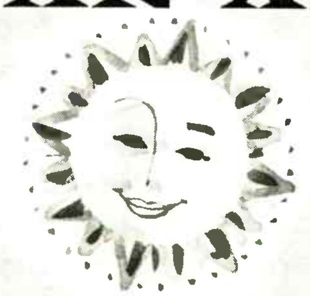
wanted:

engineers, physicists and mathematicians

FOR CAREERS WITH

GUIDED MISSILES RANGE DIVISION

PAN AM



in **Florida**

Pan Am's Guided Missiles Range Division (*GMRD*) is a technical organization with operations at Patrick Air Force Base, Cape Canaveral and the chain of down range tracking stations through the Carribean to Ascension Island.

For the past seven years, GMRD has been responsible for planning, development and operation of the Atlantic Missile Range—the largest and most complex test range in the world.

There are now new, immediate, career positions at the Atlantic Missile Range for Electronic Engineers, Mathematicians, Systems Engineers and Physicists with B.S., M.S. and Ph.D. degrees with experience in one or more of the following fields:

CW-RADAR
PULSE RADAR
COMMAND EQUIPMENT
ANALOG DISPLAY EQUIPMENT
RADIO COMMUNICATIONS
INFRA-RED EQUIPMENT
METRIC OPTICS
INSTRUMENTATION PLANNING
OPTICAL DATA REDUCTION
ANALOG AND DIGITAL DATA HANDLING

Investigate the several advantages of working with the Guided Missiles Range Division of Pan American, including the unique advantage of 90% world-wide air travel discounts.

Address your resume, including telephone number, in confidence to:

Dr. Gilbert S. Blevins
Dept. C-31, Guided Missiles Range Division
Pan American World Airways, Inc.
Patrick Air Force Base, Florida

PAN AM

GUIDED MISSILES RANGE DIVISION

PATRICK AIR FORCE BASE, FLORIDA

NEWS New Products

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

(Continued from page 142A)

compatible with existing telemetry and industrial receiving instruments.

This solid state transducer is not in production and is not available for immediate delivery. Although the firm is said to have solved the technical problems of performance, the manufacturing costs are relatively high. In order to reduce these costs, they have undertaken a program to improve production techniques. Information regarding the commercial availability of these units will be released in the near future.

The firm states that they have solved the technical problems of design and construction as well as the temperature performance of solid state transducers. The combination of the proven performance of strain gage devices and the high output (30 to 40 times greater than conventional units) of the solid state transducer represents an instrument which will be in large demand in industrial processes and automation and for the ever-increasing missile and space investigation programs.

Single-Sideband Exciter/Driver

Kahn Research Laboratories, Inc., 81 S. Bergen Place, Freeport, N. Y., has a production model SSB-58-1A Single-Sideband Exciter/Driver system for high level AM ground station transmitters which offers a practical and economical solution to military high power and super power SSB communications. Based upon high efficiency Class C sideband amplification, its advantages over conventional linear SSB systems have been confirmed by extensive commercial and government use during the past seven years. Not only is it designed for use with new AM transmitters at reduction of two-to-one or more in overall SSB system costs, but existing AM and CW transmitters can be converted to SSB operation at even greater savings.

Improved undesired sideband rejection at higher power levels, reduced size and weight of equipment, lower tube costs and relative insensitivity to varying antenna loads and tuning errors are other major advantages of Class C SSB. Either single or two independent 6 kc sidebands are developed for use with doppler-free multichannel FSK teletype, data, voice and facsimile transmissions. Operating frequency range covers 1 to 30 mc. Also available as extremely compact system for low power AM aircraft or shipboard transmitters.

(Continued on page 147A)

NEWS New Products

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

(*Continued from page 144A*)

Microwave Attenuator

Antenna & Radome Research Associates (ARRA), 27 Bond St., Westbury, N. Y., has introduced a miniature "S" band continuously variable attenuator with an insertion loss of less than 0.5 db. This unit is 3" in diameter by $1\frac{1}{8}$" high and weighs about 14 ounces. Designed primarily for airborne applications, this unit

features a simple locking device and a dial plate calibrated in degrees to simplify resetting. Design is of the Pi Line Vamp type with full attenuation achieved in less than 180° of revolution. VSWR for bandwidths of 15% or less is 1.3 maximum on the input end. Total attenuation available is 20 db minimum at about 3 kmc. Connectors are type N male. Average power capacity is 10 watts. Peak power is 5 kw. Delivery is about four weeks.

Telemetry Preamplifier

Using newly developed high gain bandwidth ceramic triodes, the TP-5 Telemetry Preamplifier, a product of **LEL, Inc.**, 380 Oak St., Copiague, L. I., N. Y., provides a gain of 26 db, a nominal noise figure of 3.5 db over the passband of 215–260 mc telemetering band. The integral power supply provides regulated heater voltage. The entire assembly, contained in a weatherproof housing for antenna tower mounting, requires a mounting space of $6\frac{1}{2}" \times 6\frac{1}{2}" \times 6"$ and weighs 6 lbs. The TP-5 is also available rack mounted.

(*Continued on page 148A*)

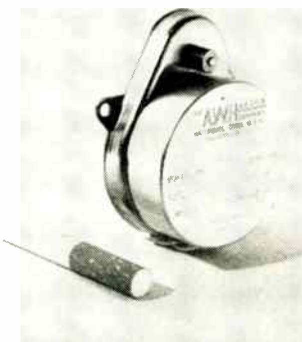
NEWS New Products

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

(Continued from page 147A)

Miniaturized Motor

A miniaturized motor, $\frac{7}{8}''$ long and designed to allow the direction of rotation to be established by electrical control alone, is the latest addition to the line of synchronous ac timing motors manufactured by **The A. W. Haydon Co.,** 232 N. Elm St., Waterbury, Conn.



Designated as the 42100 Series Commercial 60-cps ac timing motor, this five-ounce device was designed for applications where reliability, performance and space are of prime consideration. Currently being used in many applications including the company's own commercial programmers, long-life repeat cycle timers, time delay relays, elapsed time indicators, stop clocks, and so forth, the motor provides fast starting and stopping, eliminating pre-starting and clutching in many applications.

The company claims these short motors are available with no sacrifice in running torque, nominally specified as 30 ounce/inches. Available in voltages ranging from 6 to 230 volts at 20 ma maximum, the unit's rotor speed is 300 rpm with output speeds of 300 rpm to $\frac{1}{6}$ rph. Maximum temperature rise is 40°C, while power input is stated at 2.5 watts maximum. (All specifications stated at a frequency of 60 cps. A 25 or 50 cps version is also available.)

Two basic gear trains are available, with output torque independent of gear train material. The standard gear train has brass wheels and steel pinions, while the optional hard steel gear train will carry up to three times normal load, thereby obtaining greater usable torque. The standard shaft is $\frac{1}{8}$ inch diameter by $\frac{3}{8}$ inch long. A large number of variations to the standard shaft are offered and special shafts can be supplied on demand.

Additional applications for this light weight motor are in chart drives, control equipment, and any application requiring an accurate time base or source of rotary timing power.

(Continued on page 150A)

NEWS New Products

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

(*Continued from page 148A*)

Oscilloscope Cameras

Analab Instrument Corp., 30 Canfield Rd., Cedar Grove, N. J., has announced the availability from stock of their new line of oscilloscope cameras featuring binocular viewing. Polaroid® film backs, a choice of three lenses, and a wide variety of accessories add to the versatility of the cameras. In addition, the cameras offer flexibility through the use of a building-block design that permits the user to start out with a basic camera and later alter its characteristics.

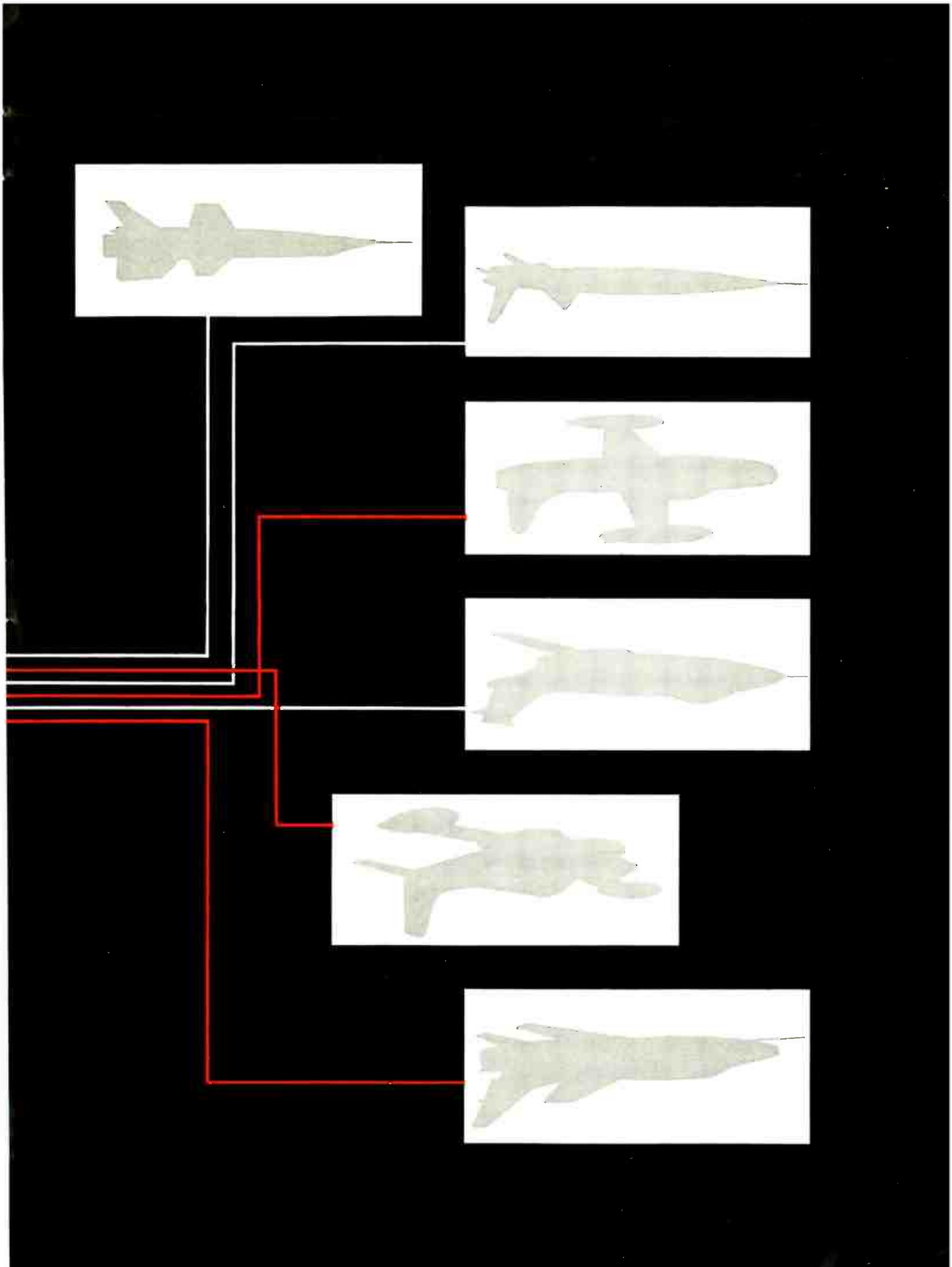
Analab's first units are the Types 3000, 3001, 3002, and 3003 Oscilloscope Cameras designed for use on all existing oscilloscopes. The basic building block for all cameras is the binocular-viewing periscope that mounts on the CRT bezel of the oscilloscope. The periscope contains a dichroic mirror that reflects 80% of the light from the CRT screen of the camera and transmits 20% of the light for simultaneous binocular viewing without parallax.

The camera unit mounts on top of the periscope by means of a slip hinge that permits the two units to be separated readily. The slip hinge arrangement also permits easy access to the lens and shutter.

A choice of either of two mounting plates for the variety of camera backs permits the selection of either a 1:07 or a 1:09 object-to-image ratio. The camera backs all slide into position on the mounting plates. A detent mechanism permits either three or ten equally-spaced positions so that the camera can record multiple exposures on a single frame.

The Polaroid camera back permits finished-print recording in one minute on paper-based films and in two minutes with positive transparency film. A low-cost special adapter lens provides off-scope recording. With this "Lensette" adapter the user may photograph his lab-bench setup for each experiment and paste this photograph

(*Continued on page 151A*)



working...
with systems that know how to adapt

Like man—Honeywell's Adaptive Autopilot can accommodate, in fact, can adapt. Anthropomorphic? Perhaps—nonetheless, this system has the capacity to change its own parameters through an internal process of measurement, evaluation, and adjustment. It operates independently of air data information and complex gain scheduling—and is unaffected by changes in aerodynamic characteristics. Simply, it adjusts itself in response to its own performance.

Without major modifications, the Adaptive Autopilot can be adapted to air vehicles of all types—from business aircraft, helicopters, drones, supersonic fighters and bombers, and missiles—to the latest hypersonic research vehicles.

Adaptive flight control systems are another example of Honeywell Aeronautical's accomplishments in advanced control. Other developments include electrically suspended gyros, guidance control and environmental control systems, and "decision-making" pre-launch checkers. The division's competence has been demonstrated by contributions to Sergeant, Thor, Atlas, Titan, Polaris, Centaur, Wagtail, F-104, B-58, WS 117L, X-15, F-101, B-66 and WF2. Current expansion has created openings for engineers and scientists in these and similar programs throughout the country and abroad.

living...
with people that know how to live

Examples of the variety of aircraft and missiles that can be flown with Honeywell Adaptive Autopilots

Honeywell engineers and their families find facilities are readily accessible for the things they enjoy. Those who appreciate the arts enjoy the Twin Cities as a cultural center . . . listening to the Minneapolis Symphony Orchestra—browsing through the Minneapolis Institute of Art or Walker Art Center—enjoying opera and ballet at Northrup Memorial Auditorium, or legitimate theatre at one of the area's fine playhouses.

For the sports minded, the Twin Cities offer Big Ten Football, professional hockey, and in 1961, National League Football. And with its Metropolitan Stadium, one of the country's finest baseball parks, major league baseball seems inevitably near.

Outdoor enthusiasts enjoy this area for its easy access to hunting, fishing, golfing, tennis, swimming, boating—or in winter, skating, skiing, sledding, ice fishing, and ice boating. And the popularity of sailing and yachting is evidenced by the existence of many fine clubs of this type.

These are some of the activities that make this an area for "living." And they're an important part of the life enjoyed by Honeywell engineers and their families.

Inquiries on opportunities with Honeywell Aeronautical will get prompt and confidential attention. Write Mr. J. H. Burg, Dept. 801, Aeronautical Division, 1433 Stinson Blvd., Minneapolis 13, Minnesota.

Honeywell

H Military Products Group

To explore professional opportunities in other Honeywell operations coast to coast, send your application in confidence to H. K. Ekstrom, Honeywell, Minneapolis 8, Minn.

NEWS New Products

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

(Continued from page 150A)

in his notebook along with the actual oscillograms also recorded. This provides ready reference should it ever be necessary to repeat the experiment.

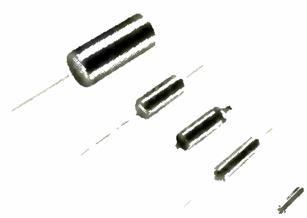
In addition to the Polaroid type camera backs, a complete series of Graflex backs may be used with these cameras. A special Analab adapter permits the use of either a Graflex roll-film holder, a film-pack holder, or a cut-film holder to accommodate conventional films for high-speed scope photography.

The Types 3000, 3001, 3002, and 3003 Cameras are all identical except for the lens used and the degree of minification. A complete Analab scope camera featuring a Polaroid back with an f/1.9 lens is \$405.00. The same system with a newly developed f/1.9 flatfield lens that has minimal field distortion even at wide apertures is \$445.00.

A Type 3670 carrying case is available for \$35.00.

High Altitude Capacitors

A special high altitude line of hermetically sealed mylar film capacitors has been introduced by **Scientific Electronics, Inc.,** 3810 Colasset St., Burbank, Calif.



This line of HI-LAR capacitors is designed for high altitude space probe applications where exceptional reliability and durability are required. A design combination of mylar high insulation characteristics and high dielectric strength permit the HI-LAR capacitors to be employed in critical circuitry where compactness is necessary. In addition to space age applications, these components are being used in instrument quality recorders, pulse networks and computers. The line is available in 100, 200, 400 and 600 volt increments in range from 0.0047 to 1.0 μf.

Because of the special construction, insulation resistance is greater than 50,000 Megohms/μf at 25°C in an operating range between −55 to +125°C. Other characteristics include dielectric absorption 0.1%, retrace stability less than 0.2% and inherent noise unmeasurable. Dielectric strength characteristics are 200% over rated voltage at 25°C for 2

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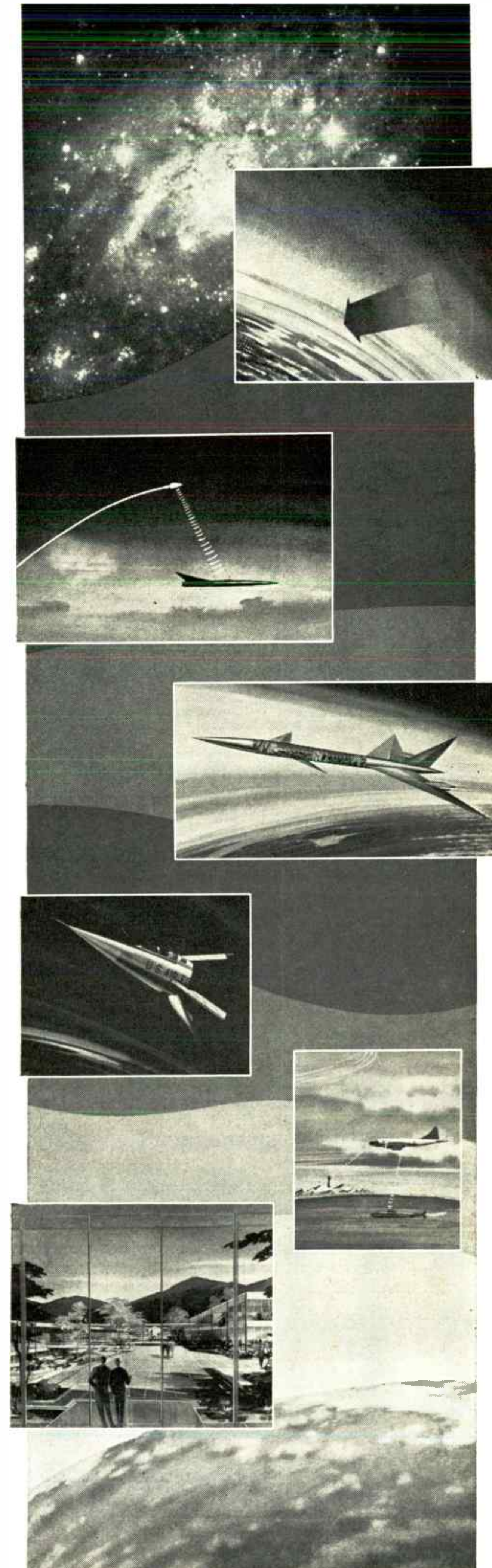

NEWS New Products

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

(Continued from page 154A)

minutes or a flash test of 250% over rated voltage.

The engineering design derating characteristics are zero volts at 85°C, 50% at 125°C for units under 200 volts, 25% for 400 volt units and 35% for 600 v units. Dissipation factor is less than 0.5 at 25°C at 100 cps.

Lightweight Recorder

Voltrec, a small compact lightweight recorder, has been designed by **Voltron Products, Inc.**, 1020 S. Arroyo Parkway, Pasadena, Calif., to monitor voltage or frequency with an accuracy of 0.5%. The Voltrec consists of two sections, a miniature inkless chart recorder and an expansion network. The recorder consists of a d'Arsonval meter with a free moving pointer.

A motor driven striker cam presses the pointer against sensitized paper once each two seconds, producing a record. Transducers for expanded scale recording of ac volts, dc volts, frequency and temperature are provided.

Delivery time is 30 days. Price of the dc recorder is $170; ac type $185; 60 cps frequency recorder $260; 400 cps frequency recorder $220; temperature recorder $190; chart paper, 6 rolls, $13.50.

Precision Capacitors

A new line of precision capacitors for precise circuitry has been developed by **Arco Electronics, Inc.**, 64 White St., New York 13, N. Y. to provide compact units of "standard" accuracy. Tolerance is ±0.1%. Long term operating stability is ±0.05%. These capacitors may be expected to maintain reliable operation throughout life of the equipment. High "Q" is also a principal feature of the line. All capacitors are hermetically sealed in bathtub type enclosures with compression glass seals. Calibration is made at 23°C at a frequency of 1 kc.

Type CPM capacitors have silvered mica dielectric. Standard voltage rating is 500 vdcw; but other voltage ratings, both higher and lower, are available to special order. Dissipation factor at 1 kc is less than 0.1%. Operating temperature range is −55°C to +125°C without derating. Temperature coefficient is 40 +15 ppm/°C.

Type CPP capacitors have polystyrene dielectric. Specifications hold for the entire temperature range of −55°C to +85°C. Dissipation factor is 0.02% at 1 kc. Dielectric absorption is 0.1% maximum. Tem-

(Continued on page 158A)

PHYSICISTS
ENGINEERS
MATHEMATICIANS

WHAT'S BEHIND THE NEW EXCITEMENT IN SCIENTIFIC AND ENGINEERING GROUPS AT STROMBERG-CARLSON

A Division of General Dynamics

New R & D programs in 34 critical electronic areas are behind it! Also:

...big increase in research budgets
...new experimental facilities
...significant technical advances are nearing application
...vigorous support given R & D personnel by engineer-oriented management

Every senior engineer and scientist who feels he can contribute to the expansion of man's capabilities in any of the areas below is invited to contact Stromberg-Carlson:

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- Single Sideband Communications
- Synchronous Data Transmission
- ASW Techniques
- Machine Tool Automation
- Radio Data Links
- High Intensity Sound Generators
- Air Acoustics
- Shaped Beam Display Systems
- High-Speed Automatic Missile Check-Out Equipment
- Super-Speed Read-Out and Printing Equipment
- Tacan Equipment
- Electro Acoustics & Transducers
- Logic Systems
- Sound Systems
- RF Equipment
- Precision Hi-Fi Components

RESEARCH

- Paramagnetic Resonance
- Thin Photoconductor Films
- Ferroelectricity
- Propagation and Coding
- Speech Analysis
- Bandwidth Compression
- Hydro-Acoustic Transducers
- Molecular Electronics
- Defect Solid State Physics
- Parametric Devices
- Tunnel Diode Logic
- Scatter Propagation Analysis
- Plasma Physics

We are particularly interested in people with advanced degrees. If you have a Physics, Electrical Engineering or Mathematics degree and experience in one or more of the above areas, you are invited to discuss the positions currently open. Please write details of your background and experience in complete confidence to Mr. Maurice Downey.

STROMBERG-CARLSON

A DIVISION OF **GENERAL DYNAMICS**

1476 N. Goodman St., Rochester 3, New York

NEWS New Products

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(Continued from page 156A)

perature coefficient is -120 ± 15 ppm/°C, making these units adaptable to matching with positive temperature coefficient precision resistors. Standard voltage rating is 400 vdcw.

For further information write to the firm.

SSB and RF Service Triode

United Electronics Co., 42 Spring St., Newark 4, N. J., a subsidiary of Ling-Altec, Inc., has announced a new triode, Type 572. This ruggedized tube is a one way replacement for the prototype 811A in most applications. A specially processed electronic graphite anode permits the 572 a 50% increase in plate dissipation over the 811A.

Rugged characteristics have been achieved by the utilization of a hard glass envelope, heavy-duty tungsten lead wires, a non-frangible filament and enclosed getter traps. The 572 withstands greater overloads, resists shock and vibration and provides longer life.

Applications for the 572 are found in communications, ground support equipment, sonar, ultrasonics and wherever reliability and ruggedness are required. This new tube is particularly useful in zero bias audio Class B service, and in single sideband RF service.

Temperature Coefficient Calibration Chamber

Conrad, Inc., subsidiary of Compton Manufacturing Co., 141 Jefferson St., Holland, Mich., announces availability of a production model temperature coefficient calibration chamber for thermistor and capacitor tests. The equipment capability is for stabilizing within ± 0.005°C stability. The equipment is also available with additional controls to permit stability of ± 0.001°C.

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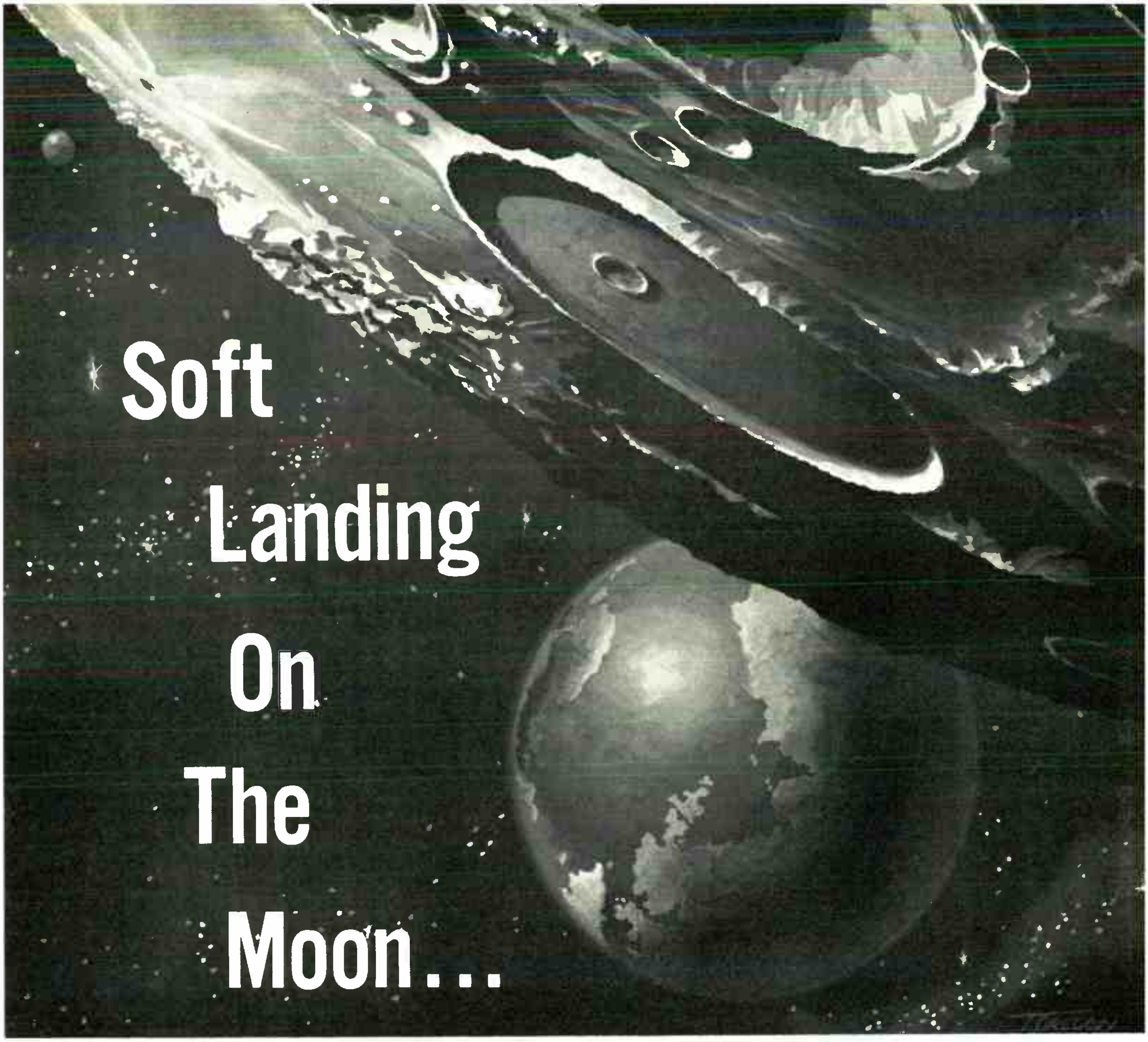

A CHALLENGE TO EXCITE ANY ELECTRONICS ENGINEER WORTH HIS SALT... THE KIND RYAN NEEDS RIGHT NOW

Who will engineer a missile navigation system to make a "soft landing" on the moon? This is the kind of problem in avionics we're tackling here at Ryan – the kind that gets future-thinking electronics engineers excited.

Already Ryan is the world leader in continuous wave doppler navigation systems. This is one of several important reasons why Ryan is now the largest electronics firm in San Diego – *and the fastest growing!* If you are an electronics engineer ambitious to help advance the art, as well as your own career, through vital "frontier work" – we want you *right now* here at Ryan Electronics.

Ryan Electronics employs over 2000 people and has over one-third of the company's $149-million backlog of business. Under the leadership of some of America's most prominent scientists and engineers, Ryan is probing beyond the known . . . seeking solutions to vital problems of space navigation.

Expanding facilities of Ryan Electronics at San Diego and Torrance in Southern California are among the most modern in the West. You enjoy living that's envied everywhere, plus facilities for advanced study. Send your resume or write for brochure today: Ryan Electronics, Dept. 4, 5650 Kearny Mesa Road, San Diego 11, Calif.

DIVISION OF RYAN AERONAUTICAL COMPANY
SAN DIEGO & TORRANCE · CALIFORNIA

RYAN ELECTRONICS

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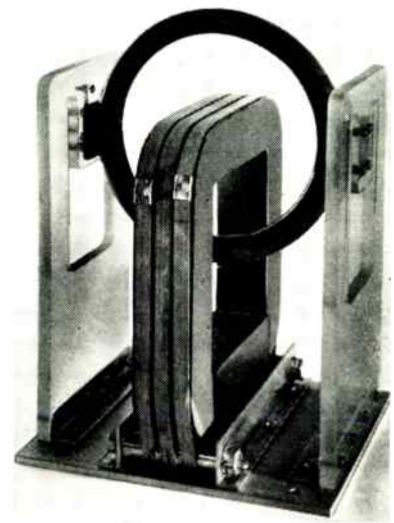
The temperature range available is from +250°C to −75°C. Instrumentation for recording the controlled condition can be supplied as well as automatic temperature conditioning.

High Voltage Transformer



Pearson Electronics, Inc., 707 Urban Lane, Palo Alto, Calif., has announced a new low capacitance, high voltage insulated filament transformer which is rated at 300 kv pulse immersed in oil, or 30 kv pulse, ac, or dc in air. A variety of output voltages and currents are available including the common heater voltages for high power klystrons, TWT's hydrogen thyratrons, high voltage diodes, magnetrons, and so forth.

Micropulse Delay Line

The Ralph M. Parsons Co., Electronics Div., 151 South De Lacey Ave., Pasadena, Calif., announces a new miniature lumped-constant delay line with a wide variety of general applications.

These high-density units are available in two standard configurations; cylindrical with a nine-pin header for standard tube

(Continued on page 162A)

NEWS New Products

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

(Continued from page 160A)

socket mounting, and rectangular for use with printed circuitry. The cylindrical model is 1.0 inch in diameter by 2.0 inches long, and weighs 1.5 to 2.0 ounces depending on required delay. Volume and weight of the rectangular models are comparable. Units are completely potted.

Units having delay times of 0.5, 1.0, 2.0, and 3.0 microseconds are available off-the-shelf. The total delay is accurate to ±3 per cent. The rise-time to delay-time ratio is 5 per cent with an attenuation of 1 db. Leakage resistance is more than 100 megohms between line and ground at 500 volts dc. Input impedances available are 300, 500, and 1000 ohms, ±10 per cent.

Many modifications of standard units are available for special applications, including other header arrangements, delay times, input impedances, or as many as six taps.

Rixon Shifts Engineering Heads

Myrick

Ray

Joseph C. Myrick, vice president of **Rixon Electronics, Inc.**, 2414 Reedie Dr., Silver Spring, Md., has been named to head a newly formed Research and Development Division. Homer A. Ray, Jr., who has been a staff engineer and assistant to C. J. Harrison, operations vice president, has been named as the new engineering director in his place, according to J. L. Hollis, Rixon president.

Myrick, one of the company founders, has always been a strong force behind new ideas at Rixon.

Ray has had twenty years of experience in a variety of electronic communications projects. Some of his most outstanding designs have been in the field of antennas, transmission networks and related devices.

High Voltage Power Supplies

A new line of regulated high voltage power supplies known as the TRHV series has been announced by **Del Electronics Corp.**, 521 Homestead Ave., Mount Vernon, N. Y. These supplies are all solid state. Referencing and regulating circuitry is accomplished at low voltage. The operating temperature range is $-20°C$ to $+55°C$. Units are completely self contained and require 115 volts excitation to be operative.

The TRHV line presently consists of 5 models with output voltage ratings of from 1 kv at 5 ma to 10 kv at 1 ma. They are rugged and feature good line, load and thermal stability. Other features include: regulation for 10% line change; 0.25% (input 115 volts $\pm 10\%$ 60 cps); regulation from no load to full load less than 0.25%; ripple is 0.5% rms at maximum current; temperature stability is better than 0.05%/°C; sealed construction; no warm-

(Continued on page 164A)

NEWS New Products

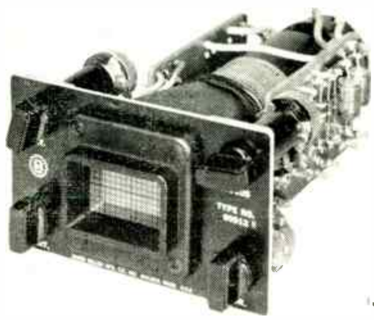
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 163A)

up time required; small size and light weight.

Additional data on the Del TRHV line of power supplies can be secured by writing to the firm.

Panel Scopes



A new series of compact Module Oscilloscopes using the 3UP1 $2\frac{1}{4}'' \times 1\frac{9}{16}''$ and the 3XP1 $3'' \times 1\frac{1}{2}''$ rectangular tubes has just been released by **The James Millen Mfg. Co., Inc.,** 150 Exchange St., Malden, Mass. The new 'scopes take up a panel space of $3'' \times 5''$ for the No. 90912-R and $3'' \times 6''$ for the No. 90913.

Transistorized Power Supply

A new line of tiny, lightweight, plug-in power supplies, actuated by "fleapower" sources as small as 1.5 volt penlight cell or a 1.3-volt mercury cell, and delivering output voltages as high as 20,000 volts, has been announced by **Victory Electronics, Inc.,** 50 Bond St., Westbury, N. Y. Each of the new units is about an inch in diameter and less than 3 inches long. These components are light and rugged.

Output currents range as high as 120 microamperes, depending on the model and the external circuit parameters. This power is capable of operating high-voltage instruments used in both terrestrial and in spaceborne, telemetered instrumentation and control, including: Geiger tubes, infrared detectors, ionization chambers, scintillation counters, maser frequency meters, cathode ray tubes and photomultipliers.

(Continued on page 166A)

NEWS New Products

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

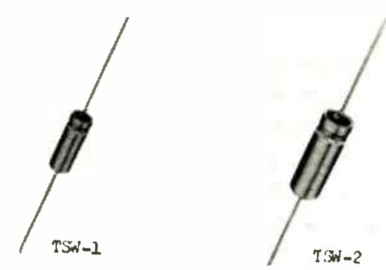
(Continued from page 164A)

Full encapsulation provides environmental protection; completely isolated input and output; silicon rectifiers throughout; 7-pin miniature plug-in base; and a MIL-approved potentiometer for precise adjustment of output.

Three models, currently available, cover ranges from 1.5 to 6.0 volts at 4 to 65 mils input; outputs from 800 to 20,000 volts dc in a wide selection of output-current values. The smallest in capacity and lowest in price ($93.80) will deliver 22 microamperes at 1200 volts from a 1.5-volt source, or 1500 volts at 7.5 from a six-volt source. Other units with 120 microampere output will deliver voltages as high as 1000 volts. A unit at 20,000 volts from a 1.5 input source in a miniature package is also available.

Wet Tantalum Capacitor

Believed to be the industry's first leakage and corrosion resistant wet tantalum capacitor, production of a 100-volt dc hermetically sealed wet electrolyte, sintered anode tantalum capacitor series is announced by **U. S. Semiconductor Products,** 3540 W. Osborn Rd., Phoenix, Arizona.



Immediate delivery of the new TSW line is available in ratings from 6 to 100 working volts and capacitance from 270 to 4.7 μf at 85°C. Applicable MIL Specs are met or exceeded in leads, moisture resistance, temperature and immersion. Supplied with or without insulating sleeve, the bare tube case is 0.188″ diameter, 0.525″ length for TSW1 model, and 0.282″ diameter, 0.720″ length for TSW2 model.

High capacitance derives from the large surface area of the sintered slug; unusual mechanical strength results from its internal weld and glass compression seal. Shelf life tests up to 2000 hours demonstrate no significant changes in performance characteristics.

For complete specifications, delivery dates and prices, write U. S. Semiconductor Products.

Marquardt to Systron-Donner Board

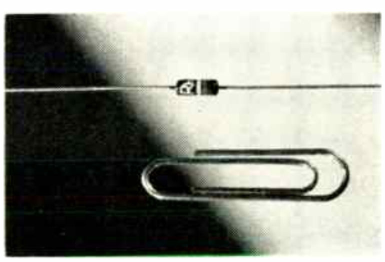
Roy E. Marquardt, president of Marquardt Corporation of Van Nuys, Calif., has been elected to the Board of Directors of **Systron-Donner Corp.**, 950 Galindo St., Concord, Calif. In 1944 Roy Marquardt founded the company which bears his name. He is a graduate of California Institute of Technology, where he received his bachelor's (1940) and Master of Science (1942) degrees in Aeronautical Engineering.

In 1942 he was appointed Engineer in Charge of Naval Research at Northrop Aircraft, Inc. As a result of research in methods of cooling engines mounted within the wings of an airplane, he rediscovered and expanded on the principles of the ramjet engine.

In order to pursue development of the ramjet, Mr. Marquardt accepted appointment as Director of Aeronautical Research at the University of Southern California. His company was organized at this time to provide the development and manufacturing requirements for a ramjet research program initiated by the Navy under USC sponsorship. The company currently employs approximately 4600 people at nine divisions in Van Nuys, Pomona, and Monrovia, California, and Ogden, Utah.

All-Purpose Silicon Diode

The 1N661A manufactured by **Rheem Semiconductor Corp.**, 350 Ellis St., Mountain View, Calif., is said to combine the low leakage specifications of the best "general purpose" types with the switching speed of the best "computer" types. This combination of the best features of two previously distinct diode families allows standardization.



The 1N661A also provides several orders of magnitude improvement in design safety margins for both switching time and reverse leakage. For example, the reverse leakage at +100°C is 30 times lower than any existing computer type. In addition, its recovery time is 25 times faster than the best general purpose type.

This new diode will serve as an upgraded replacement for a very large number of existing types. For example, it will meet the complete test specifications for General Purpose Types: 1N458, 1N459, 1N462, 1N463, and Computer Types: 1N625, 1N626, 1N627, 1N628, 1N629, 1N643, 1N659, 1N660, 1N661, 1N662, 1N778, 1N779, 1N789, 1N790, 1N793, 1N797, 1N801, 1N803, 1N806, and 1N807.

(Continued on page 168A)

NEWS New Products

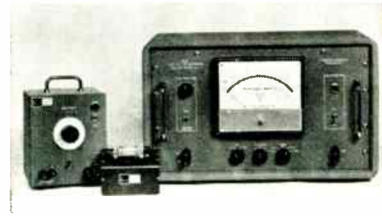
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your IRE affiliation.

(Continued from page 167A)

At 200 volts the maximum reverse leakage is 0.025 μa at $+150°C$. Recovery time is 0.3 microsecond to 400 kohms under the difficult switching conditions of $+30$ ma to -35 v in the JAN 256 circuit. This new diode is sealed in the standard glass package and certified to MIL-S-19500B. Price: $2.74 at 100 quantity level. Availability: Immediate from stock in production quantities.

Capacitor Tester

Direct reading of unknown capacitors in microfarads and their loss factor angle in degrees can be achieved by use of Type A101 Capacitor Tester and Type 405 Precision Phase Meter developed by **Ad-Yu Electronics Lab., Inc.,** 249 Terhune Ave., Passaic, N. J. This combination is very suitable for incoming inspection because the operating procedure is very simple, only one knob adjustment is needed before obtaining the answers directly in degrees and in microfarads. Loss factor angle is defined as the complementary angle of the power factor angle, namely, loss factor angle equals 90° minus power factor angle. In addition to the feature of simple operation in measuring unknown capacitors, both Type A101 and Type 405 are suitable for a number of other applications, such as measuring the Q of choke coils, phase shift of transformers, and other components in servo systems.



The specifications are as follows: The voltage across the unknown capacitor is a combination of 0.5 volt rms ac and 1.5 volt dc or less. The frequency of the ac voltage can be 60 cps, 120 cps, and 400 cps; other frequencies from 20 cps to 1 kc can be supplied with slight modifications. The accuracy of the capacitor measurement is $\pm 3\%$ for capacitors having loss factor angle less than 10°, more error for capacitors with larger loss angle. The loss factor measurement is $\pm 1°$ or $\pm 3\%$. There are three capacitance ranges: 0.1 to 15, 10 to 150 and 100 to 1500 μf.

Photoconductive Cells

A new, low resistance series of photoconductive cells hermetically sealed in glass, the "L" type, is now available from **Clairex Corp.,** 19 W. 26th St., New York 10, N. Y. Particularly useful in transistor and other low voltage applications, the

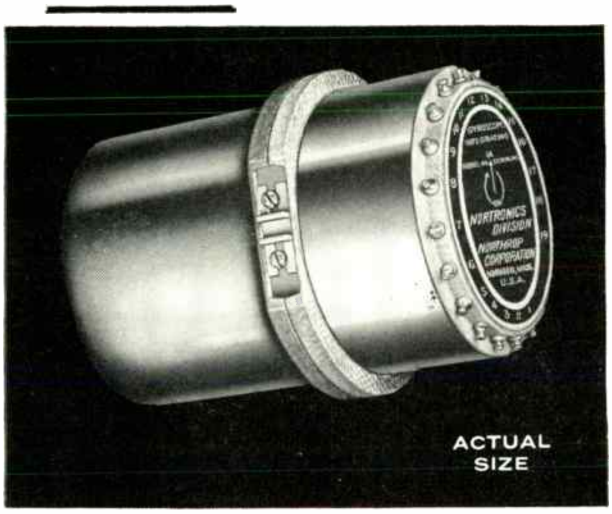
(Continued on page 170A)

AUTOPILOT OR ANTENNA PROBLEM...?

IF YOUR SPECS CALL FOR:

A miniature integrating gyro designed for economy, with superior off-null characteristics, low power consumption, and fast warm-up time . . .

THIS GYRO IS THE ANSWER



← GI-H5 miniature integrating gyro
Size ←
Cost ←
Perf ←

MEASURED CHARACTERISTICS **1.** Total Drift: less than 5° /hour (uncompensated), less than 4° /hour per G. **2.** Storage Temperature: −65°F to +200°F (higher temperature models available). **3.** Extremely stable thermally, with warm-up time to 5% of damping from −65°F within 7 minutes. **4.** Characteristic Time Constant for 6 degree memory angle unit = 0.87 milliseconds. **5.** P.M. Torquer with better than 0.05% linearity and 1 rad/sec peak steady-state torque. Unpack it, put it in the system – it's ready to go. Availability – 90 days. Write or TWX for detailed test data on the GI-H5 (Norwood 835-U, or field offices listed below).

NORTRONICS
A Division of
NORTHROP CORPORATION

PRECISION PRODUCTS DEPARTMENT
NORWOOD, MASSACHUSETTS
Field Offices

Highway #46
Teterboro, New Jersey
Telephone: ATlas 8-1750
TWX-Hasbrouck Heights 871-U

2486 Huntington Drive
San Marino, California
Telephone: ATlantic 7-0461
TWX-Alhambra 9619-U

NEWS New Products

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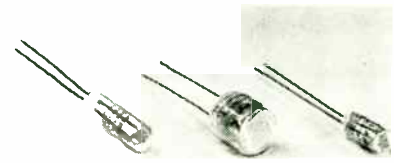
(Continued from page 168A)

units are rated at 60 volts maximum with a resistance as low as 40 ohms at 100 foot candles.

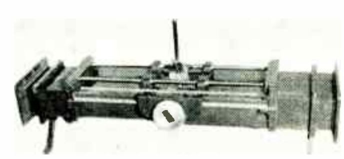


The standard "L" line consists of seven types, five of the Cadmium Selenide variety and two of Cadmium Sulphide. They provide a choice of resistance, speed and spectral response to suit most applications. The new units also provide a power dissipation range to ½ watt, continuous, and high ratios of light to dark current even at low light levels. They are available in the three package configurations of the 400, 500 and 600 series, ranging in diameter from 0.25 to 0.5 inch and in length from 0.5 to 1.0 inch. Prices range from $1.15 to $4.00 depending upon quantity.

All design parameters are stable and closely controlled. Literature is available direct from the manufacturer.

Slotted Line Impedance Meters

A new line of precision instruments to provide reliable and accurate VSWR measurement on the larger waveguides including the WR2300 and WR2100 types has been announced by **Schutter Microwave Corp.**, 80 E. Montauk Highway, Lindenhurst, N. Y.



The instruments are designed with adjustable mounts for ease of alignment, incorporate ball bearing carriages for smooth probe travel and are rugged in construction to insure long life operation.

These units are offered at comparatively lower cost and with prompt delivery. Additional information on those instruments such as graphical, technical data and cost schedules may be obtained directly from the company.

Crystal Filters

Systems Inc., 2400 Diversified Way, Orlando, Fla. engaged in electronics since 1956, introduces its newest family of products, precision crystal filters.

(Continued on page 172A)

NEWS New Products

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

(*Continued from page 170A*)

Typical of the filters offered is this new 10 mc band-pass filter No. BP-10000-40.

Developed for application in transistorized IF amplifiers, the BP-10000-40 is manufactured to meet applicable MIL specifications—operates from −55°C to 65°C in salt laden atmosphere. The hermetically sealed unit is capable of withstanding 50 G shock.

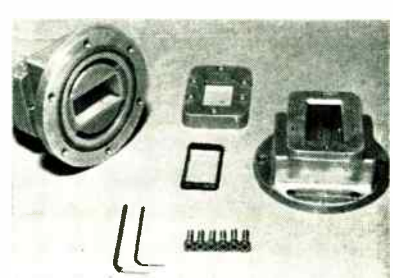
Specifications: Center frequency, 10 mc; 6 db bandwidth, 36 kc; Stopband rejection, 70 db; Ripple, 0.5 db; 60 db bandwidth, 80 kc maximum; Dimensions, ½"×1½"×2"; Tab mounting for printed circuit board applications.

Checkout Equipment Brochure

Packard Bell Electronics, 12333 W. Olympic Blvd., Los Angeles 64, Calif., has released a new brochure entitled "Automatic Electronic Checkout Equipment," which is available to all interested engineers. The design, development and production of this equipment is based on Packard Bell's experience on the Thor and Polaris missiles programs in addition to test and other checkout equipment. Brochure is available on request.

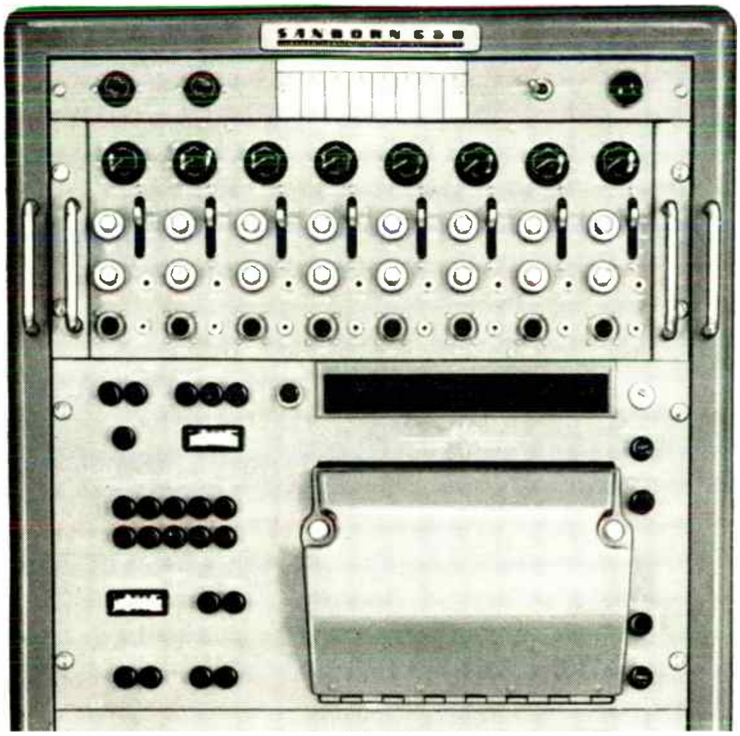
Solderless Waveguide Flanges

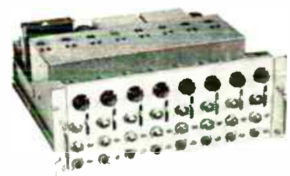
Microtech, Incorporated, Milldale Rd., Cheshire, Conn., a subsidiary of Talley Industries, Inc., announces the development of waveguide flanges which are attached to the waveguide without soldering or brazing.



According to George R. Houk, president of the firm, this new flange has elec-

(*Continued on page 174A*)





8-channel Amplifier

+

2000 cps Optical Recorder

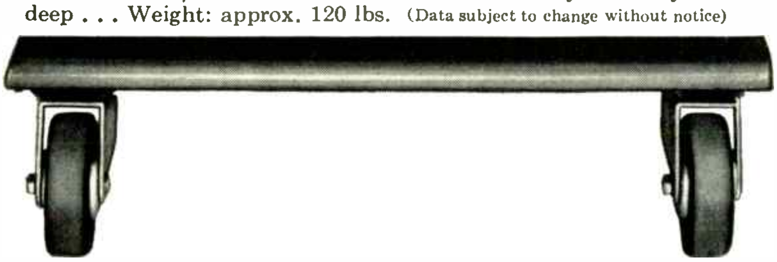
= a new 0 to 5000 cps direct writing system

SANBORN "650" SERIES

The Sanborn "650" is the first *complete* multi-channel high-speed optical recording system with medium gain general purpose amplification for each channel. Either the 8-channel amplifier or the recorder may be used separately. Together they provide a max. sensitivity of *2.5 mv/in* and a frequency response of *DC to 5000 cps* (within 3 db at 4 in peak-to-peak) in a multi-channel "direct writing" system.

MODEL 658-3400 GENERAL PURPOSE AMPLIFIER. Here is the first sensitive multi-channel amplifier designed specifically for use with high frequency optical galvanometers — those in the Sanborn "650" and any similar recorder. The single chassis has 8 separate channels, each one complete from floating and guarded signal input to galvanometer output. They include front-end modulator and input transformer, medium gain carrier amplifier, demodulator, filter and driver amplifier. An internal pre-emphasis circuit increases galvanometer frequency range from 2000 cps to 5000 cps in the "650" recorder. The all transistorized circuitry is mounted on easily serviced printed plug-in cards. The Amplifier chassis has an output transfer chassis on the rear which simplifies coupling to optical recorders of other manufacturers. External damping resistors are easily added when required.
Specifications: Sensitivity: 7.2 ma/mv input, max. . . . Attenuation: X2, 5, 10, 20, 50, 100, 200, 500, 1000, 2000 . . . Common Mode Performance: tolerance — 500 volts max; rejection — 140 db for DC . . . Input Resistance: 100,000 ohms all ranges floating and guarded.

MODEL 650 1- TO 24-CHANNEL OPTICAL RECORDER. The Model 650 Recorder provides high frequency direct writing recording, flexible housing and wide application possibilities. It may be used separately with from 1 to 24 plug-in type galvanometers of various natural frequencies. When used *with the 658-3400 Amplifier*, the recorder is equipped with eight 2000 cps galvanometers — extended to 5000 cps by the amplifier pre-emphasis circuit — for wide range, high speed, wide deflection recording. The recorder has nine electrically controlled (local or remote) chart speeds, beam interrupters for trace identification, timing lines at 0.01 or 0.1 sec intervals; amplitude lines with manual washout from ¼, ½, ¾ or all of the record; full chart width deflection for each trace and trace overlap.
Specifications: Input Sensitivity: 17.5 ma/inch (with 2000 cps galvanometers) . . . Chart Speeds: 0.25, 0.5, 1.0, 2.5, 5.0, 10, 25, 50 and 100 inches/second . . . Dimensions: 19" wide by 17½" by 16½" deep . . . Weight: approx. 120 lbs. (Data subject to change without notice)



Complete data is available from Sanborn Sales-Engineering Representatives located in principal cities throughout the United States, Canada and foreign countries.

SANBORN COMPANY
INDUSTRIAL DIVISION
175 Wyman Street Waltham 54, Mass.

At the WESCON SHOW — Booths 608-609

NEWS
New Products

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

(Continued from page 172A)

trical characteristics equal to its equivalent brazed flange. The flanges called "EASY-TACH," can be attached to waveguide by inexperienced personnel using Allen wrenches. In addition to being mechanically and electrically equivalent, the flange is also pressure tight. After assembly, the flange joint will stand greater than 230 lbs. straight pull.

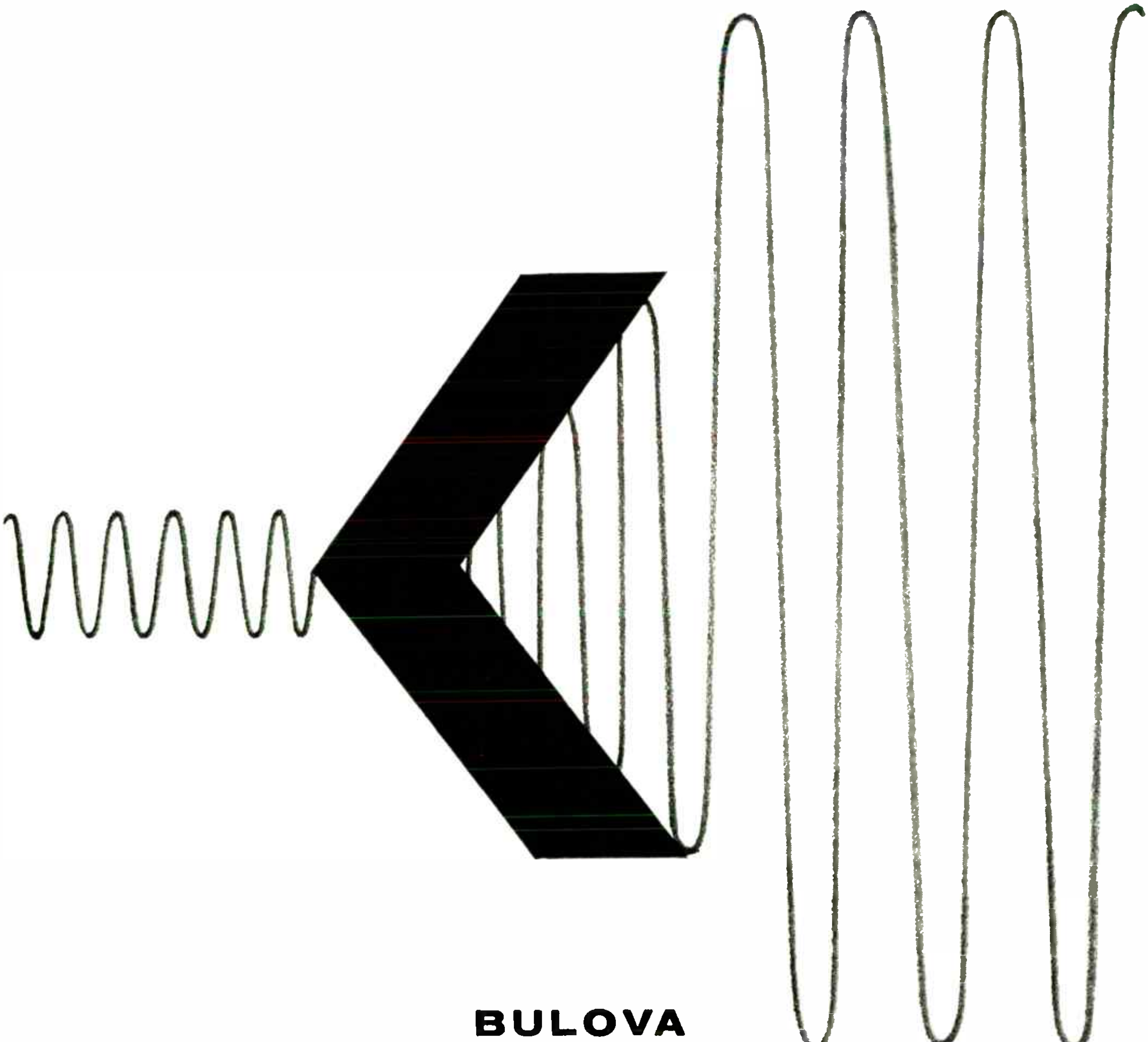
They are ideal for field installations since assemblies can be fabricated on the spot.

First models produced are ETP-137 plate and ETC-137 choke equivalent to UG-344/U and UG-343/U respectively. They will be available in either aluminum or brass.

Multiplier Phototube

Research and development by the **Electronic Tube Div., Allen B. DuMont Laboratories, Inc.,** 750 Bloomfield Ave., Clifton, N. J., has been carried out to produce eight-inch virtually flat end-window multiplier phototubes for low level radiation counting on the entire human body or other large surfaces. This type of tube also has extensive industrial applications

(Continued on page 176A)



NEWS New Products

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(Continued from page 174A)

where maximum light-gathering power is necessary. The flat end screen of large diameter permits excellent optical coupling to conventional large size scintillation crystals. Initial prototypes of the tube (K1979) were developed with technical cooperation from the contractor, the New York Operations Office of the Atomic Energy Commission under contract AT (30-1)-1336. The prototypes are undergoing evaluation.

DuMont's Tube Division officials emphasize that these prototypes are interim products in a continuing development program, and that further refinement

(Continued on page 178A)

NEWS New Products

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(Continued from page 176A)

of materials and other structural changes are expected in forthcoming versions.

The type K1979 as delivered under the contract is a 10-stage multiplier phototube with a photo cathode of S-11 spectral response. Its steel cone reduces radioactive background count and affords partial magnetic and electrostatic shielding. It features a focusing electrode (shield) for optimum photoelectron collection.

Tentative specifications are available and requests are invited dealing with modifications of the existing type. Write Electronic Tube Sales Dept., Allen B. DuMont Laboratories, Inc.

Stereo/Mono Recorder

A new combination stereophonic/monophonic recorder, the Model 354, was introduced by **Ampex Professional Products Co.**, Redwood City, Calif. The 354 will record and reproduce two track stereo, two track mono (for full track use), or half track mono; providing essentially two professional recorders in one—stereophonic and monophonic. The price of the Model 354, unmounted, is $1775.00, $300 above a monophonic Ampex recorder.

Features include smaller size, greater operating convenience, lighter weight and lower cost. Smaller size now permits a stereo-console model, a portable model with room for mixer or other accessories, or rack mount in smaller space. Greater operating convenience and control is made possible by front panel adjustments of bias, record equalization and playback level, eliminating need to remove electronics from console or case. New record selector switch, side-by-side positioning of VU meters, and record level reset indicators are among the other features giving the operator speedier, more complete control during recording sessions. The tape transport is the same used throughout the 350–351 series which has proven itself with more than 9000 units.

Accessories to the Model 354 include a four position stereo/mono mixer, remote control unit, plug-in equalizers, plug-in input units and plug-in microphone preamplifiers.

Transistor Curve Generator

Trans-Western Electronics, formerly Western Instruments, P. O. Box 1473, Ventura, Calif., has designed a transistor curve generator to generate a collector curve for low and medium power transistors. The Model 81 generates a single curve in the grounded-emitter configuration for display on an external oscilloscope. The unit supplies collector voltages continuously variable to 40 volts, base currents from 20 microamperes to 10 milliamperes and provides collector currents from 200 to 500 ma. The unit operates from a 117 volt ac source and requires no batteries.

Availability: 30 days. Price is $118.50.

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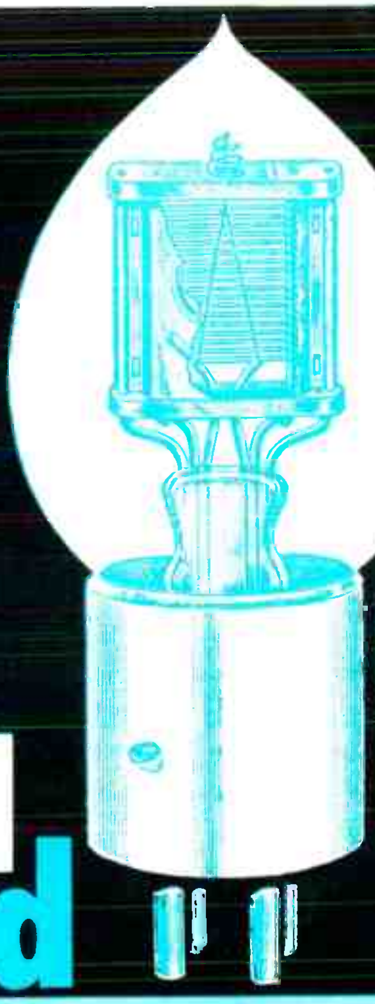


THEN and NOW

"Then," as it refers to the electronics industry, means . . . "just a very few years ago." The radically swift advances made in those few years have changed the whole concept of component design and performance.

Capacitor users will find the new capacitors which modern circuitry demands at Pyramid Electric Company, where miniaturization with highest reliability are among our proven engineering and production accomplishments. A few of the modern capacitor products are shown here.

AT WESCON—BOOTH 728

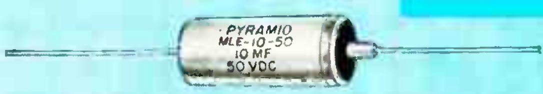
type TAD... solid tantalum high reliability capacitors. They are available in a wide range of sizes and values, are hermetically sealed, and have a long shelf and operating life.

temperature range: −80°C to +85°C
+125°C (with derating)

Write for bulletin ET-7

type TAK-H . . . wet electrolyte tantalum capacitors with built-in rugged construction, seep and vibration proof, and made to meet specifications MIL-C-3965.

Write for bulletin ET-8



type MLE . . . true miniature electrolytic capacitors. They are metal clad of all-welded construction, made to last ten years or longer.

temperature range: −40°C to +85°C

Write for bulletin EL-6

type 107...Mylar* dielectric capacitors–the smallest film capacitors made. Non-hydroscopic polyester dielectric and wrapper with thermosetting resin end seals.

temperature range: −55°C to +125°C

Write for bulletin MY-2

*Mylar—reg. duPont TM

For full details write or call: **Sales Department S-100**

PYRAMID ELECTRIC COMPANY

ORANGE STREET, DARLINGTON, SOUTH CAROLINA • In Canada: Wm. Cohen, Ltd., 8900 Tanguay Street, Montreal • Export: Morhan Exporting Company, 485 Broadway, New York 13, N. Y.

NEWS New Products

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

(Continued from page 178A)

Non-Overload Amplifier

A new low-cost N-371 linear non-overload amplifier with integral discriminator, has been designed primarily for use in routine scintillation or proportional counting systems by **Hamner Electronics Company, Inc.**, P. O. Box 531, Princeton, N. J.

The amplifier operates linearly over a full 100 volt output range so that a standard pulse height analyzer can also be driven. Overload protection is provided so that when combined with Hamner's N-601 or N-603 single channel analyzers, spectrum analysis work can be performed.

Amplifier specifications are as follows: Gain, 7000; Pulse Shaping, Double RC Clipping; Linearity, Better than 0.2%, 4–100 volts; Hum and Noise, Less than 0.5 volt peak to peak at maximum gain.

Discriminator specifications are: Range, 3–100 volts by 10 turn helipot; Linearity, 0.5%, 3–100 volts; Resolving Time, Better than 1 μs.

The N-371 is available from stock at a price of $490. For detailed specifications please contact the firm.

Constant Voltage Transformer

Freed Transformer Co., Inc., is now offering a 60 cps 500 va constant voltage transformer with an input of 95 to 130 volts and an output of 115 volts. This unit has a line regulation of $\pm 1\frac{1}{2}$%.

The transformer can be hermetically sealed for military applications. For further information on this and other constant voltage transformers, write to the firm.

(Continued on page 182A)

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A. D. Ehrenfried and associates
Technical product planning—product and market evaluation—market development programs—all by graduate engineers.
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NEWS New Products

(Continued from page 180A)

Daly Named V. P. of Magnecord

Hugh J. Daly has been named vice-president of Magnecord Sales for Midwestern Instruments. In his new position Daly will be responsible for the sales and marketing program for the entire Magnecord product line manufactured by **Midwestern Instruments**, 41st & Sheridan Rd., Tulsa 18, Oklahoma. Daly, formerly manager for Magnecord, Inc. of Chicago, came to Midwestern Instruments in 1957. Midwestern purchased the assets of Magnecord, Inc. in December 1956, and moved the operation to Tulsa. The Magnecord tape product line was placed into production along with Midwestern's other electronic products, and Daly moved to Tulsa to become General Sales Manager for the Magnecord portion of MI's products.

Daly, a native of Chicago, has a degree in business administration from the University of Kentucky. He served as an officer in the Army Signal Corps during World War II. His work concerned radar, radio and electronics in general.

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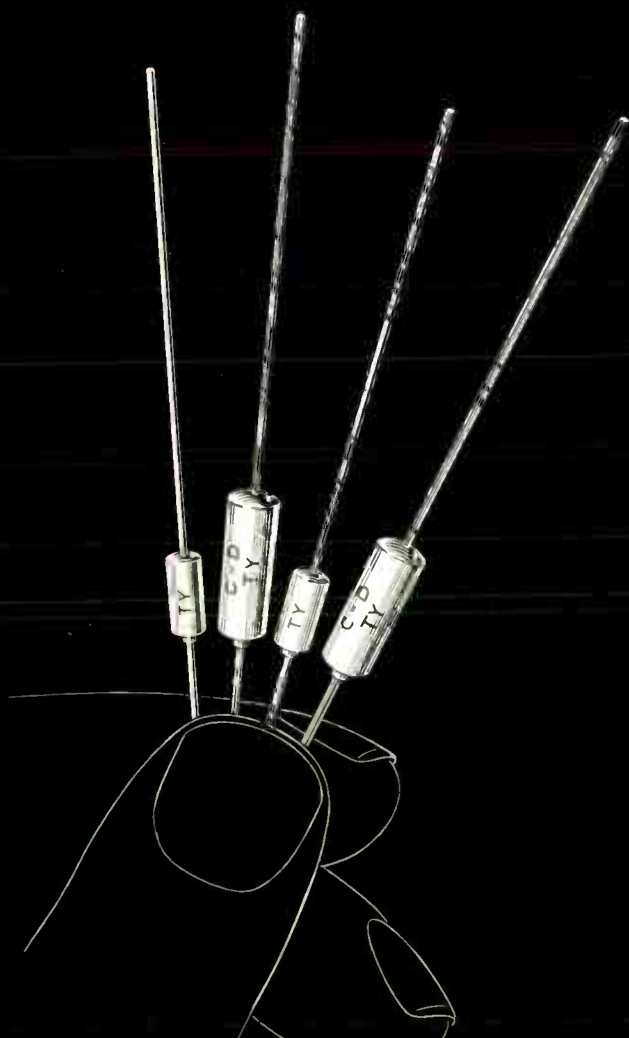

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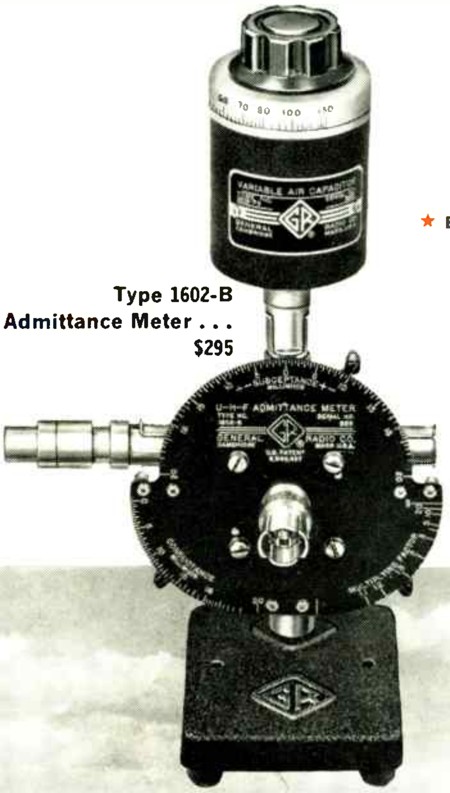
Its design is basic . . . three coaxial lines, one containing a conductance standard, one a susceptance standard, and one for connection to the unknown, are fed from a voltage source at a common junction point. Each of the lines contains an adjustable loop which samples the field within the line. In making measurements, these loops are adjusted for a null with the aid of an appropriate null detector. (G-R Type DNT Detector recommended.) At null, the settings of the conductance and susceptance loops times a multiplying factor established by a third loop gives the value of the unknown.



Type 1602-B Admittance Meter . . . $295

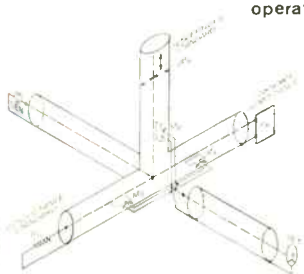
★ **WIDE FREQUENCY RANGE** . . . 20-1500 Mc; direct reading from 41-1500 Mc; useful for matching to 2000 Mc.

★ **DIRECT READING RANGES** that are independent of frequency
- **Conductance:** 0.2 to 1000 millimhos
- **Susceptance:** ±0.2 to ±1000 millimhos

With ¼-wavelength line between unknown and Meter, scales become direct reading in **Resistance** from 1 to 5000Ω, and **Reactance** from ±1 to ±5000Ω.

★ **EASY TO USE** . . . no sliding balances to chase . . . only three levers to adjust

★ **UNCOMPLICATED CONSTRUCTION** guarantees long, reliable operation and insures that basic ±3% accuracy will be held indefinitely.



★ **SMALL, LIGHTWEIGHT, PORTABLE** . . . ideal for antenna measurements

★ **A WIDE VARIETY OF ACCESSORIES** available to extend versatility:
- Balun for measurements on balanced lines and circuits.
- Component Mount for measuring circuit elements.
- Terminations for measuring reflection coefficient.
- Adaptors ranging from BNC to 3⅛-inch rigid line for measurements with any connector system.
- Oscillators and detector systems for complete frequency coverage.

A tribute to the Admittance Meter's versatility is its use at Grumman Aircraft, Bethpage, Long Island. Grumman engineers were faced with the problem of making accurate measurements on developmental aircraft antennas without influencing, by their physical presence, the antenna's radiation pattern or impedance characteristics. As a solution, they mounted an Admittance Meter, a G-R Unit Oscillator, and DNT Detector System inside an aircraft model. Pull cords connected to the Admittance Meter's controls were run out to a remote point where the operator could make his measurements without disturbing the setup. By adjusting the cords and using a surveyor's transit to read the instrument scales, accurate measurements could readily be made.

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