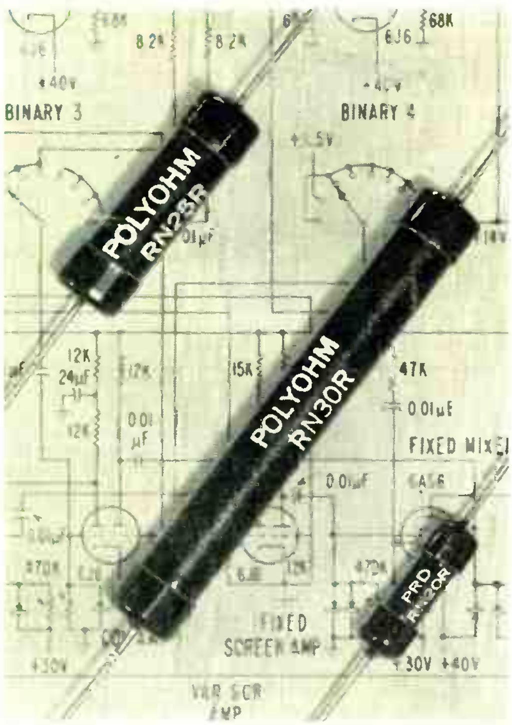
JULY · 1955

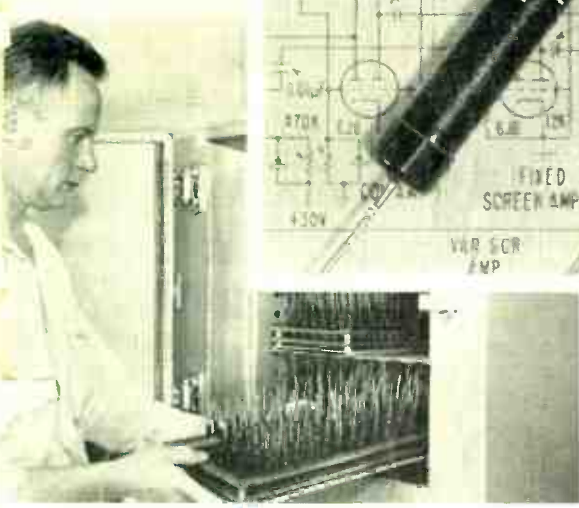
# Proceedings OF THE IRE

Ben R Cole

HIGH TEMPERATURE RESISTOR

New techniques in evaporating metal films on glass have led to high-temperature one-per-cent resistors which take full power at ambient temperatures up to 120 degrees C. The resistive element is protected by a high-temperature silicone varnish.



Polytechnic Research and Development Co.

Volume 43 Number 7

## IN THIS ISSUE



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The Institute of Radio Engineers

*Operate at temperatures to 125°C* without voltage derating

*Withstand dielectric test of twice rated voltage*

*Insulation resistance higher than* any other *metallized paper capacitor*

*Self healing dielectric*

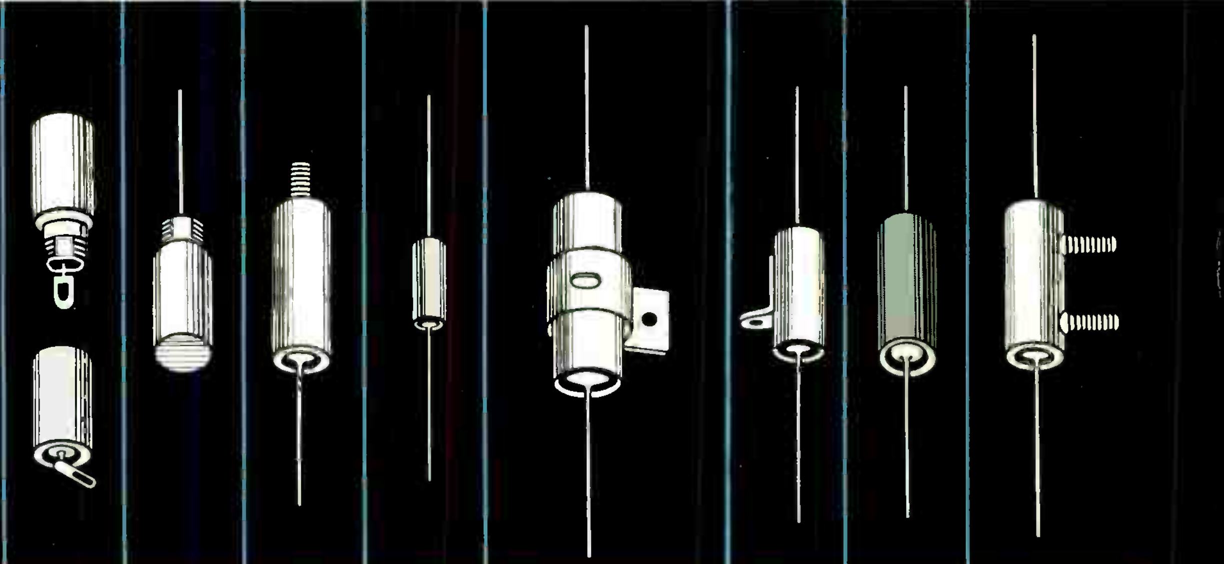
**Here are the finest capacitors which the present state of the art can produce.**

In the application of stringent quality controls, Sprague has gone so far as to metallize its own paper . . . the only commercial manufacturer to do this. Thus Sprague is the only capacitor manufacturer with *complete* control over the end product. And in no other type of capacitor does quality in manufacture play so important a part in performance.

*subminiature, metal-clad*

# SPRAGUE metallized paper capacitors

A complete range of ratings and sizes, hermetically sealed with glass-to-metal solder-seals in corrosion-resistant cases, is available in numerous mounting and terminal styles. Write for Engineering Bulletin 224 on your letterhead.



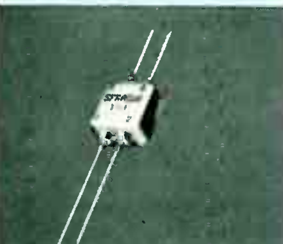
**SPRAGUE®** *world's largest capacitor manufacturer*

Sprague Electric Company | 235 Marshall Street | North Adams, Massachusetts

Export for the Americas: Sprague Electric International Ltd., North Adams, Massachusetts. CABLE: SPREXINT.

# choose from this complete line of

**Type 10Z**
**tubular pulse transformer**



**Type 15Z**
**miniature bathtub pulse transformer**

**Type 20Z**
**drawn-shell bathtub pulse transformer**

**Type 41Z**
**plug-in pulse transformer**

NOW YOU CAN CHOOSE from eighteen standard pulse transformers in four major construction styles, all in quantity production at Sprague. The standard transformers covered in the table below offer a complete range of characteristics for computer circuits, blocking oscillator circuits, memory array driving circuits, etc.

These hermetically sealed units will meet such stringent military specifications as MIL-T-27, and operate at temperatures up to 85°C. Special designs are available for high acceleration and high ambient temperature operation. In addition, the electrical counterparts of each transformer can be obtained in lower cost housings designed for typical commercial environment requirements.

Complete information on this high-reliability pulse transformer line is provided in Engineering Bulletin 502A, available on letterhead request to the Technical Literature Section, Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

## ELECTRICAL CHARACTERISTICS OF SPRAGUE PULSE TRANSFORMERS

| Type No. | Turns Ratio | Pulse Width μ seconds | Rise Time μ seconds | Primary Inductance | Leakage Inductance | Repetition Rate | Load and Output | Typical Applications |
|---|---|---|---|---|---|---|---|---|
| 10Z1 | 5:1 | 0.1 | 0.04 | 200 μH | 5 μH | 1 to 2 MC | 15 volts 100 ohms | Used in digital computer circuitry for impedance matching and inter-stage coupling. Pulses are of sine wave type. |
| 10Z2 | 4:1 | 0.07 | 0.03 | 200 μH | 20 μH | 1 to 2 MC | 20 volts 100 ohms | |
| 10Z3 | 1:1 | 0.07 | 0.03 | 125 μH | 12 μH | 1 to 2 MC | 20 volts 200 ohms | |
| 10Z4 | 3:1 | 0.07 | 0.03 | 160 μH | 15 μH | 1 to 2 MC | 20 volts 100 ohms | |
| 10Z6 | 4:1 | 0.1 | 0.04 | 200 μH | 6 μH | 1 to 2 MC | 17 volts 100 ohms | |
| 10Z12 | 1:1 | 0.25 | 0.02 | 200 μH | 2 μH | 12KC | 100 volts | Blocking Oscillator |
| 10Z13 | 1:1 | 0.33 | 0.07 | 240 μH | 2 μH | 2KC | 50 volts | Blocking Oscillator |
| 10Z14 | 7:1:1 | 0.50 | 0.05 | 1.2 mH | 20 μH | 1MC | 25 volts | Impedance Matching |
| 15Z1 | 3:1 | 5.0 | 0.04 | 7.5 mH | 22 μH | 10 KC | 10 volts 100 ohms | Impedance Matching and Pulse Inversion |
| 15Z2 | 2:1 | 0.5 | 0.07 | 6 mH | 15 μH | | 40 volts | Blocking Oscillator |
| 15Z3 | 5:1 | 10.0 | 0.04 | 12 mH | 70 μH | 10 KC | 10 volts | Impedance Matching |
| 15Z4 | 1:1.4 | 6.0 | 0.1 | 16 mH | 15 μH | 0.4 KC | 15 volts | Blocking Oscillator |
| 20Z1 | 5:5:1 Push-Pull | 1.5 | 0.25 | 4.0 mH | 0.3 MH | | 5 volts 10 ohms | Memory Core Current Driver |
| 20Z3 | 6:1 | 1 to 4 | 0.22 | 18 mH | 0.8 MH | 250 KC (max.) | 21 volts 200 ohms | Current Driver |
| 20Z4 | 6:1:1 | 1 to 7 | 0.25 | 55 mH | 0.3 MH | 50 KC (max.) | 22 volts 400 ohms | Current Driver and Pulse Inversion |
| 20Z5 | 3.3:3.3:1 Push-Pull | 2.4 | 0.2 | 2.8 mH | 0.2 MH | | 2.5 volts 6 ohms | Memory Core Current Driver |
| 20Z6 | 11:1 | 6.0 | 0.2 | 90 mH | 0.2 MH | 50 KC (max.) | 10 volts 75 ohms | Current Transformer |
| 41Z1 | 7:1:1 | 0.50 | 0.05 | 1.2 mH | 20 μH | 1 MC | 25 volts | Impedance Matching |

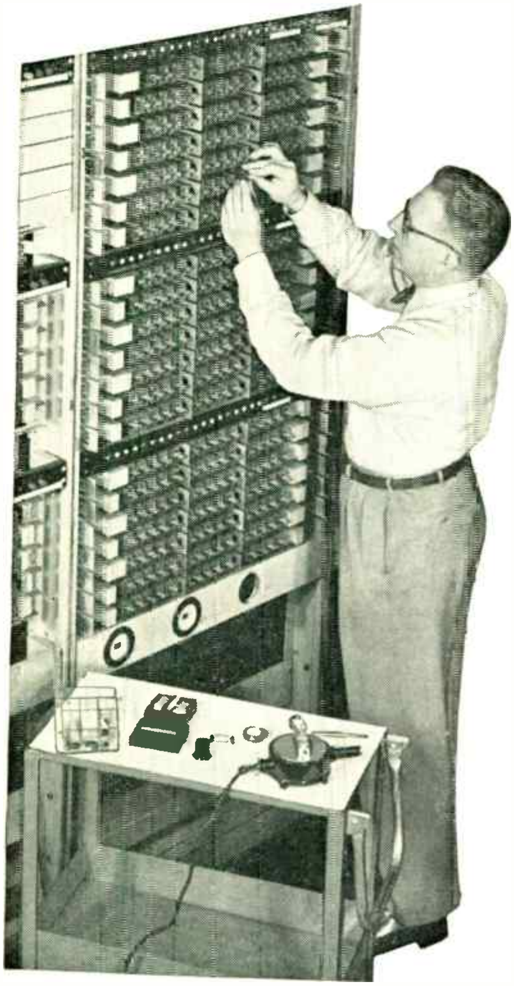
Sprague, on request, will provide you with complete application engineering service for optimum results in the use of pulse transformers.

**SPRAGUE®**

Export for the Americas: Sprague Electric International Ltd., North Adams, Mass. CABLE: SPREXINT

PROCEEDINGS OF THE I.R.E. July, 1955, Vol. 43, No. 7. Published monthly by the Institute of Radio Engineers, Inc., at 1 East 79 Street, New York 21, N.Y. Price per copy: members of the Institute of Radio Engineers, one additional copy, $1.00; non-members $2.25. Yearly subscription price: to members, one additional subscription, $9.00; to non-members in United States, Canada and U.S. Possessions $18.00; to non-members in foreign countries $19.00. Entered as second class matter, October 26, 1927, at the post office at Menasha, Wisconsin, under the act of March 3, 1879. Acceptance for mailing at a special rate of postage is provided for in the act of February 28, 1925, embodied in Paragraph 4, Section 412, P. L. and R., authorized October 26, 1927.

Table of Contents will be found following page 80A



# *He's "fingerprinting" a relay contact*

Bell Laboratories microchemists have perfected an ingenious new technique for "fingerprinting" relay contacts, the tiny switches on which a dial telephone system critically depends.

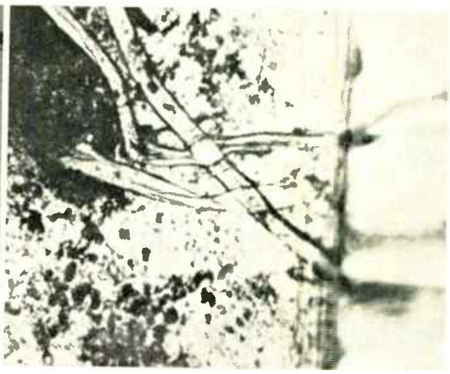
Using a portable test set, a chemist makes a plastic print of a contact. On-the-spot examination of the print with a microscope and chemical reagents quickly reveals the effects, if any, of arcing, friction, dust or corrosive vapors. While the chemist studies the print, urgently needed contacts continue in service. Findings point the way to improve relay performance.

This is another example of how Bell Telephone Laboratories research helps to keep your telephone system the world's best.

Above, Bell Laboratories microchemist applies plastic disc in heated clamp to relay contact. Imprint reveals contours of surface and picks up contaminants, if any. Part of portable test set is shown on table. Contacts, shown in small sketches, are of precious metal fused to base metal.

Preparing disc for microscopic examination. On-the-spot examination may reveal acid, alkali, sulfur, soot or other polluting agents peculiar to an area.

A microscopic look at disc often provides lead to nature of trouble. Unlike actual contact, print can be examined with transmitted light and high magnification.



Here the plastic disc has picked up microscopic lint that insulates contact, stops current. (Picture enlarged 200 times.) Traces of contaminants are identified in microgram quantities. Inert plastic resists test chemicals that would damage contact.

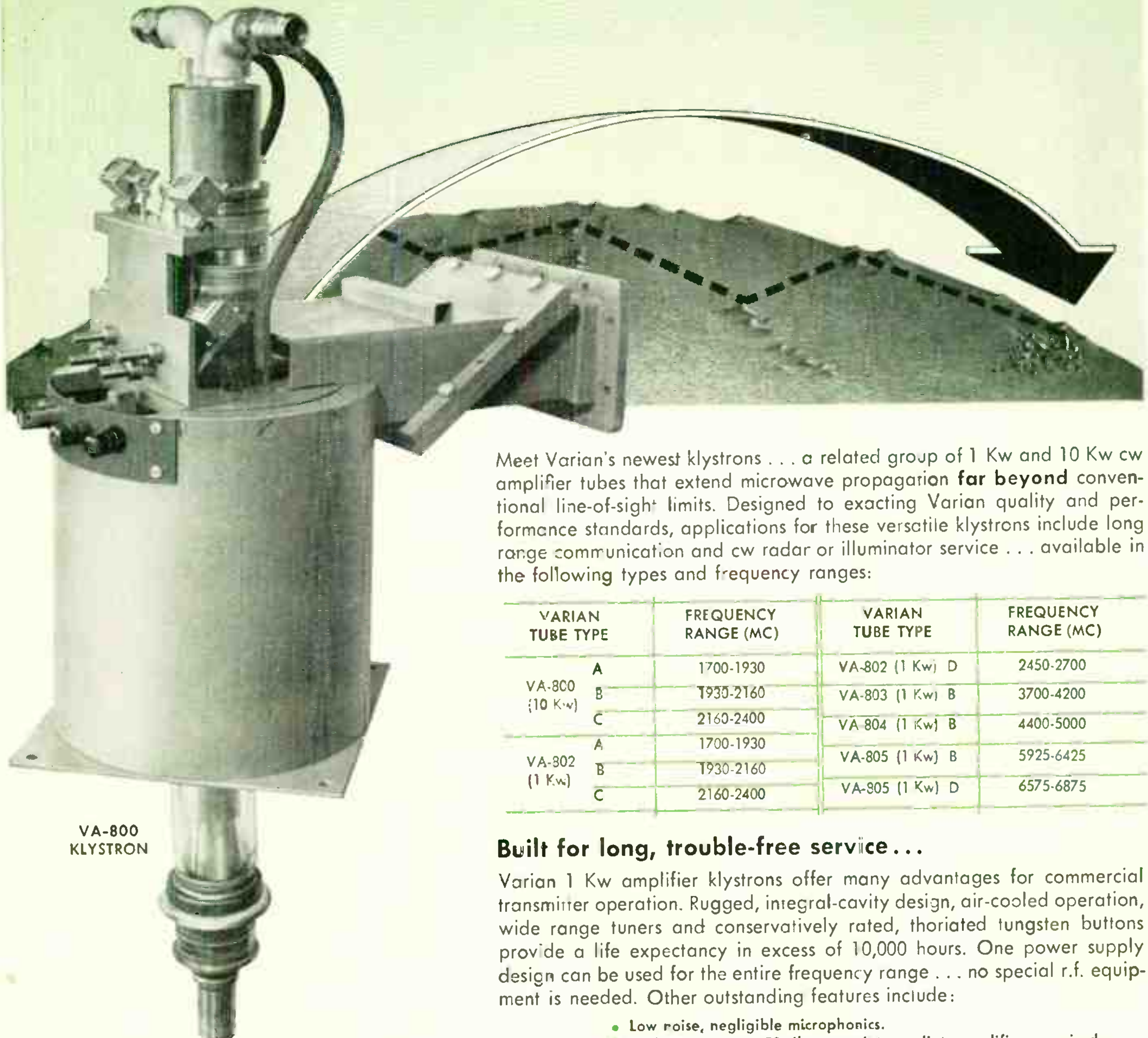
## Bell Telephone Laboratories

*Improving telephone service for America provides careers for creative men in scientific and technical fields*

***NEW*** **VARIAN KLYSTRONS**

# ADD SEVEN LEAGUE BOOTS

## to microwave transmission...



VA-800
KLYSTRON

Meet Varian's newest klystrons . . . a related group of 1 Kw and 10 Kw cw amplifier tubes that extend microwave propagation **far beyond** conventional line-of-sight limits. Designed to exacting Varian quality and performance standards, applications for these versatile klystrons include long range communication and cw radar or illuminator service . . . available in the following types and frequency ranges:

| VARIAN TUBE TYPE | FREQUENCY RANGE (MC) | VARIAN TUBE TYPE | FREQUENCY RANGE (MC) |
|---|---|---|---|
| VA-800 (10 Kw) A | 1700-1930 | VA-802 (1 Kw) D | 2450-2700 |
| VA-800 (10 Kw) B | 1930-2160 | VA-803 (1 Kw) B | 3700-4200 |
| VA-800 (10 Kw) C | 2160-2400 | VA-804 (1 Kw) B | 4400-5000 |
| VA-802 (1 Kw) A | 1700-1930 | VA-805 (1 Kw) B | 5925-6425 |
| VA-802 (1 Kw) B | 1930-2160 | VA-805 (1 Kw) D | 6575-6875 |
| VA-802 (1 Kw) C | 2160-2400 | | |

### Built for long, trouble-free service...

Varian 1 Kw amplifier klystrons offer many advantages for commercial transmitter operation. Rugged, integral-cavity design, air-cooled operation, wide range tuners and conservatively rated, thoriated tungsten buttons provide a life expectancy in excess of 10,000 hours. One power supply design can be used for the entire frequency range . . . no special r.f. equipment is needed. Other outstanding features include:

- Low noise, negligible microphonics.
- High gain — over 50 db . . . no intermediate amplifiers required.
- Standard waveguide output — permits direct coupling.
- High efficiency and simplicity of installation.

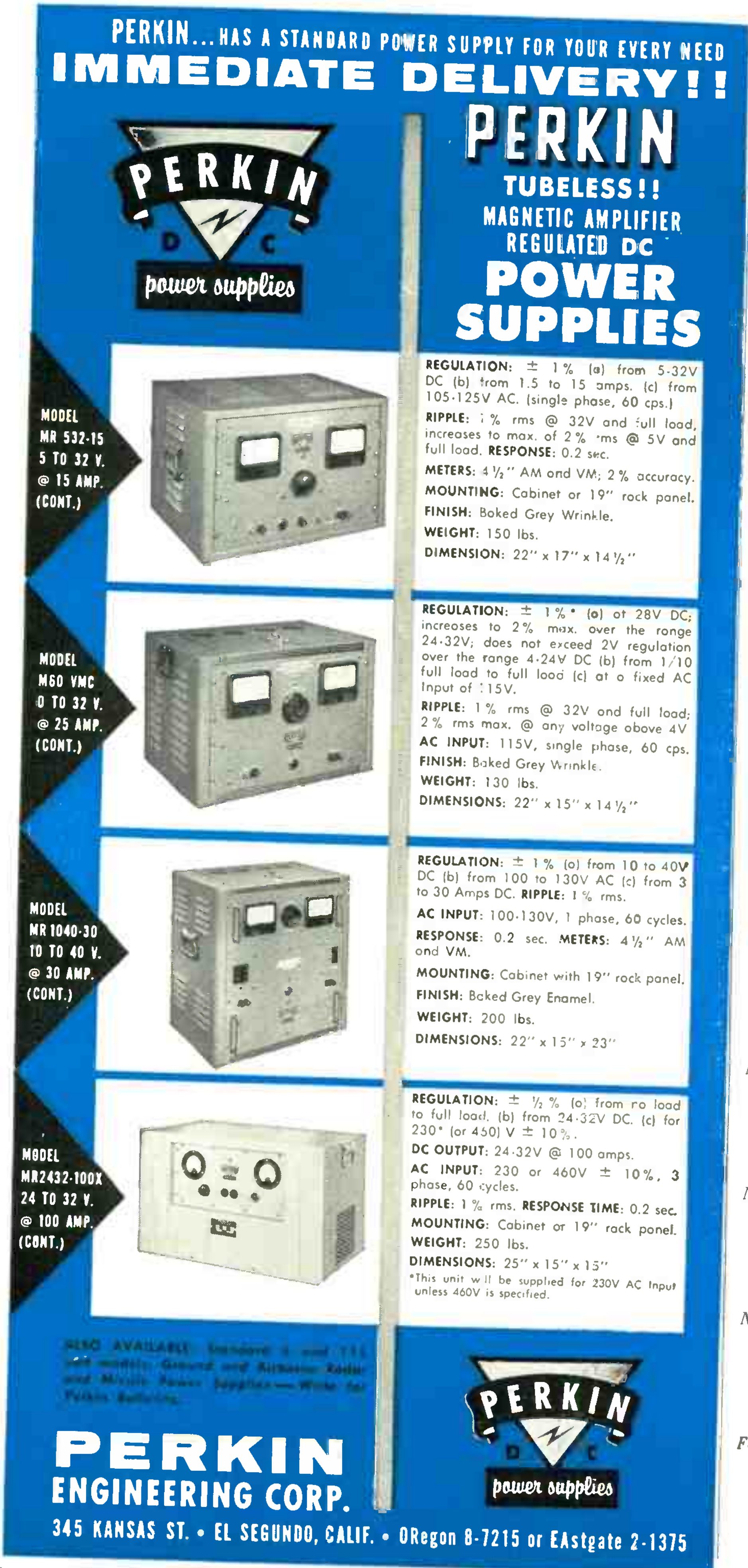
*EXTEND* ***YOUR*** *MICROWAVE HORIZONS . . . Write today for complete specifications and technical information on the new Varian 1 Kw and 10 Kw amplifier klystrons . . . data on the Varian V-42 and other* ***high power*** *klystrons is also available. Address our Applications Engineering Department or contact your nearest Varian representative*

THE MARK OF LEADERSHIP

**VARIAN associates**

PALO ALTO 2, CALIFORNIA

KLYSTRONS, TRAVELING WAVE TUBES, BACKWARD WAVE OSCILLATORS, R.F. SPECTROMETERS, MAGNETS, STALOS, UHF WATERLOADS, MICROWAVE SYSTEM COMPONENTS, RESEARCH AND DEVELOPMENT SERVICES




## Meetings with Exhibits

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

Δ

*Aug. 24-26, 1955*
**Western Electronic Show & Convention,** Civic Auditorium, San Francisco, Calif.
*Exhibits:* Mr. Mal Mobley, 344 N. La-Brea, Los Angeles 36, Calif.


*Sept. 12-16, 1955*
**Tenth Annual Instrument Conference & Exhibit,** Shrine Exposition Hall & Auditorium, Los Angeles, Calif.
*Exhibits:* Mr. Fred J. Tabery, 3442 So. Hill St., Los Angeles 7, Calif.

*Sept. 26-27, 1955*
**IRE Sixth Annual Meeting of the Professional Group on Vehicular Communications,** Hotel Multnomah, Portland, Ore.
*Exhibits:* Mr. Henry S. Broughall, General Electric Co., 2727 N.W. 29th Ave., Portland, Ore.

*October 3-5, 1955*
**National Electronics Conference,** Sherman Hotel, Chicago, Ill.
*Exhibits:* Mr. G. J. Argall, c/o DeVry Technical Institute, 4141 Belmont Ave., Chicago 41, Ill.

*Oct. 31-Nov. 1, 1955*
**IRE East Coast Conference on Aeronautical & Navigational Electronics,** Lord Baltimore Hotel, Baltimore, Md.
*Exhibits:* Mr. C. E. McClellan, Westinghouse Electric Corp., Air Arm Div., Friendship International Airport, Baltimore, Md.

*Nov. 3-4, 1955*
**Annual Electronics Conference, Kansas City Section,** Town House Hotel, Kansas City, Kans.
*Exhibits:* Mr. Charles V. Miller, Bendix Aviation Corp., P.O. Box 1159, Kansas City 41, Mo.

*Nov. 7-9, 1955*
**Eastern Joint Computer Conference (IRE-AIEE-ACM),** Hotel Statler, Boston, Mass.
*Exhibits:* Mr. J. D. Porter, Digital Computer Lab., Barta Building, M.I.T., Cambridge, Mass.

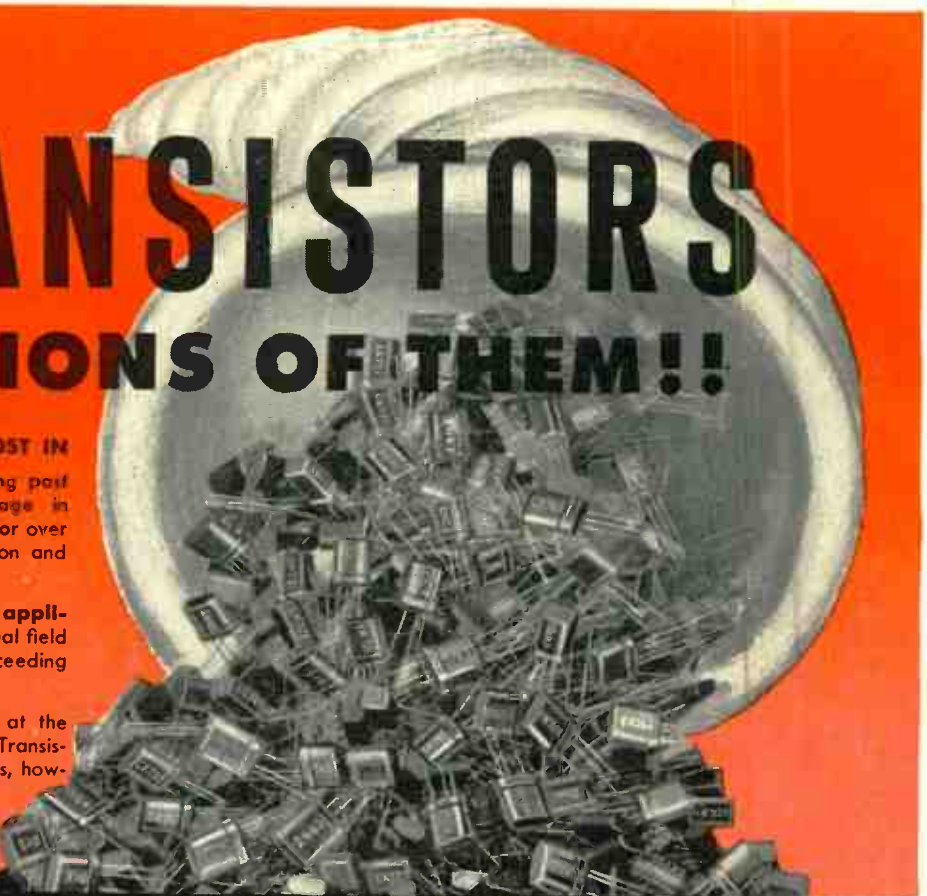
*Nov. 28-30, 1955*
**Instrumentation Conference & Exhibit,** Atlanta Biltmore Hotel, Atlanta, Ga.
*Exhibits:* Mr. W. B. Wrigley, Engineering Experiment Station, Georgia Institute of Technology, Atlanta, Ga.

*Feb. 9-11, 1956*
**Eighth Annual Southwestern IRE Conference and Electronics Show,** Municipal Auditorium, Oklahoma City, Okla.
*Exhibits:* Mr. Charles E. Harp, P.O. Box 764, Oklahoma City, Okla.

RAYTHEON

# TRANSISTORS

## MILLIONS OF THEM!!



**RAYTHEON IS FIRST AND FOREMOST IN**

— **mass production.** Raytheon is long past the experiment and development stage in Germanium PNP Junction Transistors — for over 2 years has had the quantity production and quality control techniques and resources

— **proved reliability in commercial application,** based on billions of hours of actual field performance and a record of success exceeding that of many reliable vacuum tubes

— **range of characteristics.** Look at the chart. You'll find one or more Raytheon Transistors that meet your specific requirements, however exacting.

**LOW FREQUENCY TRANSISTORS PLASTIC CASE**

| TYPE | Collector Volts | Collector Meg. ohms | Collector Cutoff μA | Emitter mA | Base ohms | Base Current Ampl. Factor | Max. Noise Factor db | Alpha Freq. Cutoff mc. | Max. Junction Temp. °C | Temp. Rise °C/mW |
|---|---|---|---|---|---|---|---|---|---|---|
| CK721 | —6 | 2.0 | 6 | —1.0 | 700 | 45 | 22 | 0.8 | 70 | 0.25 |
| CK722 | —6 | 2.0 | 6 | —1.0 | 350 | 22 | 25 | 0.6 | 70 | 0.25 |
| CK725 | —6 | 2.0 | 6 | —1.0 | 1500 | 90 | 20 | 1.2 | 70 | 0.25 |
| CK727 | —1.5 | 1.0 | 6 | —0.5 | 700 | 45 | 12 | 0.8 | 70 | 0.25 |

**LOW FREQUENCY TRANSISTORS — HERMETICALLY SEALED CASE**

| TYPE | Collector Volts | Collector Meg. ohms | Collector Cutoff μA | Emitter mA | Base ohms | Base Current Ampl. Factor | Max. Noise Factor db | Alpha Freq. Cutoff mc. | Max. Junction Temp. °C | Temp. Rise °C/mW |
|---|---|---|---|---|---|---|---|---|---|---|
| 2N63 | —6 | 2.0 | 6 | —1.0 | 350 | 22 | 25 | 0.6 | 85 | 0.58 |
| 2N64 | —6 | 2.0 | 6 | —1.0 | 700 | 45 | 22 | 0.8 | 85 | 0.58 |
| 2N65 | —6 | 2.0 | 6 | —1.0 | 1500 | 90 | 20 | 1.2 | 85 | 0.58 |
| 2N106 | —1.5 | 1.0 | 6 | —0.5 | 700 | 45 | 12 | 0.8 | 85 | 0.58 |

**HIGH FREQUENCY TRANSISTORS — HERMETICALLY SEALED CASE**

| TYPE | Collector Volts | Collector Cutoff μA | Emitter mA | Extrin. Base Resis. ohms | Base Current Ampl. Factor | Alpha Freq. Cutoff mc. | Max. Junc. Temp. °C | Temp. Rise °C/mW | Coll. Capac. μμf | Gain at 455kc db | Gain at 2 mc db | Rise time* μsecs | Decay time* μsecs |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 2N112 (CK760) | —6 | 1 | —1.0 | 75 | 40 | 5 | 85 | 0.62 | 14 | 32 | 18 | 0.05 | 0.06 |
| 2N113 (CK761) | —6 | 1 | —1.0 | 75 | 45 | 10 | 85 | 0.62 | 14 | 33 | 20 | 0.04 | 0.05 |

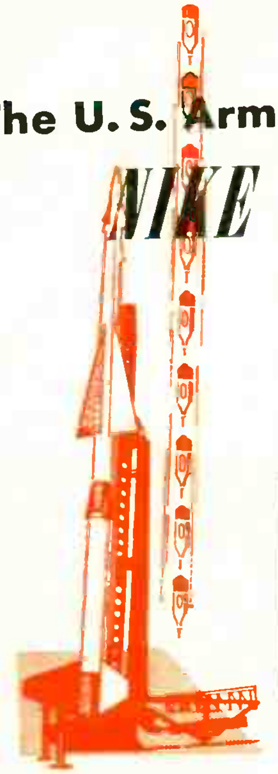
*measured in circuit which will be supplied on request

Note: above characteristics are average except where noted

*There are more — several times more*

**RAYTHEON TRANSISTORS**

*in use than all other makes combined*



# The U. S. Army Ordnance *NIKE* thinks itself to a kill with "Brains" that rely on . . . RAYTHEON Tubes

Nike, as graceful as the Greek goddess for which she is named, locates, pursues and destroys hostile aircraft. Nike reaches far beyond conventional anti-aircraft weapons; outmaneuvers fighters or bombers alike — actually thinks her way to the kill

**From bottom to top:**

Nike blasts off

Nike reaches full flight speed in seconds

Unerringly, the Nike system's electronic "brain" takes her to the target.

Selection of Western Electric Company as prime contractor for the U. S. Army's Nike guided missile systems was logically based on the necessity for supreme reliability of manufacture and of consequent performance.

Selection of Raytheon Subminiature Tubes by Western Electric was dictated by that same necessity. A number of the subminiature tubes that go into the Nike system's superhuman "brain" are Raytheon Tubes.

No pains were spared, no tests overlooked in securing the very finest, most dependable tubes for the Army's *Nike.* Think, then, of your own tube applications and their needs whether they be for low microphonics, low power, long life, extreme reliability under severe service conditions or a combination of requirements. Will you be satisfied with anything less than the best? Specify Raytheon Quality Subminiature Tubes.

## RAYTHEON Flat Press Subminiature Tubes ...the tubes with the SEAL of RELIABILITY

The long, flat press glass to metal seal is a Raytheon development that reduces glass strain, button cracking, lead burning, lead corrosion and lead breakage. Its in-line leads permit easier socketing and easier wiring. It is ideal for printed circuitry.

RAYTHEON®

*Excellence in Electronics*

**RAYTHEON MANUFACTURING COMPANY**

Receiving Tube Division — Home Office: 55 Chapel St., Newton 58, Mass., Bigelow 4-7500

For application information write or call the Home Office or: 9501 Grand Avenue, Franklin Park (Chicago), Illinois, TUxedo 9-5400

589 Fifth Avenue, New York 17, New York, Plaza 9-3900 • 622 South La Brea Ave., Los Angeles 36, California, WEbster 8-2851

RAYTHEON MAKES ALL THESE:

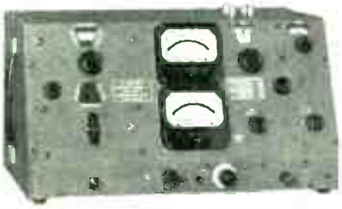
RELIABLE SUBMINIATURE AND MINIATURE TUBES · SEMICONDUCTOR DIODES AND TRANSISTORS · NUCLEONIC TUBES · MICROWAVE TUBES · RECEIVING AND PICTURE TUBES

# BRC

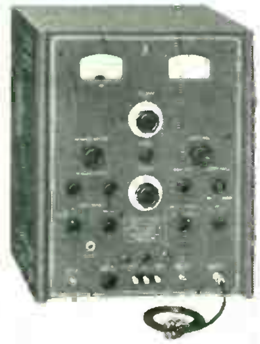
## PRECISION ELECTRONIC MEASURING INSTRUMENTS

**RX METER:** Wide Frequency Band RF Bridge
**Q METERS:** Low, Medium, High and Very High Frequencies
**UNIVERTERS:** Low, Medium and High Frequency Converters
**SIGNAL GENERATORS:** Frequency and Amplitude Modulated For Aircraft Navigation, Mobile and TV Receivers. Precision broad band sweeps with markers
**SIGNAL GENERATOR CALIBRATORS:** RF Voltage Standard in the low microvolt range over a wide frequency range


Q METER TYPE 190-A


Q METER TYPE 260-A


FM-AM SIGNAL GENERATOR TYPE 202-B


SWEEP SIGNAL GENERATOR TYPE 240-A


RX METER TYPE 250-A


RF VOLTAGE STANDARD TYPE 245-A

**Q METERS**

| Type | Freq. Range | Q Range | Tuning Capacity Range | Q Accuracy | Price |
|---|---|---|---|---|---|
| 260-A | 50 kc to 50 mc | 10 to 625 | 30-450 mmf | 5% to 30 mc | $725. |
| 190-A | 20 mc to 260 mc | 5 to 1200 | 7.5 to 100 mmf | 7% to 100 mc | $625. |

**FM-AM SIGNAL GENERATORS**

| Type | Freq. Range | Output Range | Modulation FM | Modulation AM | Application | Price |
|---|---|---|---|---|---|---|
| 202-B | 54-216 mc | 0.1 to 200,000 μv | 0-240 kc | 0-50% | General | $975. |
| 202-C | 54-216 mc | 0.1 to 200,000 μv | 0-240 kc | 0-50% | Mobile | $1090. |
| 202-D | 175-250 mc | 0.1 to 200,000 μv | 0-240 kc | 0-100% | Telemetering | $980. |

**SWEEP SIGNAL GENERATOR**

| Type | Freq. Range | Output Range | Modulation FM | Modulation AM | Markers | Price |
|---|---|---|---|---|---|---|
| 240-A | 4.5 to 120 mc | 1.0 to 300,000 μv | ±1% to ±30% Center Freq. | 30% | Crystal & Pip. | $1375. |

**OMNI-RANGE SIGNAL GENERATOR (Crystal Monitored)**

| Type | Freq. Range | Output Range | Modulation | Application | Price |
|---|---|---|---|---|---|
| 211-A | 88-140 mc | 0.1 to 200,000 μv | 0-100% am | Omni-Range Rcvrs. | $1800. |

**GLIDE SLOPE SIGNAL GENERATOR**

| Type | Freq. Range | Output Range | Modulation | Application | Price |
|---|---|---|---|---|---|
| 232-A | 329-335 mc | 1.0 to 200,000 μv | 0-100% am | Glide Slope Rcvrs. | $1500. |

**WIDE BAND IMPEDANCE MEASURING EQUIPMENT—RX Meter**

| Type | Freq. Range | R Range | C Range | L Range | Price |
|---|---|---|---|---|---|
| 250-A | 0.5 to 250 mc | 15 to 100,000 ohms | 0-20 μμf | 0.001 μh-100 mh | $1250. |

**RF VOLTAGE STANDARD**

| Type | Freq. Range | Calibrated Out. | Output Impedance | Application | Price |
|---|---|---|---|---|---|
| 245-A | 0.1 to 1,000 mc | 0.5,1.0,2.0 mv | 50 ohms | Calibrates Signal Generators | $315. |

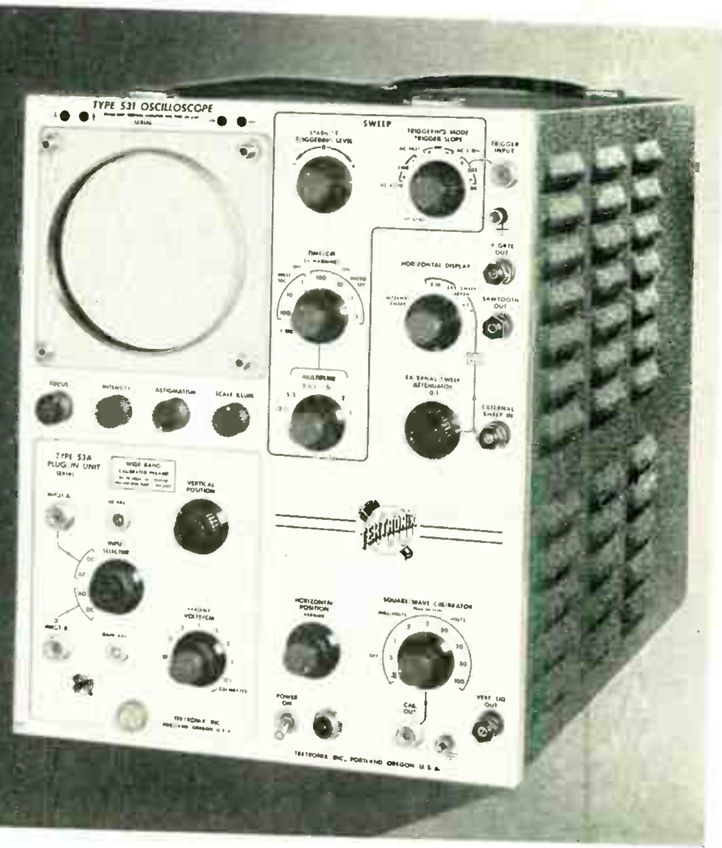
**UNIVERTERS**

| Type | Freq. Range | Output Range | Modulation FM | Modulation AM | Accessory to | Price |
|---|---|---|---|---|---|---|
| 207-A | 0.1 to 55 mc | 0.1 to 100,000 μv | 0-240 kc | 0-50% | 202-Band 202-C | $345. |
| 207-B | 0.1 to 55 mc | 0.1 to 100,000 μv | 0-240 kc | 0-50% | 202-D | $345. |
| 203-B | 0.1 to 25 mc | 1.0 to 100,000 μv | 1.5 to 30 mc | 30% | 240 A | $345. |

**BOONTON RADIO Corporation**
BOONTON·N·J·U·S·A·


ON LINE
ON TIME
MEMO
From: Engineering & Production
To: Sales Department
Subject: Size 8 and 22 Synchros
As promised last November, our size 8 and 22 standard synchros are now in production and available for 3 weeks or less delivery.
In addition, we have added the BuOrd type size 15 transmitter to the line.
Size 15
1.437" diameter
ACTUAL SIZE
RECEIVER
cppc
RESOLVER
cppc
Size 11
1.062" diameter
ACTUAL SIZE
CONTROL TRANSFORMER
cppc
Size 10
.937" diameter
ACTUAL SIZE
BuOrd Size 15
1.437" diameter
ACTUAL SIZE
BONUS
MK.22 MOD 1
CLIFTON PRECISION PRODUCTS CO. INC.
Next
HERE!
TRANSMITTER
TYPE CG-22-A-11
cppc
Size 22
2.161" diameter
ACTUAL SIZE
Next
HERE!
Size 8
.750" diameter
ACTUAL SIZE
Look to cppc for Synchro Progress
cppc
CLIFTON PRECISION PRODUCTS CO., INC.
CLIFTON HEIGHTS
PENNSYLVANIA

# 5 BIG REASONS why you should own this oscilloscope




The Tektronix Type 531 Oscilloscope is far ahead in performance characteristics, and is capable of a much wider range of applications than the ordinary general-purpose laboratory oscilloscope.

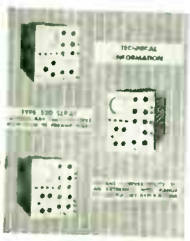
**1 THE TYPE 531 EXCELS** ***in vertical-amplifier characteristics***—with the Type 53B Plug-in Preamplifier it offers accurately calibrated sensitivity to 0.05 v/cm from dc to 10 mc, 0.035-μsec risetime... to 0.005 v/cm from 5 cycles to 9 mc, 0.04-μsec risetime.

**2 THE TYPE 531 EXCELS** ***in sweep characteristics***—Miller-runup circuitry generates linear sweeps in the extremely wide range of 0.02 μsec/cm to 12 sec/cm (600,000,000-to-1 ratio), with 24 accurately calibrated sweeps from 0.1 μsec/cm to 5 sec/cm. 5x magnifier is accurate on all ranges.

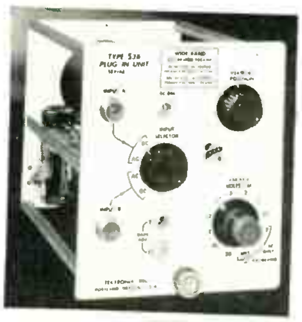
**3 THE TYPE 531 EXCELS** ***in triggering facilities***—offering amplitude-level selection, automatic triggering, and 30-mc sync in addition to all standard triggering modes.

**4 THE TYPE 531 EXCELS** ***in writing characteristics***—new Tektronix precision metallized crt with 10-kv accelerating potential provides high brightness, improved focus, and excellent linearity. (Recorded writing rate exceeds 175 cm/μsec).

**5 THE TYPE 531 EXCELS** ***in versatility***—Quick change plug-in preamplifiers and inherent oscilloscope capabilities combine to convert the Type 531 to applications normally requiring separate highly-specialized instruments. Available plug-in units provide for dual-trace... low-level differential... wide-band differential...and micro-sensitive applications in addition to wide-band high-gain applications. Current development work promises greatly-extended capabilities through new designs in plug-in units.



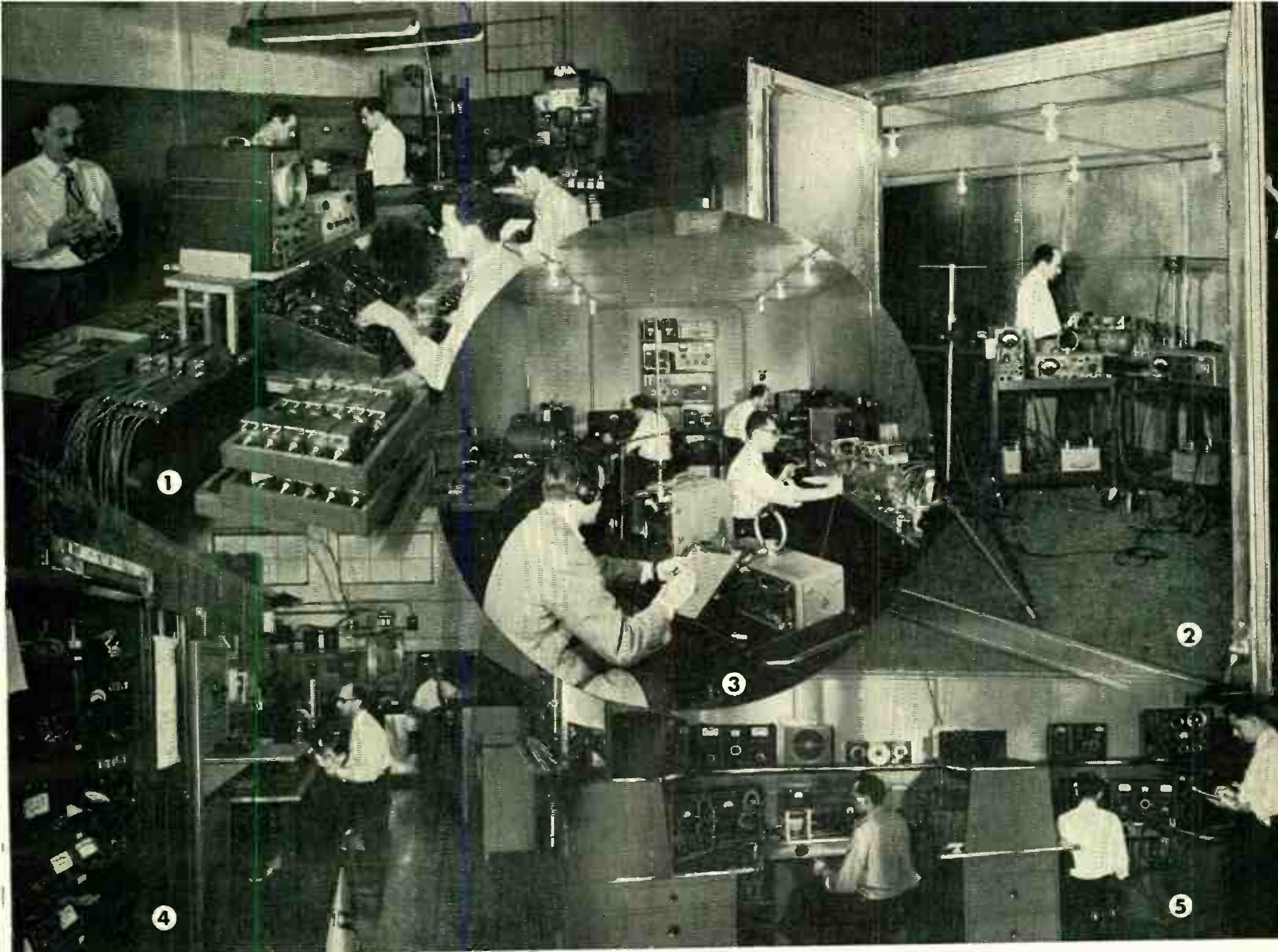
New 16-page booklet contains full specifications on the Type 531 and Type 53B, as well as all Plug-In units and other Oscilloscopes in the Type 530 Series.
Please call your Tektronix Field Office or Representative, or write direct for free copy.



Type 531 Oscilloscope—$995
Type 53B Plug-In Unit—$125
prices f.o.b. Portland (Beaverton), Oregon

**Tektronix, Inc.**

**P. O. BOX 831B, PORTLAND 7, OREGON**
**CYpress 2-2611 • Cable: TEKTRONIX**




*filtered by* FILTRON

Guided Missile

Jet Aircraft

T-38 Sky Sweeper
Self-Aiming
Anti-Aircraft Gun

Atomic Submarine
U.S.S. NAUTILUS

# FROM GUIDED MISSILES TO ATOMIC SUBMARINES...

## FILTRON PROVIDES EXACTLY THE FILTER THEY NEED

1 Electrical Test
2 Shielded Laboratory Measurements
3 Screen Room Interference Testing
4 Environmental Testing
5 Attenuation Test Consoles (per MIL-STD-220)

From the best equipped Radio Interference Laboratories in the world—staffed with the most experienced Radio Interference Engineers, tomorrow's electric and electronic components and systems are made "Radio Interference Free" today.

FILTRON's exceptional facilities are available for the Radio Interference testing AND filtering of your equipment to meet Military Radio Interference Specifications.

Combining engineering facilities, application experience, and manufacturing ability, Filtron competently handles RF interference problems from start to finish.

FILTRON's four plants, with complete production facilities—capacitor manufacturing, coil winding, metal fabricating and stamping, tool and die department, assembly division—are producing more RF interference filters than ever before.

FILTRON—the most dependable name in RF Interference Filters—is the choice of engineers, manufacturers, and military and commercial laboratories the world over.


*Send for our free 20-page RADIO INTERFERENCE FILTER CATALOG.*

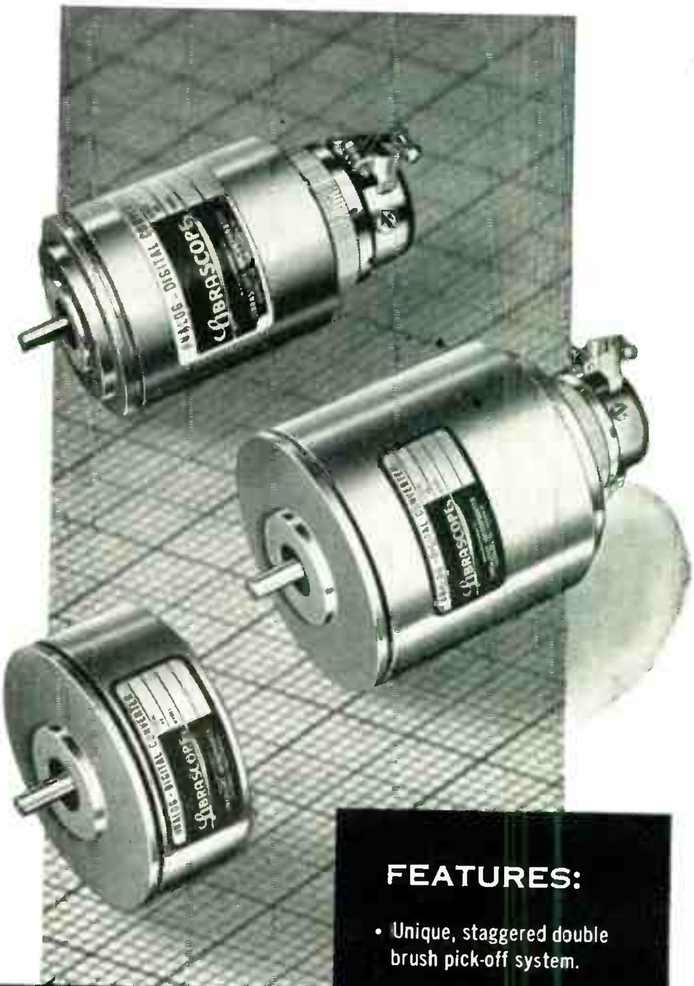
Main Plant, Flushing, N. Y.


**FILTRON CO., INC., FLUSHING, LONG ISLAND, NEW YORK**

**PLANTS IN FLUSHING, NEW YORK, AND LOS ANGELES, CALIFORNIA**

**Flexibility in Application**
**Versatility in design...**

# packaged analog-digital converters




**Shaft Position to Digital Converters** features reliability, long life, non-ambiguity and speed makes these converters ideal for computers or data handling systems where serial read-out is preferred. Librascope converters transmit information at almost any rate desired up to 1 mc and in some cases above, and may be multiple time-shared, holding extra circuitry to a minimum. All units quickly adjustable, syncro-mounted. Available in Binary, Gray code or Binary decimal code as shown in chart below. Special units may be designed to your order.

*Write for catalog information.*

| CODE | MODEL* | RESOLUTION PER INPUT SHAFT REV. | RESOLUTION OVER FULL RANGE | DIMENSIONS DIAMETER X LENGTH |
|---|---|---|---|---|
| BINARY | 7 digit | 128 | 1 part in 128 | 2"x2 49/64" |
| | 13 digit | 128 | 1 part in 8192 | 2"x3 19/32" |
| | 17 digit | 128 | 1 part in 131,072 | 2"x4 13/16" |
| | 19 digit | 128 | 1 part in 524,288 | 2"x4 13/16" |
| BINARY CODED DECIMAL | 0-2000 | 200 | 1 part in 2000 | 3 1/16"x4 27/32" |
| | 0-3600 | 200 | 1 part in 3600 | 3 1/16"x4 27/32" |
| | 0-20,000 | 200 | 1 part in 20,000 | 3 1/16"x4 27/32" |
| | 0-36,000 | 200 | 1 part in 36,000 | 3 1/16"x6 3/8" |
| GRAY | 8 | 256 | 1 part in 256 | 3 1/16"x1 11/16" |
| *SPECIAL UNITS AVAILABLE | Precision gearing<br>Shaft Speed: 120 rpm continuous<br>Operating temp: —55° C to +75° C<br>Shock and Vibration: up to 15 G, 5 to 500 cps. | | Life Expectancy: Function of lead current. For 13 digit unit @ 2 ma. per brush, life approx. 5x10⁸ breaks or makes at approx. 120 rpm. | |

## FEATURES:

- Unique, staggered double brush pick-off system.
- Reads out serially into relays or single or multiple scan matrices.
- Analog-digital or digital-analog operation.
- May be time-shared.
- Syncro-mounted.
- Associated circuitry can be designed to fit your data-handling problems.

*Engineers, physicists and mathematicians interested in challenging California careers, contact Mac McKeague, Personnel Director.*

LIBRASCOPE
COMPUTERS, COMPONENTS AND CONTROLS

A SUBSIDIARY OF GENERAL PRECISION EQUIPMENT CORPORATION

LIBRASCOPE, INC. • 808 WESTERN AVENUE • GLENDALE, CALIFORNIA

# MOISTURE PROOF

## *Connect safely with Cannon under all moisture conditions*

**Splash Proof • Potting Connectors • Watertight**

Protection against moisture is becoming more and more important. You'll find a wide variety of moisture-proof connectors in the Cannon line . . . connectors that solve a wide range of moisture problems . . . from those where only minimum protection is required to those where potting of connectors is needed to give maximum safety and performance. Certain Cannon connectors are even designed for complete submersion in such applications as underwater geophysical exploration.

Typical Cannon Connector Designed for Potting

**Potting** connectors in the AN Series include 12 designs, each in 16 sizes, in both plugs and receptacles, with pin or socket contacts, in the CA group. Potting may be applied, also, to the "K" miniatures, specials and other types. Other moisture-proof connectors in the AN Series include the popular AN-E, OA (Ordnance), special aircraft AF and F types.

For **"average"** moisture resistance, the XKW and BRS Series are recommended.

Where complete **watertight** protection is needed, you may select from the heavy-duty W Series in three AN insert sizes. They may be submerged. 2E (Signal Corps type) is available as a moisture sealed power connector.

Our engineers are available to help you with your moisture protection or potting problems. Write TODAY!

AUDIO | RUGGED SERVICE | UNDER WATER

10th YEAR

*first in connectors*

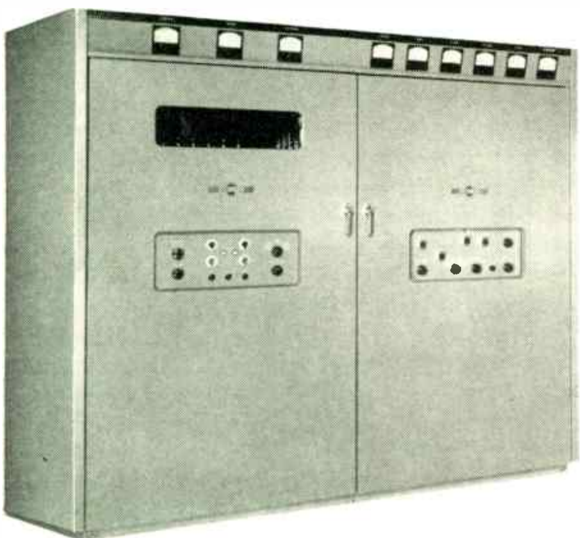
# CANNON PLUGS

Please refer to Dept. 377

CANNON ELECTRIC COMPANY, 3209 Humboldt Street, Los Angeles 31, California.
Factories in Los Angeles; East Haven; Toronto, Canada; London, England. Representatives and distributors in all principal cities.

# *New* VHF Communication Equipment from *Collins* Research and Development in SSB

Single sideband's spectrum and power economies offer a solution for many of today's communication problems. The effects of selective fading and interference due to multipath transmission are minimized through suppression of the carrier and concentrating the r-f power in the intelligence carrying sidebands. Both frequency spectrum space is conserved and the probability of adjacent channel interference reduced by the narrower bandwidth requirements of SSB communications systems. An extensive research and development program at Collins has produced equipment having optimum performance characteristics to realize the full advantages of SSB. Exciters and receivers utilize stabilized master oscillators slaved to a precise frequency standard. Transmitters are easily tuned, efficient and reliable. Channeling facilities incorporate the Collins Mechanical Filter for sideband selection. Typical of Collins equipment available for single sideband communication circuits are:



**205G-1 TRANSMITTER**

## 205G-1 TRANSMITTER

The new Collins 205G-1 Communication Transmitter has many outstanding features. Linear operation with low distortion permits multiplex RTTY and/or voice operation with minimum of interference between channels. The transmitter is manually tuned for fixed frequency operation. Only four tuned circuits are employed and each covers the 30 to 60 mc frequency range. All sub-units and components are accessible from the front of the cabinet.

FREQUENCY RANGE: 30 to 60 mc.

TUNING: Manual over ±2.5% range.

TYPE OF EMISSION: $A_1$, $A_3^b$, or teleprinter signals.

POWER OUTPUT: 20 kw carrier or peak envelope power.

OUTPUT IMPEDANCE: 52 ohms with up to 2 to 1 SWR.

DRIVE REQUIREMENTS: 0.5 watt at carrier frequency at 52 ohms.

SSB DISTORTION: 3rd order distortion products at least 30 db below one tone of a two-tone test signal at 20 kw P.E.P.

HARMONIC OUTPUT: 2nd harmonic is at least 35 db down.

## 50P-1 RECEIVER

The 50P-1 fixed frequency communication receiver consists of an RF amplifier using high Q circuits, first and second i-f amplifiers and mixers. The 250 kc i-f output feeds accessory equipment for recovering the RTTY and voice signals. Maximum rejection of adjacent channel interference together with minimum inter-modulation and cross modulation is provided. When used with the Collins 708B-1 Stabilized Master Oscillator and Collins 40K-1 Frequency Standard, the total frequency error is maintained at less than one part in 100 million (0.000001%) making possible better utilization of spectrum space and attainment of better signal-to-noise ratio by allowing bandwidth requirements to be reduced to a minimum.

FREQUENCY RANGE: 20 mc to 50 mc.

OUTPUT FREQUENCY: 250 kc i-f output.

AMBIENT TEMPERATURE RANGE: 0 to 50 degrees C.

SIZE: 3½" high, 17¼" wide, and 7" deep.

MOUNTING: Relay rack.

POWER INPUT: Supplied by external type 426B-1 power supply — 50 watts.

Complete terminal equipment for generalized data transmission including teletypewriter using synchronous detection techniques is also available to provide a fully integrated system.

***Write for additional information.***

**COLLINS RADIO COMPANY** | **CEDAR RAPIDS, IOWA**

261 Madison Avenue, NEW YORK 16, NEW YORK
1200 18th St. N.W., WASHINGTON, D. C.
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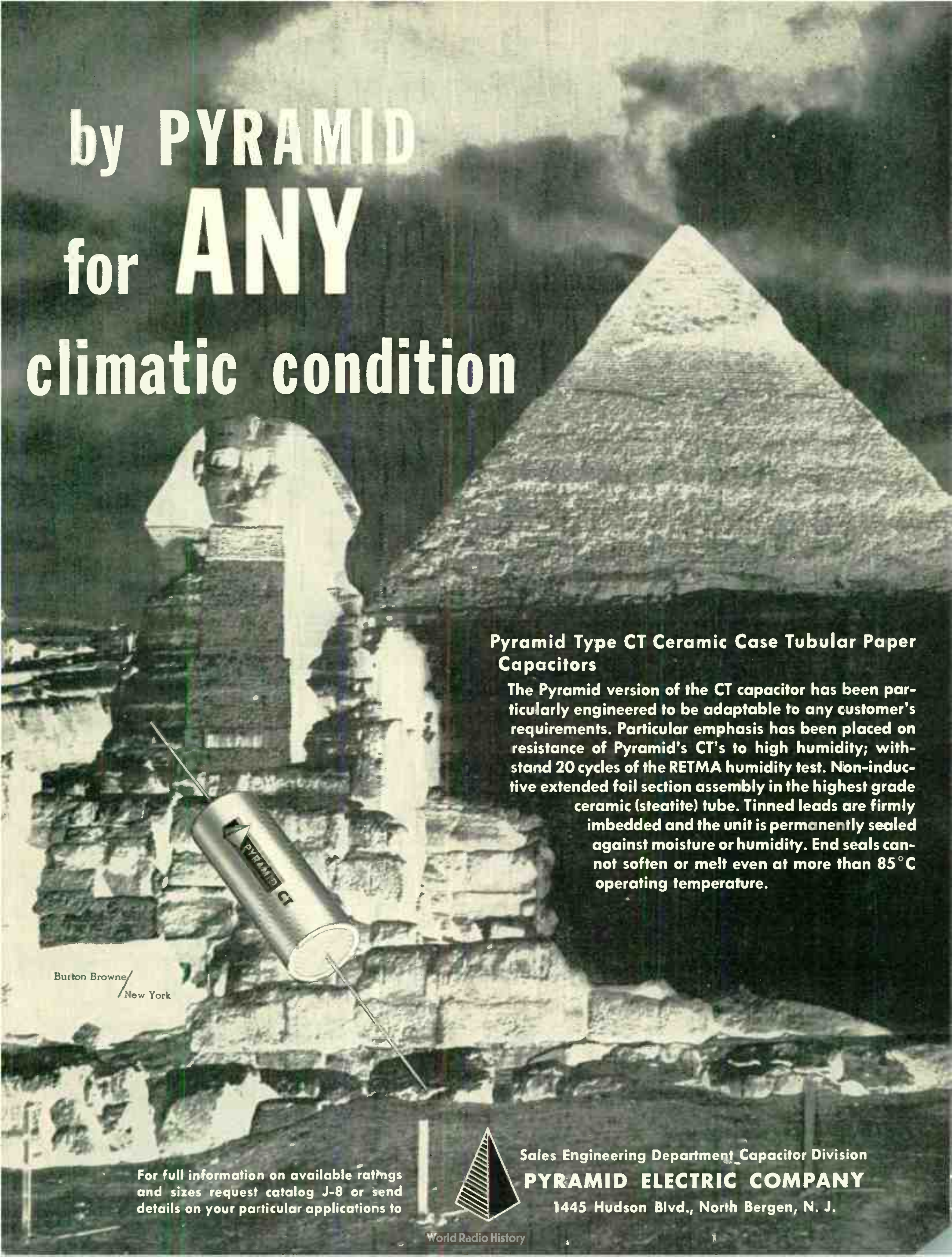
2700 W. Olive Avenue, BURBANK, CALIFORNIA
COLLINS RADIO COMPANY OF CANADA, LTD.
74 Sparks Street, OTTAWA, ONTARIO


by PYRAMID
for ANY
climatic condition
PYRAMID CT
Pyramid Type CT Ceramic Case Tubular Paper Capacitors
The Pyramid version of the CT capacitor has been particularly engineered to be adaptable to any customer's requirements. Particular emphasis has been placed on resistance of Pyramid's CT's to high humidity; withstand 20 cycles of the RETMA humidity test. Non-inductive extended foil section assembly in the highest grade ceramic (steatite) tube. Tinned leads are firmly imbedded and the unit is permanently sealed against moisture or humidity. End seals cannot soften or melt even at more than 85°C operating temperature.
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For full information on available ratings and sizes request catalog J-8 or send details on your particular applications to
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PYRAMID ELECTRIC COMPANY
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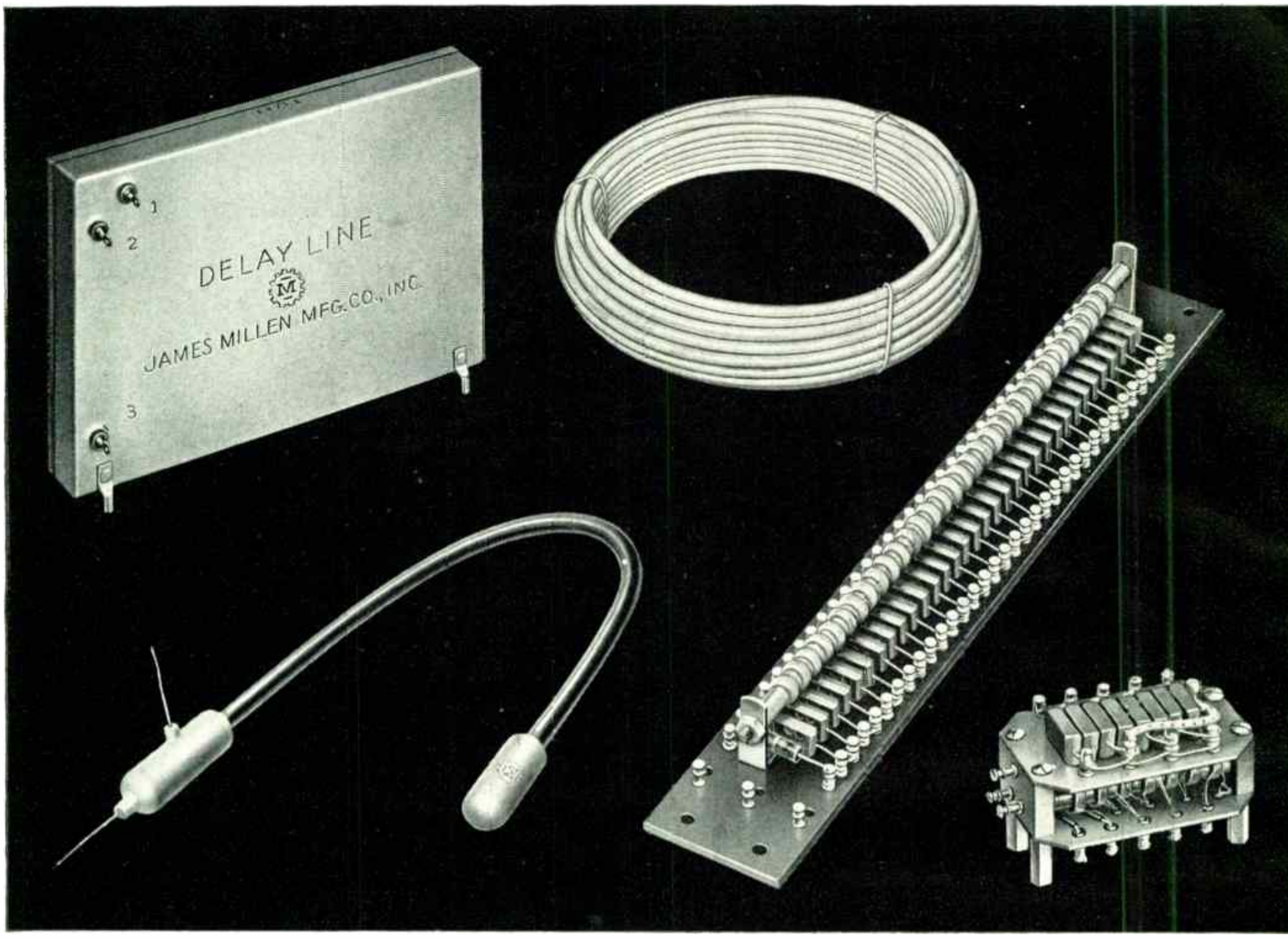



## *"Designed for Application"*

# Delay Lines and Networks

The James Millen Mfg. Co., Inc. has been producing continuous delay lines and lump constant delay networks since the origination of the demand for these components in pulse formation and other circuits requiring time delay. The most modern of these is the distributed constant delay line designed to comply with the most stringent electrical and mechanical requirements for military, commercial and laboratory equipment.

Millen distributed constant line is available as bulk line for laboratory use and in either flexible or metallic hermetically sealed units adjusted to exact time delay for use in production equipment. Lump constant delay networks may be preferred for some specialized applications and can be furnished in open or hermetically sealed construction. The above illustrates several typical lines of both types. Our engineers are available to assist you in your delay line problems.

**JAMES MILLEN MFG. CO., INC.**
MAIN OFFICE AND FACTORY
**MALDEN, MASSACHUSETTS, U.S.A.**

# Direct, automatic power readings

## SPECIFICATIONS

**Power Range:** 5 ranges, front panel selector. Full scale readings of .1, .3, 1, 3 and 10 mw. Also continuous readings from —20 to +10 dbm. (0 dbm = .001 watt). Power range may be extended with attenuators or directional couplers in microwave system.

**External Bolometer:** Frequency range depends on bolometer mount. Bolometers can operate at resistance levels of 100 or 200 ohms and can have positive or negative temperature coefficients. Any dc bias current up to 16 ma is available for biasing positive or negative temperature coefficient bolometers. Dc bias current is continuously adjustable and independent of bolometer resistance and power level range.

Suitable bolometers are:

Instrument fuses: *-hp-* G-28A 1/100 amp fuse:

Barretters: Sperry 821, Narda N821B or N610B, PRD 610A, 614, 617 or 631C.

Thermistors: W. E. D166382 and 32A3, V. E. Co. 32A3, 32A5, Narda 333, 334.

**Accuracy:** ±5% of full scale reading.

**Power:** 115/230 v ±10%, 50/1,000 cps, 75 watts.

**Dimensions:** Cabinet Mount: 7⅜" wide, 11½" high, 12¼" deep. Rack Mount: 19" wide, 7" high, 12½" deep.

**Weight:** Net 20 lbs. Shipping 32 lbs. (cabinet mount).

**Price:** $250.00.

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

***CW or pulsed power***

***Wide frequency range***

***No calculations***

***Assured accuracy***

***Operates with wide variety of bolometers***

## *New! -hp- 430C*
## Microwave Power Meter

Here is the newest, finest, most dependable source of instantaneous microwave power readings available today. The new *-hp-* 430C gives you power readings direct in db or mw and completely eliminates tedious computations or troublesome adjustment during operation. The instrument measures either pulsed or CW power on either waveguide or coaxial systems. Operation is entirely automatic, stability is extremely high, and the meter may be used with a wide variety of bolometer mounts having either positive or negative temperature coefficients. The broad nominal measuring range can be extended to higher powers by means of directional couplers and attenuators.

For measurements of CW or pulsed power, *-hp-* 430C uses either an instrument fuse, barretter or thermistor as a bolometer element. Operation may be at either 100 or 200 ohms. Power is read direct in milliwatts from 0.02 to 10 mw, or in dbm from —20 to +10 dbm.

**HEWLETT-PACKARD COMPANY**

**3341D PAGE MILL ROAD • PALO ALTO, CALIFORNIA, U. S. A.**

**Cable "HEWPACK"**

***FIELD REPRESENTATIVES IN ALL PRINCIPAL AREAS***

**hp® ELECTRONIC MEASURING INSTRUMENTS**

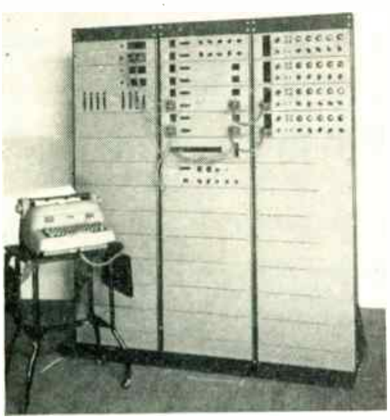
# NEWS and NEW PRODUCTS

**July 1955**

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

## Digital-Analog Computer

**Wang Laboratories,** 37 Hurley St., Cambridge, Mass., announces design and production of the "WEDILOG," a new electronic digital-analog differential computer, with applications in aeronautical design, trajectory problems, process control, dynamic systems analysis, etc. Although it is a digital computer, "WEDILOG" operates on the principle of functional simulation, using numbers instead of voltages. Designed for problems in the physical sciences and engineering, this computer will solve linear and non-linear ordinary and partial differential equations, integral equations, and simultaneous differential and algebraic equations. "WEDILOG" combines the simplicity of problem set-up of the DC analog computer with the accuracy and resolution of a full scale digital computer. The machine resolution is five decimal digits. All numbers are handled as true number with sign.



The basic "WEDILOG" computer is made up of a combination of several types of computation units operated from a central control. The problem, itself, determines the number and types of units and how they are "patched" together. Unitized construction of these computation units enables easy expansion of the basic computer, as more comprehensive problems arise. Outputs are available as electric typewriter tabulations, multiple channel recordings, on plotting boards, or on punch cards. In addition, all variables are indicated by neon lights, readable at all times.

## Ham Tube Catalog

**Tube Div. Radio Corp. of America,** 415 S. 5th St., Harrison, N. J., has just brought out a completely revised edition of the 4-page folder "Headliners for Hams."

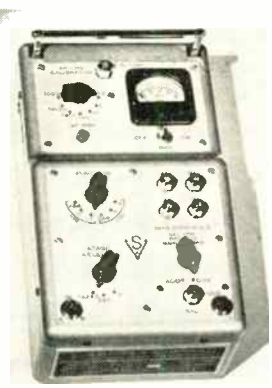
This folder covers 45 popular RCA "Hams" types—oscillators, amplifiers, frequency multipliers, voltage regulators, thyratrons, rectifiers, oscillograph types for test equipment, and camera tubes for use in amateur telecasting.

"Headliners for Hams" contains a tube line-up chart for amateur transmitters; operating conditions for class C amplifier and oscillator, modulator, and frequency-multiplier service; single-sideband tube data; and latest "Ham" ratings on popular receiving-tube types.

Members can obtain copies of "Headliners for Hams" from local RCA Tube Distributors, or direct from Commercial Engineering, Tube Division, Radio Corporation of America, Harrison, N. J.

## Oscilloscope Pre-Amplifier

The new VS-61 A oscilloscope preamplifier developed by **Volkers & Schaffer Mfg. Corp.,** Box 996, Schenectady, N. Y., is a highly sensitive, dual input, adding or differential amplifier, having substantially less than 1 $\mu$V RMS noise. It incorporates the latest results in low-noise amplifier-research. Volkers & Pedersen have shown (1955 IRE National Convention) that transistors, contrary to general experience and opinion, are inherently less noisy than vacuum tubes, if suitable operating parameters are selected. The resulting new "Hushed Transistor Amplifiers," having less than 1 $\mu$V RMS noise over a frequency band of 60 kc, provides the input stages of the VS-61 A. Additional amplification is provided by vacuum tubes.



The amplifier has a stage selector switch which provides a choice of either straight vacuum tube amplification (maximum gain 10, input impedance 100 K, frequency response 2 cps to 250 kc) or combined transistor and vacuum tube amplification (available maximum gains 200 and 1000, input impedance 1 K, frequency response 2 cps to 60 kc). All frequency responses are given for the 3 db down point. The maximum noise with straight tube operation is 5 $\mu$V RMS over the full 250 kc pass-band and with transistor and tube operation usually much less than 0.5 $\mu$V RMS over the full 250 kc pass-band.

## Power Supply

The new Model UHR-240 regulated power supply developed by **Krohn-Hite Instrument Co.,** 580 Massachusetts Ave., Cambridge 39, Mass., provides up to ½ ampere of direct current at 0–500 volts with 0.001 per cent regulation and less than 100 microvolts of ripple. The stabilization for ±10 per cent change in line voltage is better than 0.003 per cent.

The dc and low frequency impedance is less than 0.005 ohms. The ac impedance is less than 0.05 ohms in series with 0.1 microhenry (4 inches of wire). Transient response is 0.001 millisecond. Typical ten hour drift is 300 ppm plus 20 millivolts.

The ultra-high regulation applies over the entire operating range. For line voltages between 105 and 125 volts, the full maximum current can be drawn continuously at any output voltage.

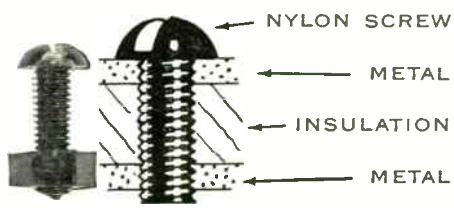
There is an additional 0–150 volt, 0–5 ma negative supply with 0.05 per cent stabilization and less than 2 millivolts of ripple. A 5–13 volt, 0–2.5 ampere dc heater supply with less than 20 millivolts of ripple is included in addition to two independent 6.3 volt ac, 10 ampere heater supplies.

The two 3½ inch front panel meters are ruggedized and hermetically sealed.

Dimensions are 17½ wide, 9 high, and 15½ inches deep—also available for rack mounting. Price is $550.00, f.o.b. factory.

## Nylon Screws & Nuts

**Weckesser Co.,** 5261 N. Avondale Ave., Chicago 30, Ill., has available from inventory 10 stock items of screws and nuts of molded black nylon.




These non-magnetic, non-corrosive, light weight screws and nuts come in sizes $\frac{6}{32}$, $\frac{8}{32}$ and $\frac{10}{32}$.

(Continued on page 20A)

## POWER AND IMPEDANCE MEASUREMENTS

# DAVEN

# OUTPUT POWER METERS

*of unexcelled accuracy and reliability*

## have many applications

### TYPE OP-182

Impedance Range: 2.5 ohms to 20,000 ohms. Remains essentially resistive over frequency range of 30 to 10,000 cps. Accuracy ±5%.

Power Range: 0.1 mw. to 5 watts in steps of 0.1 mw.

Indicating Meter: Calibrated from 0 to 50 milliwatts and from 0 to 17 db. —Zero level: 1mw.

Meter Multiplier: Will change reading of indicating meter by ratios of 0.1:1, 1:1, 10:1, 100:1, or decibel reading by —10, 0, +10, +20.

The DAVEN Output Power Meters are designed to measure the actual power delivered by an audio signal system to a given load. However, because of the characteristics of the circuit, they are admirably suited to other applications, namely:

1. **Determination of Characteristic Impedance of an A.C. Source.**
2. **Effects of Load Variation on a Signal System.**
3. **Transmission Line Equalization Measurements.**
4. **Measurement of Insertion Loss in Multi-channel Mixer and other complex circuits.**
5. **Filter and Transformer Measurements.**
6. **Radio Receiver Measurements.**

The equipment shown on this page is built to DAVEN'S well-known standards of precision. Please write for more detailed data. Let our engineering department help you on specific problems.

### TYPE OP-962

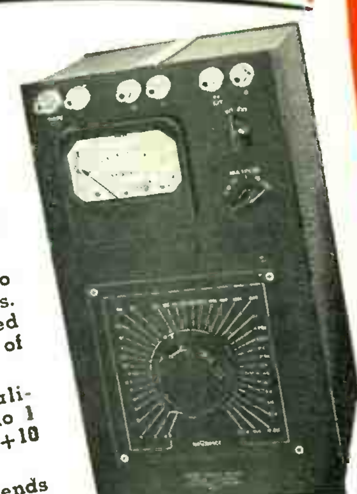
Characteristics similar to OP-961, except that it can measure up to 100 watts.

Impedance Range: 40 selected impedances between 2.5 and 20,000 ohms. Accuracy ±2% over frequency range 30 to 10,000 cycles.

Power Range: 0.1 mw to 100 watts in 0.1 mw steps. Range may be extended below 0.1 mw by use of external amplifier.

Indicating Meter: Calibrated from .01 watt to 1 watt and from —10 to +10 db. Zero level: 1mw.

Meter Multiplier: Extends range of meter from 0.01 to 100 times scale reading.



### TYPE OP-961

Impedance Range: 2.5 ohms to 20,000 ohms. Remains essentially resistive over frequency range of 30 to 10,000 cps. Accuracy ±2%.

Power Range: 0.1 milliwatts to 50 watts in steps of 0.1 milliwatts.

Indicating Meter: Calibrated from 1 to 50 milliwatts and 0 to 17 decibels. Zero level: 1mw.

Meter Multiplier: Extends the power reading of the indicating meter from 0.1x to 1,000x scale value, or the db. reading from —10 to +30 db. in steps of 2 db.

## THE DAVEN CO.

195 CENTRAL AVENUE • NEWARK 4, NEW JERSEY

## News-New Products

**These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.**

*(Continued from page 18A)*

An advantage of these screws is that they are insulators in themselves, and require no insulating sleeves, bushings or washers.

The nylon is the hardest grade and it may be used to 250°F. It resists weathering and degrading effect due to ultra-violet light because of the black dye impregnation.

### New Crystal

A new frequency control unit has been introduced by the **James Knights Co.**, Sandwich, Ill. It is named the "Thermystal," and represents the integrated packaging of crystal and oven to provide stability and environmental control within limited price range.


The JK Thermystal offers the following performance data: higher merit factor; vacuum enclosure increases Q of crystal; calibration accuracy; 1 cycle 0.0001 per cent; temperature stability; 30 to 900 kc 0.0001 per cent, 1000 kc to 150 mc 0.00005 per cent. Oven temperature varies less than 1°C over ambient range of −55°C to +85°C, secular stability; Less than 0.001 per cent per year. Crystal is specially processed and sealed in glass enclosed vacuum; low oven power; 6.3 volts at 1.5 ampere maximum. The thermostat cycles less than 3 times per minute at room temperature. The unit stabilizes in less than 10 minutes when turned on at −55°C.

*(Continued on page 145A)*

*Raytheon — World's Largest Manufacturer of Magnetrons and Klystrons*


# now available

## NEW Consolidated Data Booklets for Raytheon Magnetrons, Klystrons and Special Tubes

### WRITE FOR YOUR COPIES

These valuable free data booklets, which we will be glad to send you, list most of the principal unclassified types manufactured. They give maximum ratings, typical operating values, frequency ranges, and power levels. Indispensable to every engineer's file.

Raytheon is the world's largest producer of CW and pulse magnetrons, available in either fixed or tunable designs. A compact, efficient source of microwave power, magnetrons are being used in a growing number of applications.

Raytheon klystrons, which can be tuned mechanically or thermally, fit the widest range of requirements from 550 to 60,000 Mc.

The special tubes include backward wave oscillators, square law tubes, storage tubes and others.

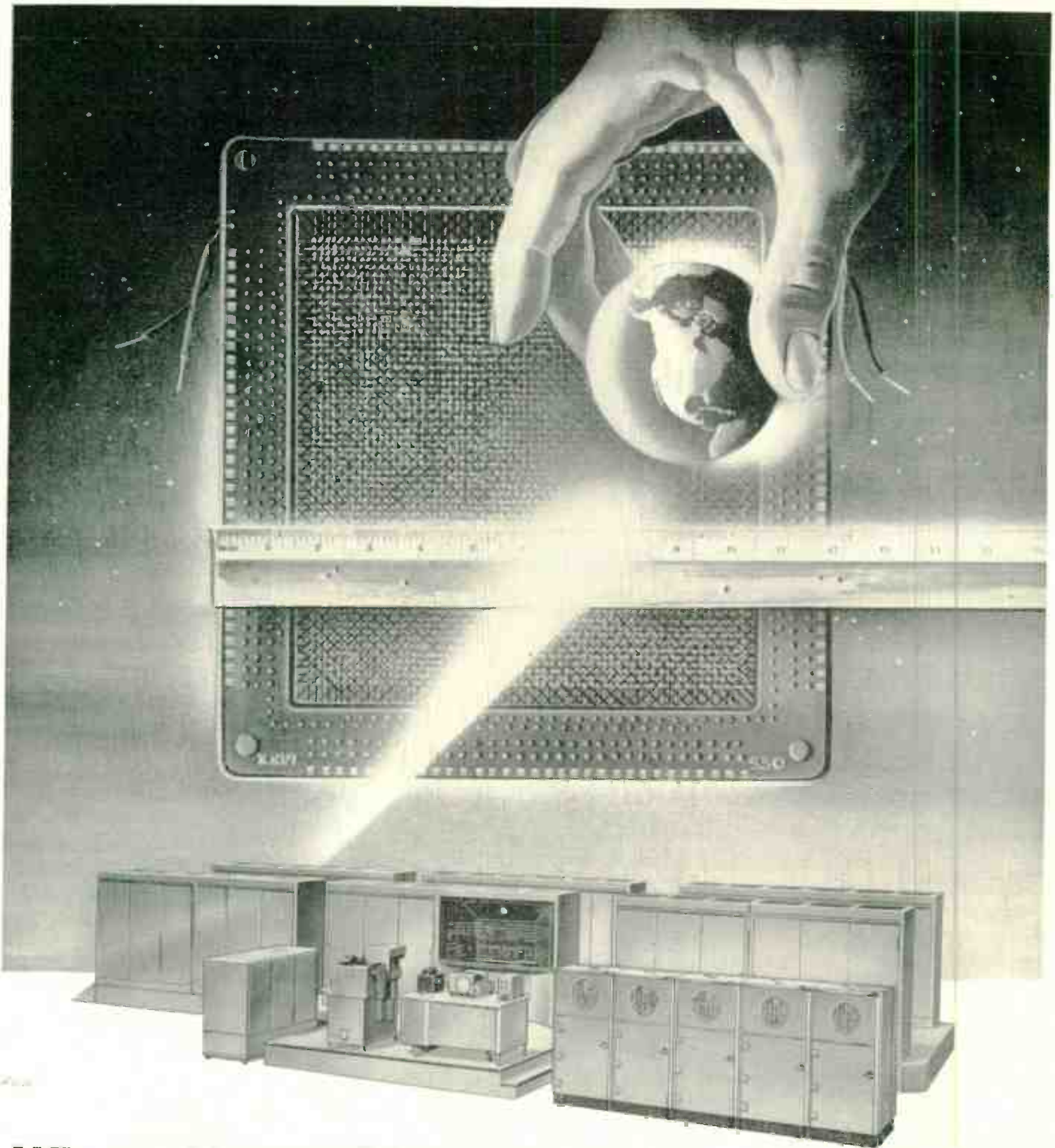
Write today for your data booklets—or telephone WAltham 5-5860. There is no obligation, of course.

*Excellence in Electronics* RAYTHEON

**RAYTHEON MANUFACTURING COMPANY**

**Microwave and Power Tube Operations, Section PL-31, Waltham 54, Mass.**

Raytheon makes: Magnetrons and Klystrons, Backward Wave Oscillators, Traveling Wave Tubes, Storage Tubes, Power Tubes, Receiving Tubes, Transistors



# What's New in Mnemonics?

The Univac Scientific Computing System

The news is that the magnetic-core memory has emerged from the computer laboratory and has been in customer use for approximately a year, passing all tests with flying colors. This new development has been pioneered by Remington Rand with the Univac Scientific—the first installation of a commercially available computer that successfully uses magnetic-core storage.

Mnemonics, says Webster, is "the art of improving the efficiency of the memory." And, as far as electronic computers are concerned, Remington Rand has clearly established its leadership in this art.

Illustrated above is a single plane of core storage, each of which holds 4,096 binary bits of information. Planes are wafer thin, and a stack capable of "remembering" 147,456 bits would measure only 13 inches in depth. The speed, economy, and reliability of this magnetic-core memory are now available in the new Univac Scientific Models 1103A and 1103B.

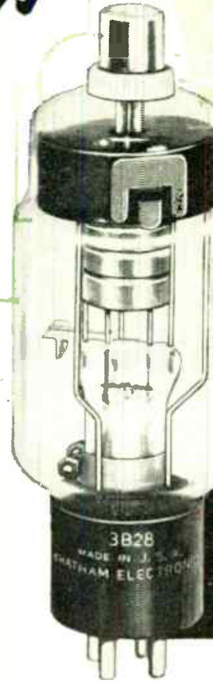
For the latest information about the Remington Rand magnetic-core memories or about the Univac Scientific, write, on your business letterhead, to . . .

Electronic Computer Department, Room 1744

**Remington Rand**
DIVISION OF SPERRY RAND CORPORATION

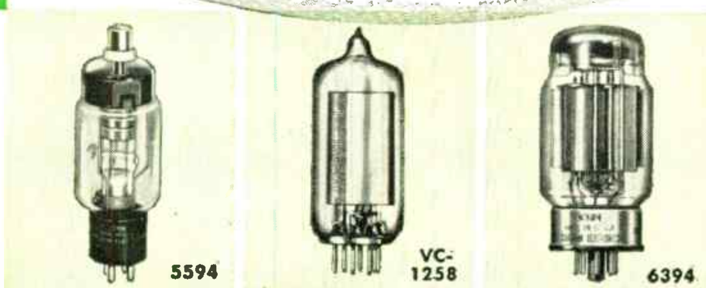
315 Fourth Avenue, New York 10, New York

# CHATHAM advance-designed yesterday — in industry-wide use today!


**AMPLIFIERS • REGULATORS • INERT GAS AND MERCURY RECTIFIERS • MERCURY, INERT GAS AND HYDROGEN THYRATRONS**


# CHATHAM
## SPECIAL-PURPOSE
# TUBES




- **3B28 RECTIFIER**
  Rugged half-wave Xenon filled rectifier. Operates in any position. Ambient temperature range −75° to +90°C. Inverse peak anode voltage 10,000, average current .25 amps. Filament 2.5v., 5 amp.
- **4B32 RECTIFIER**
  Ruggedly built, half-wave Xenon filled rectifier. Ambient temperature range −75° to +90°C. Inverse peak anode voltage 10,000, average anode current 1.25 amp. Filament 5v., 7.5 amp.
- **VC-1258 MINIATURE HYDROGEN THYRATRON**
  for pulse generation. Handles 10 kw peak pulse power.
- **6336 TWIN TRIODE**
  for voltage regulation. Features high plate dissipation, hard glass envelope.
- **6394 TWIN TRIODE**
  Similar to 6336 except 26.5 volt heater instead of 6.3 volt heater.
- **5594 XENON THYRATRON**
  Operates over wide ambient temperature range −55°C to +90°C.

*STANDARD TYPES DIRECT FROM STOCK*
*PLUS SPECIAL DESIGNS BUILT TO REQUIREMENTS*

Chatham specializes in the development of general and special purpose tubes for both electronic and industrial applications. Many of the tubes originally developed by Chatham to fill a specialized need, now number among the most widely used tubes in the industry. For complete information on Chatham tubes—either stock items or types built to your requirements—call or write today.

**Chatham Electronics**
*DIVISION OF GERA CORPORATION—LIVINGSTON, NEW JERSEY*


## ... the Sangamo *Telechief*

For critical applications such as hi-fi equipment, computers and other electronic gear . . . applications which require exceptionally high insulation resistance and unusual stability at high temperatures, your best bet is a Sangamo Telechief.

It is the molded paper tubular which, tests by leading manufacturers show, outperforms all other paper tubulars in . . .

**MOISTURE RESISTANCE**

Sangamo paper tubulars are molded in Humiditite, the remarkable plastic molding compound which gives them moisture resistance properties far superior (10 to 15 times greater) than any other molded tubular capacitor.

**HIGH TEMPERATURE OPERATION**

The resistance qualities of Humiditite also make Telechief the winner over other paper tubulars in this department. (Tests show top performance in temperatures up to 125° C.)

**HOLDING RATED CAPACITY**

The exclusive Sangamo solid dielectric impregnant enables the Telechief to hold its rated capacity under all conditions . . . makes it a really rugged paper tubular.

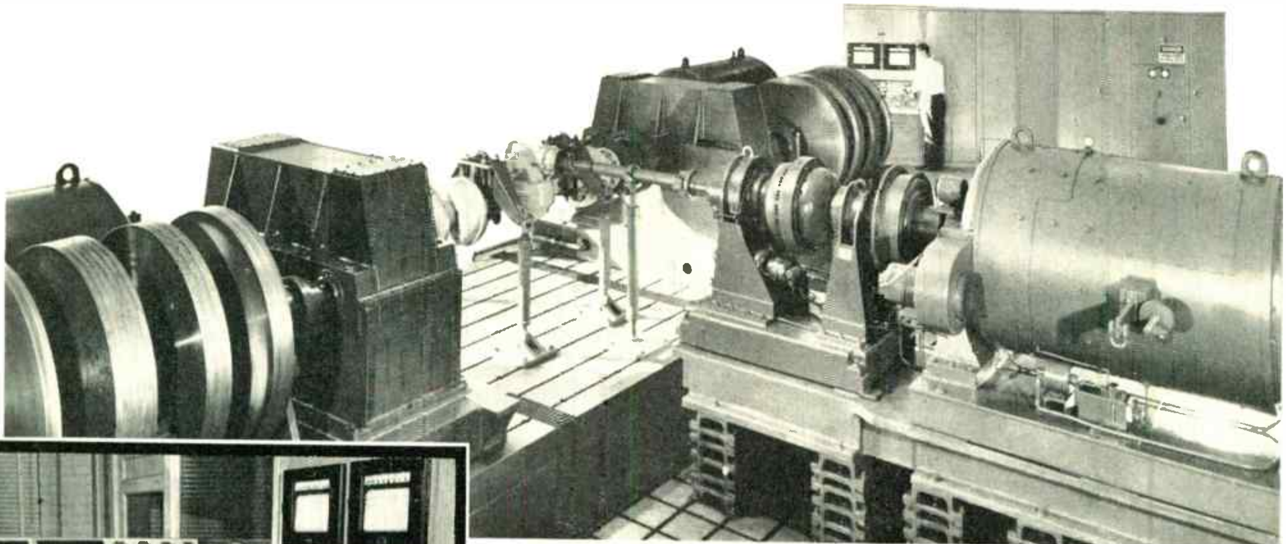
Because the Telechief outperforms other paper tubulars in all of these areas, you can be sure that here is a paper tubular which will deliver long, trouble-free capacitor life.

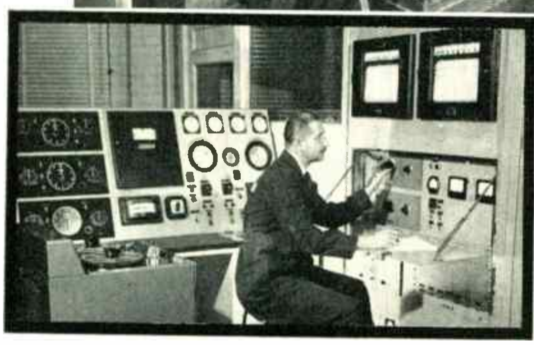
SC55-11

**SANGAMO ELECTRIC COMPANY**
MARION, ILLINOIS

# MAGNETIC TAPE RECORDING

## *helps **Road-Test** Timken Truck Axles*





An Ampex Model 306 Recorder programming axle-test dynamometers at Timken-Detroit.

Magnetic tape recordings are now being used to duplicate rugged road-tests at the Timken-Detroit Axle Division of the Rockwell Spring and Axle Company, Detroit, Michigan. A four hour tape cycle is made of actual road surface and driving conditions . . . then played back through torque and speed dynamometers — repeatedly — until a test axle breaks down.
Result: more realistic and efficient testing — better axles for today's trucks, buses and trailers.

### WHY TIMKEN CHOSE AMPEX

Timken engineers required a recording and playback medium that could give near-perfect reproduction of the original road test phenomena . . . and would playback indefinitely without introducing errors through wear and speed irregularities. They found that the Ampex F-M recorder best met these exacting requirements. Its extreme stability of tape motion, precise timing and consistent accuracy produced laboratory "road-test" results within 1% of actual conditions.

### LET AMPEX STUDY YOUR REQUIREMENTS

Ampex manufactures the most complete line of magnetic recorders for complex and sensitive automation, communication and data-handling systems. Why not let Ampex application engineers determine what magnetic tape recording can do for you?

***For further information,*** *send for our 16-page illustrated bulletin, "Data Recording, Machine Control and Process Regulation." Contact your nearest Ampex representative or write to Dept. G-1897*

AMPEX
CORPORATION

ANOTHER APPLICATION BY THE INSTRUMENTATION DIVISION OF

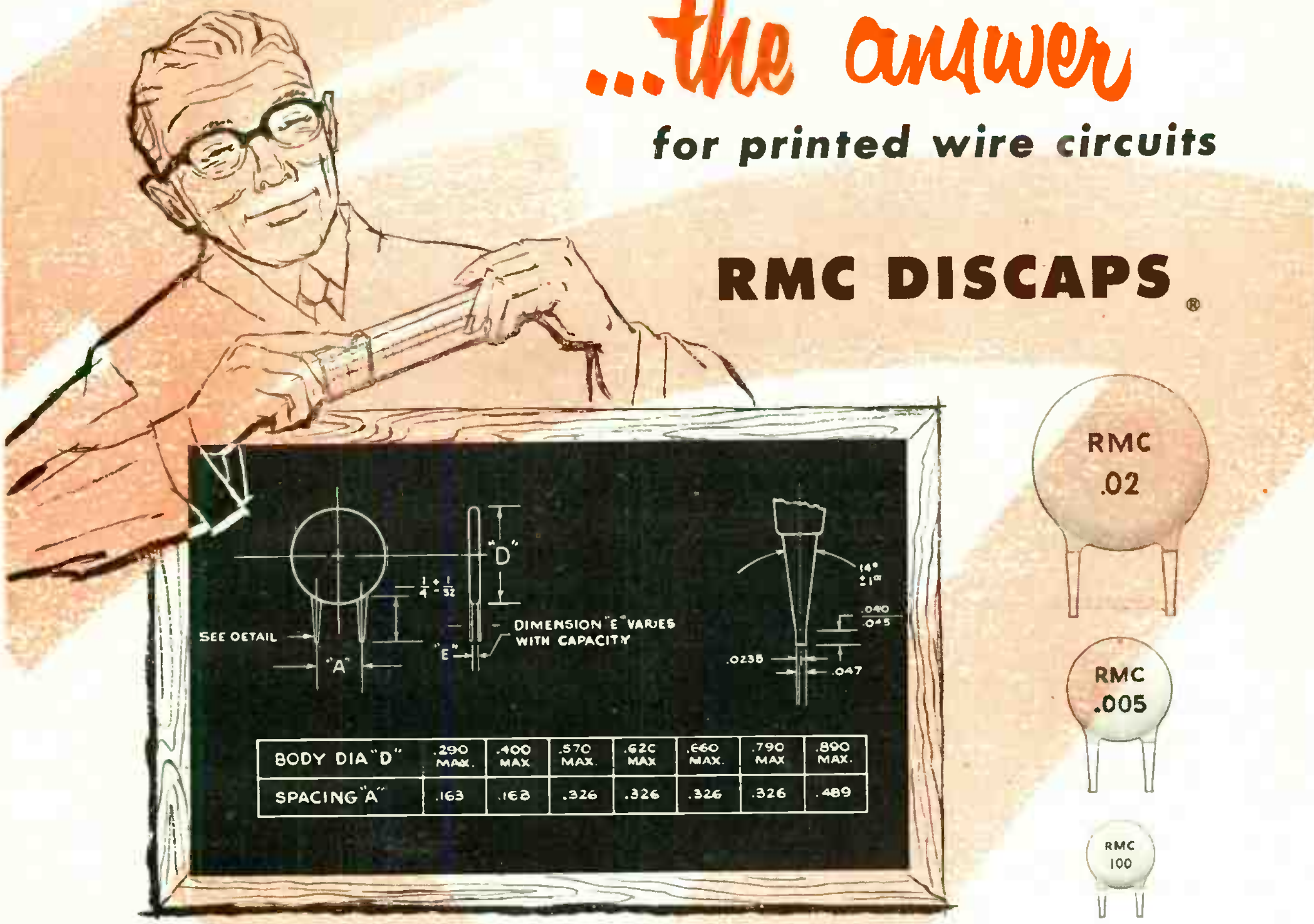
AMPEX CORPORATION • 934 CHARTER STREET, REDWOOD CITY, CALIFORNIA

**Branch Offices:** New York; Chicago; Atlanta; San Francisco; College Park, Maryland (Washington D.C. area).
**Distributors:** Radio Shack, Boston; Bing Crosby Enterprises, Los Angeles; Southwestern Engineering & Equipment, Dallas and Houston; Canadian General Electric Company, Canada.

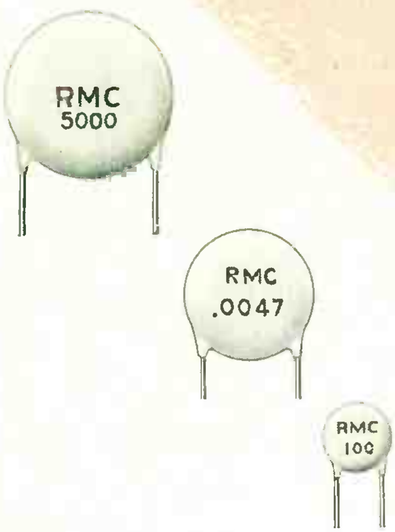
...the answer

**for printed wire circuits**

# RMC DISCAPS®




| BODY DIA "D" | .290 MAX. | .400 MAX | .570 MAX | .620 MAX | .660 MAX. | .790 MAX | .890 MAX. |
|---|---|---|---|---|---|---|---|
| SPACING "A" | .163 | .168 | .326 | .326 | .326 | .326 | .489 |




**Wedg-loc...** The exclusive wedge leads on these DISCAPS lock securely in place on printed circuit assemblies prior to the soldering operation. There is no possibility of the capacitors becoming loose or falling out and the soldered connection is always uniform.

Available in capacities between 2 MMF and 20,000 MMF, Wedg-loc DISCAPS can be furnished in temperature compensating, by-pass, and stable capacity types. Suggested hole size is a .062 square.

**Plug-in...** RMC plug-in DISCAPS are designed to simplify production line problems on printed circuits. Leads are No. 20 tinned copper (.032 diameter) and are available up to 1½″ in length. Plug-in DISCAPS are manufactured in temperature compensating, by-pass, and stable capacity types and include the mechanical and electrical features that have made standard DISCAPS the favorite of leading manufacturers.

Write today on your company letterhead for expert engineering help on any capacitor problem.

DISCAP CERAMIC CAPACITORS

RMC

**RADIO MATERIALS CORPORATION**

GENERAL OFFICE: 3325 N. California Ave., Chicago 18, Ill.

FACTORIES AT CHICAGO, ILL. AND ATTICA, IND.

Two RMC Plants Devoted Exclusively to Ceramic Capacitors

# Introducing...

## the Eimac

# 4X250B

## Radial-beam power tetrode

- **Higher Power**
- **Easier Cooling**
- **Longer Life**

ACTUAL SIZE

**4X250B,** a new, superior radial-beam power tetrode by Eimac—originators of the famous 4X150A—is now available. Unilaterally interchangeable with the 4X150A in practically all applications, this amazing new bantam for modulator, oscillator and amplifier application from low frequencies into UHF, offers these advantages:

**HIGHER POWER**—Electrical advances permit an increased plate dissipation rating of 250 watts, plate voltages to 2000 volts and doubled plate power input capabilities of 500 watts.

**EASIER COOLING**—Development of the Eimac integral-finned anode makes cooling so easy that only one-third the air-pressure and one-half the cubic feet of air are required. Forced air is unnecessary during standby periods.

**LONGER LIFE**—A newly designed, highly efficient oxide cathode and increased temperature tolerances, coupled with Eimac-developed production and testing techniques enable the 4X250B to meet the most critical standards. New techniques in grid production, high vacuum outgassing and product evaluation are among the features that insure uniform incomparable quality and more hours of top performance.

The small, rugged, versatile 4X250B is now available for existing sockets or sockets of yet-to-be-designed equipment demanding optimum quality and performance.

### TYPICAL OPERATION

(per tube, frequencies to 175mc)

**4X250B radial-beam power tetrode**

| | Class C CW FM Phone | Class C AM Phone | Class AB RF Linear |
|---|---|---|---|
| D-C Plate Voltage | 2000v | 1500v | 2000v |
| D-C Screen Voltage | 250v | 250v | 350v |
| D-C Grid Voltage | -90v | -100v | -60v |
| Zero Sig D-C Plate Current | — | — | 50ma |
| D-C Plate Current | 250ma | 200ma | 250ma* |
| D-C Screen Current | 12ma | 10ma | 5ma* |
| D-C Grid Current | 22ma | 23ma | 0ma* |
| Peak RF Grid Voltage | 114v | 125v | 60v* |
| Driving Power | 2.5w | 2.9w | — |
| Plate Power Input | 500w | 300w | 500w* |
| Plate Power Output | 400w | 240w | 325w* |

*Maximum Signal

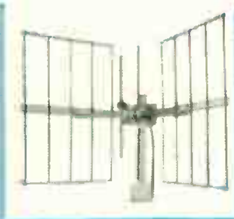
*For further details contact our Technical Services Department.*

Eimac

**EITEL-McCULLOUGH, INC.**

SAN BRUNO • CALIFORNIA

**The World's Largest Manufacturer of Transmitting Tubes**



# BARGAINS in POWER

**No single piece of radio equipment can equal the antenna for economically increasing effective power.**

One of the less expensive components in a radio communications installation is the antenna. Yet the antenna, which usually represents less than ten per cent of the total equipment cost, can multiply the effective power of every transmitter in the system *several hundred per cent*. Equally true, a poorly designed or inappropriate antenna can waste the power produced by the costly equipment behind it.

**Manufacturers of the UNIPOLE, High Gain, Corner Reflector, Parabolic and Yagi Antennas**

In planning a new system, selection of the proper antenna often will allow a lower power transmitter to achieve desired signal range. For existing systems, the use of a higher gain antenna will reduce "dead spots."

Andrew is a pioneer in designing and developing antennas. We make over 30 standard types for microwave, broadcast and mobile communications. Special models or adaptions of standard models are readily made to order.

Write or phone Andrew for a dollars-and-cents evaluation of the type of antenna that can give your installation the greatest bargain in power.

**Andrew CORPORATION**
363 EAST 75th STREET · CHICAGO 19

OFFICES: NEW YORK • BOSTON • LOS ANGELES • TORONTO

# New Mallory Cardboard Tubular Capacitors

## *...premium performance at no increase in cost*

Never before has quality like this been built into cardboard tubular electrolytics. At no increase in price, this new series developed by Mallory offers you a combination of features unique in this type of capacitor:

**Minimum size, high ripple current ratings, low RF impedance** . . . obtained through use of genuine fabricated plate anodes.

**Long life, high stability temperature rating up to 75°C** . . . due to fabricated anode and etched cathode.

**Low leakage current.**

**Low-resistance tab-to-lead wire connections** . . . welding ends danger of intermittent or high resistance connection.

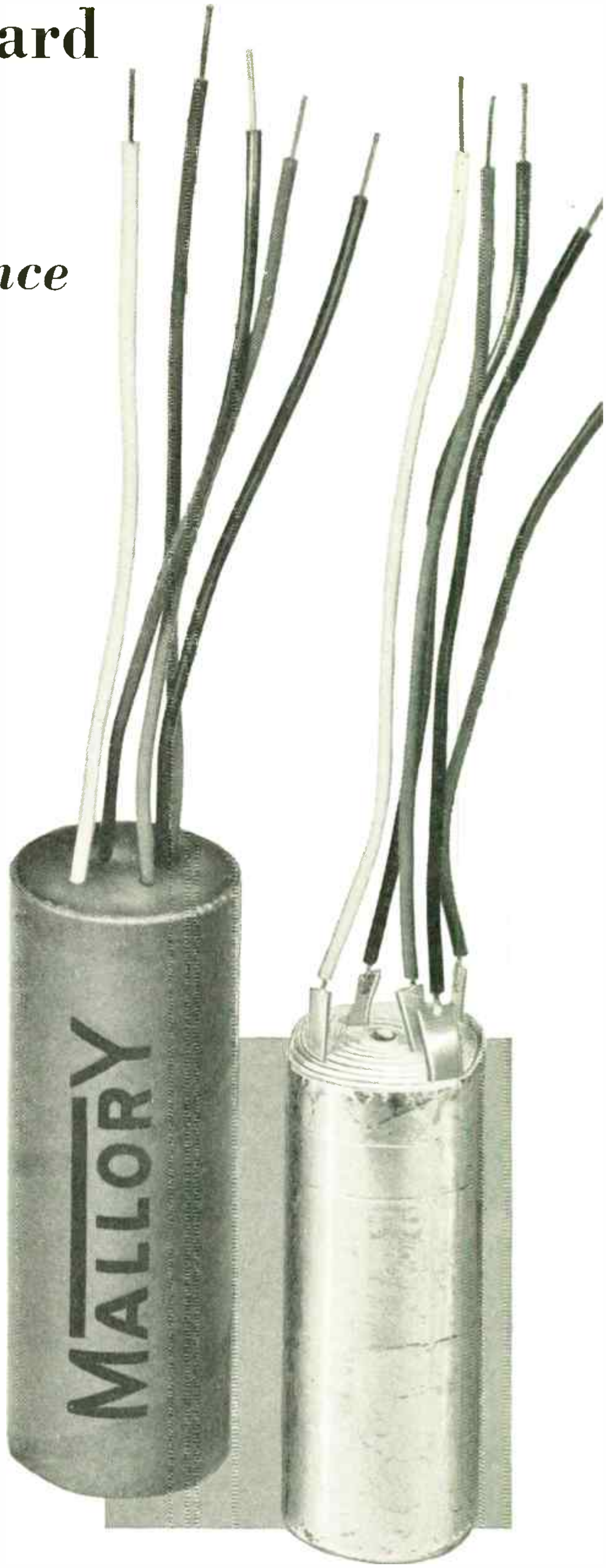
**Low moisture loss** . . . cartridge is foil wrapped; wax impregnated cardboard tube is sealed with wax at both ends.

**High dielectric strength,** exceeds U.L. requirements, due to improved low-moisture absorbent separators.

**Rugged, flexible leads** . . . covered with plastic insulation rated for 105° C, have U.L. approval.

The new series comes in single, dual, triple and quad sections, with leads all coming from one end or from opposite ends of the cartridge. A complete choice of voltage and capacity ratings is available.

For technical data, write or call Mallory today. A Mallory capacitor engineer will be glad to consult on your circuit requirements, to suggest possible cost-cutting simplifications based on Mallory's long experience in all types of applications for electronic components.




*Inside of the case is this foil-wrapped cartridge. Tabs are welded to the leads, to prevent intermittent connections.*

Parts distributors in all major cities stock
Mallory standard components for your convenience.

**Serving Industry with These Products:**

**Electromechanical**—Resistors • Switches • Television Tuners • Vibrators
**Electrochemical**—Capacitors • Rectifiers • Mercury Batteries
**Metallurgical**—Contacts • Special Metals and Ceramics • Welding Materials

P.R. MALLORY & CO. Inc.
MALLORY
P. R. MALLORY & CO., Inc., INDIANAPOLIS 6, INDIANA

Variable width — width of each of 5 pulses can be adjusted independently.

Variable delay—delay between each of 5 pulses can be adjusted independently.

CODE MODULATED MULTIPLE-PULSE MICROWAVE SIGNAL GENERATOR Model B

Pulse-time modulation—input provided in each of 5 pulse channels for external pulse-time modulation.

Variable repetition rate—repetition rate of each group of pulses can be varied.

# CODE MODULATED MULTIPLE-PULSE MICROWAVE SIGNAL GENERATOR

## Model B 950-10,750 mc

*Generates multi-pulse modulated carrier for beacons, missiles, radar... provides 5 independently adjustable pulse channels, 4 interchangeable r-f oscillator heads, precision oscilloscope, self-contained power supplies ... all in one integrated mobile instrument.*

The Polarad Model B is an essential instrument for testing beacons, missiles, radar, navigational systems such as DME, Tacan, H. F. Loran, etc., where multi-pulse modulated, microwave frequency energy with accurately controlled pulse width, delay, and repetition rate is required for coding.

**A fully integrated self-contained equipment with these features:**

**Four Interchangeable Microwave Oscillator Units** — all stored in the instrument ... each with UNI-DIAL control ... precision power monitor circuit to maintain 1 mw power output reference level ... keying circuit to assure rapid rise time of modulated r-f output ... non-contacting chokes.

**Five Independently Adjustable Pulse Channels** —each channel features variable pulse width and delay; has provisions for external pulse-time modulation.

**Precision Oscilloscope with Built-In Wide Band RF Detector** for viewing the modulation envelope and accurately calibrating the r-f pulse width, delay, and group repetition rate. Equipped with built-in calibration markers.

**Self-Contained Power Supplies**—Model B operates directly from an AC line through an internal voltage regulator. The coded multi-pulse generator is equipped with an electronically regulated low voltage DC supply. Klystron power unit adjusts to proper voltage automatically for each interchangeable band.

Contact your Polarad representative or write to the factory for detailed information.

## SPECIFICATIONS:

**Frequency Range:**
- Band 1 . . . 950 to 2400 mc
- Band 2 . . . 2150 to 4600 mc
- Band 3 . . . 4450 to 8000 mc
- Band 4 . . . 7850 to 10,750 mc

**Frequency Accuracy** . . . ±1%

**RF Power Output** . . . 1 milliwatt maximum (0 DBM)

**Attenuator:**
- Output Range . . . 0 to −127 DBM
- Output Accuracy . . . ±2db
- Output Impedance . . . 50 ohms nominal

**RF Pulse Characteristics:**
- a. Rise Time . . . Better than 0.1 microsecond as measured between 10 and 90% of maximum amplitude of the initial rise.
- b. Decay Time . . . Less than 0.1 microsecond as measured between 10 and 90% of maximum amplitude of the final decay.
- c. Overshoot . . . Less than 10% of maximum amplitude of the initial rise.

**Internal Pulse Modulation:**
- No. of Channels . . . 1 to 5 independently on or off
- Repetition Rate . . . 40 to 4000 pps
- Pulse Width . . . 0.2 to 2.0 microseconds
- Pulse Delay . . . 0 to 30 microseconds
- Accuracy of Pulse Setting . . . 0.1 microsecond
- Minimum Pulse Separation . . . 0.3 microsecond
- Initial Channel Delay . . . 2 microseconds from sync. pulse
- Internal Square Wave . . . 40-4000 pps (separate output)

**Pulse Time Modulation:**
- Frequency . . . 40-400 cps any or all channels
- Required Ext. Mod. . . . 1 volt rms min.
- Maximum deviation . . . ±0.5 microsecond

**Power Input** (built-in power supply) 105/125 v. 60 cps 1200 watts.

AVAILABLE ON EQUIPMENT LEASE PLAN

FIELD MAINTENANCE SERVICE AVAILABLE THROUGHOUT THE COUNTRY

**POLARAD** PROVEN RELIABILITY **ELECTRONICS CORPORATION** 43-20 34th STREET, LONG ISLAND CITY 1, N. Y.

REPRESENTATIVES: • Albuquerque • Atlanta • Baltimore • Bayonne • Bridgeport • Buffalo • Chicago • Dayton • Fort Worth • Los Angeles • New York
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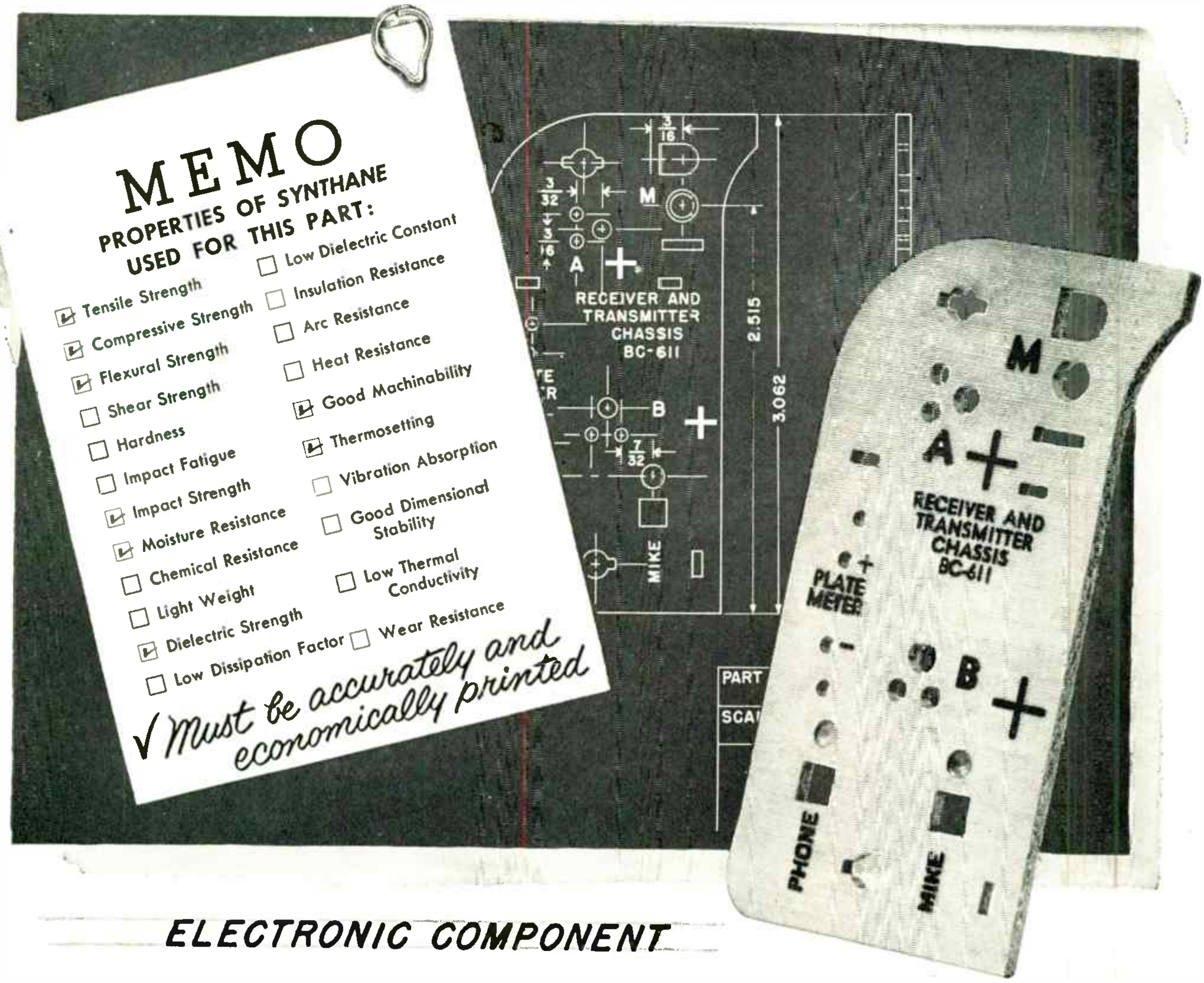
## YOU FURNISH THE PRINT, WE'LL FURNISH THE PART




## ELECTRONIC COMPONENT OF SYNTHANE HAS DURABILITY, DIMENSIONAL STABILITY, DIELECTRIC STRENGTH

Although this sturdy end plate will fit into the palm of your hand, it has in combination all the dielectric strength, the physical properties, and the printability the customer requires. It's made of *Synthane*, a laminated plastic, the same material used in hundreds of other electrical, mechanical, and chemical applications.

The blue print for this part calls for accurate machining, the punching of twenty holes of various shapes and sizes, and printing or engraving in three different colors. *Synthane* delivers finished parts exactly as specified, ready for the production line. The customer gets them promptly without problems of tooling up, waste, or rejects.

If you need components with many properties in combination, you will want to know more about *Synthane* laminates and the *Synthane* fabricating service. Send in the coupon for the full story.

**SYNTHANE CORPORATION,** 8 River Road, Oaks, Pa.

Please send my copy of the Synthane catalog.

Name__________

Title__________

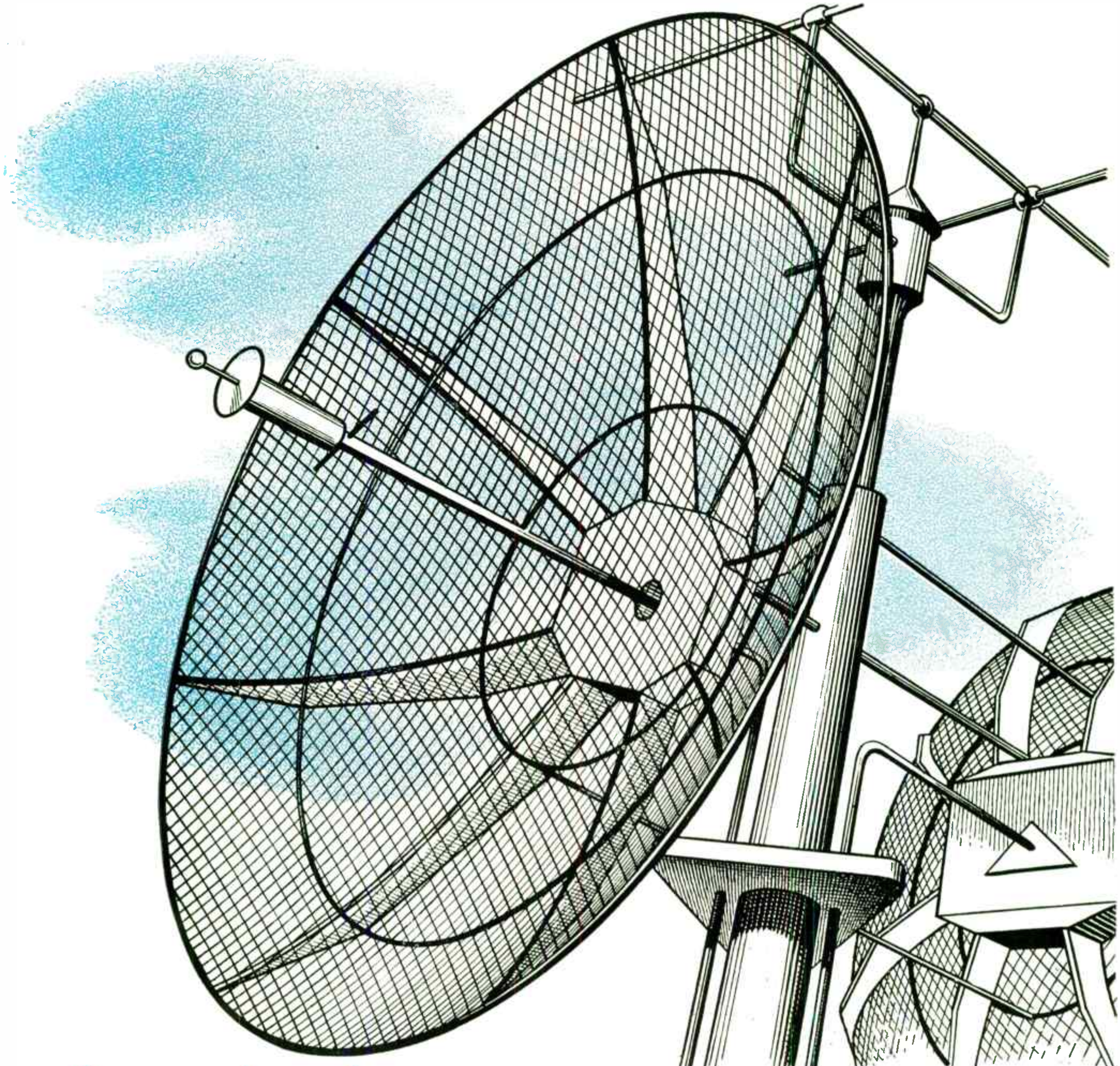
Company__________

Address__________

City______ Zone____ State____

**SYNTHANE** LAMINATED PLASTICS

**SYNTHANE CORPORATION • OAKS, PENNSYLVANIA**



## *In magnetrons, too*

# CRUCIBLE PERMANENT MAGNETS

## *provide maximum energy . . . minimum size*

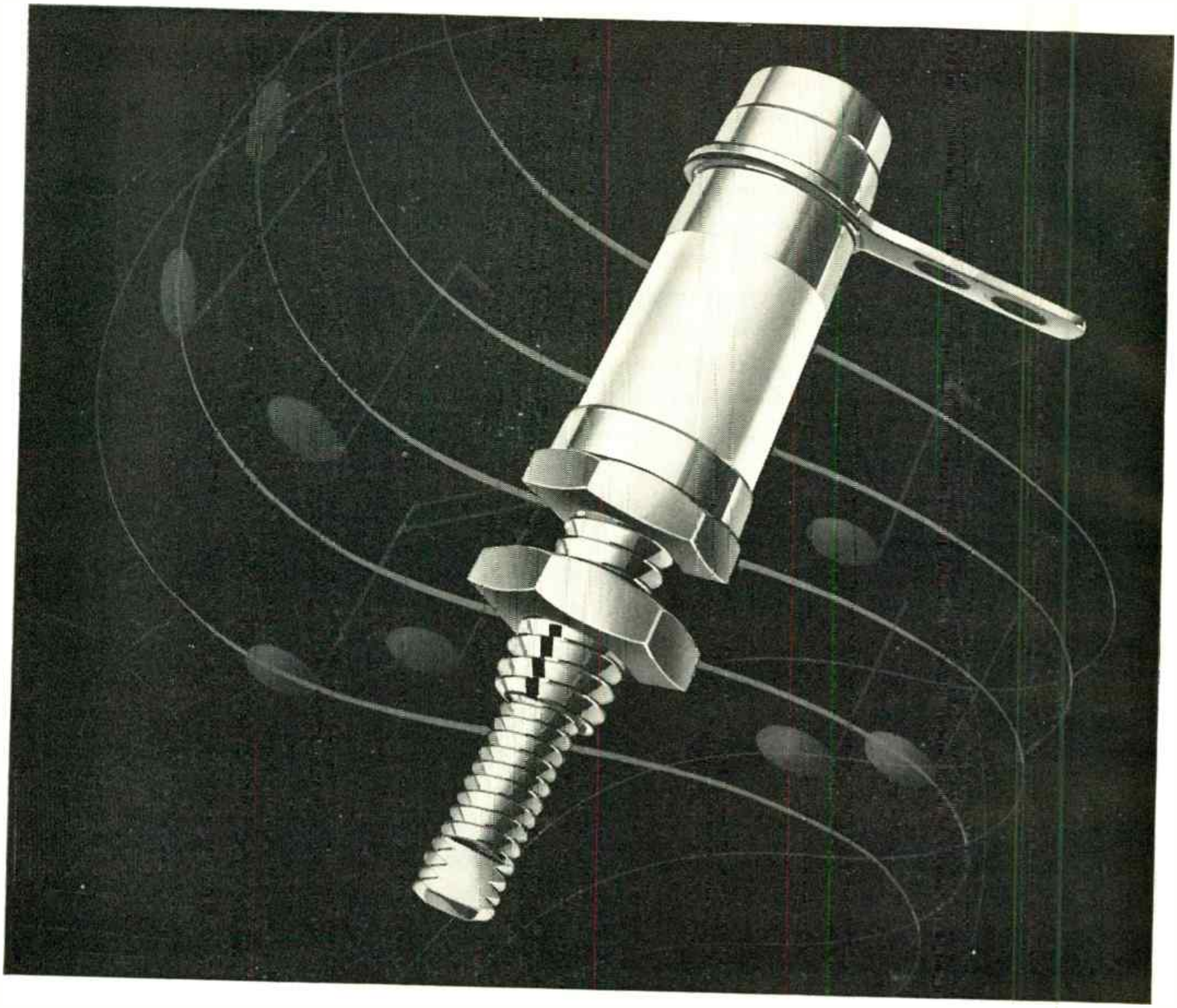
In designing magnetrons for radar systems — or any very high-frequency oscillator application — you can be sure of a *consistently higher* energy product with Crucible alnico magnets. *This means more power from a minimum size magnet!*

Crucible alnico permanent magnets are made to meet practically any size requirement from a fraction of an ounce to several hundred pounds. And they're sand cast, shell molded or investment cast to the exact size, shape, tolerance and finish you need.

Crucible has been one of the largest producers of permanent magnets since the development of alnico alloys. Its unsurpassed magnet experience is backed by over 50 years of fine steelmaking. That's why the best solution to magnet problems starts with a call to Crucible. *Crucible Steel Company of America, Henry W. Oliver Building, Pittsburgh 22, Pa.*

**CRUCIBLE** first name in special purpose steels

**Crucible Steel Company of America**



# Mighty midget "tunes up" for major performance

This miniaturized CST-50 variable ceramic capacitor outperforms capacitors several times larger. C.T.C.'s unique design includes a *tunable element* which virtually eliminates losses due to air dielectric. This results in wide minimum to maximum capacity range of 1.5 to 12 MMFD.

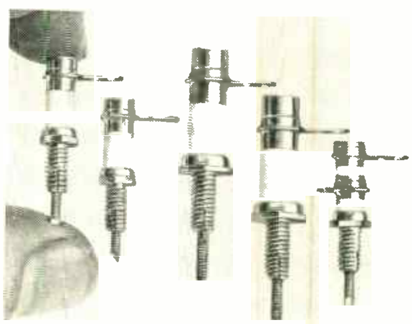
This tuning sleeve is at ground potential and can be locked firmly to eliminate undesirable capacity change. Each CST-50 is provided with a ring terminal with two soldering spaces.

This is *but one* of a versatile family of C.T.C. ceramic capacitors of this type, built to C.T.C.'s *quality control* production standards for guaranteed performance.

All C.T.C. components — standard or custom — are subject to this precision manufacture. Other C.T.C. components include coil forms, coils, terminal boards, terminals, diode clips, insulated terminals and hardware. C.T.C. engineers are glad to consult on *your* component problem. Write *now* for sample specifications and prices to Sales Engineering Department, Cambridge Thermionic Corporation, 456 Concord Ave., Cambridge 38, Mass. On West Coast, contact E. V. Roberts, 5068 West Washington Blvd., Los Angeles 16 or 988 Market St., San Francisco, Calif.

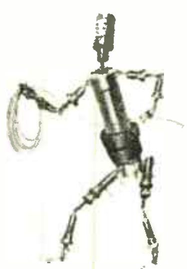
*C.T.C. Capacitor Data: Metallized ceramic forms.*

CST-50, in range 1.5 to 12.5 MMFDs.
CST-6, in range 0.5 to 4.5 MMFDs.
CS6-6, in range 1 to 8 MMFDs.
CS6-50, in range 3 to 25 MMFDs.
CST-50-D, a differential capacitor with the top half in range 1.5 to 10 MMFDs and lower half in range 5 to 10 MMFDs.



**CAMBRIDGE THERMIONIC CORPORATION**

*makers of guaranteed electronic components, custom or standard*



# All wound up

*trying to make your own R. F. Choke Coils?*

## *Standardize on* Jeffers R.F. Choke Coils-

*save time, labor and expense*

Now you can stock a wide range of R. F. choke coils just as you do resistors, capacitors and other similar components.

Jeffers Electronics is ready to deliver to you a complete line of R. F. choke coils with the widest range of inductance values available. No longer do you have to waste time, labor and money on slow, tedious hand assembly from miscellaneous forms, wires and coatings.

Instead you receive a standardized product from Jeffers, completely assembled and ready for use. Coils that are well made, too. Insulated copper wire instead of bare wire for windings . . . husky, molded jackets instead of those fastened by glue. All windings are soldered to leads . . . shorted end-turns completely eliminated.

Why not give Jeffers R. F. choke coils a try on your next order? *Write today for our specification sheets.*

### Other Jeffers Products

ceramic capacitors • disc capacitors
high voltage condensers • capristors

### Other Speer Products for the Electronics Industry

anodes • contacts • resistors • iron cores
discs • brushes • molded notched* coil forms
battery carbon • graphite plates and rods

*Patented

JEFFERS ELECTRONICS

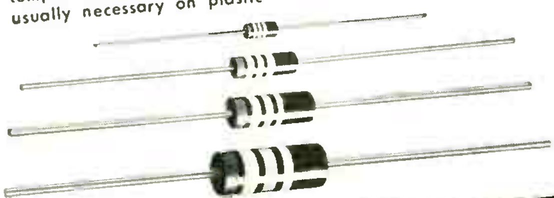
Jeffers Electronics Division
Speer Carbon Co.
Du Bois, Pennsylvania

*Other Divisions:* Speer Resistor, International Graphite & Electrode

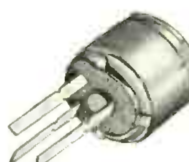
# ALLEN-BRADLEY QUALITY COMPONENTS for PRINTED CIRCUITS

## FIXED MOLDED RESISTORS

Stable characteristics, uniform size, accurate lead wire dimensions, special uniform solder coating of leads, and mechanical strength are essential requirements for components used for printed circuit applications. It is on these features that the Quality reputation of Allen-Bradley molded resistors was established. Made in 1/10, ½, 1, and 2 watt sizes, meeting all applicable RETMA and MIL-R-11A specifications, and rated at 70C ambient temperature, these resistors do not require the derating usually necessary on plastic board assemblies.



## SMALL POTENTIOMETERS



Type GP

Type T

Allen-Bradley Type GP and Type T adjustable molded resistors are designed for printed circuits where quality—not price—is the important requirement.

They are available in resistance values from 100 ohms to 5 megohms in standard tapers.

Both are rated at ½ watt.

The Type GP potentiometer is ½ inch in diameter and is supplied with gold-plated lugs.

The Type T potentiometer is 1 inch in diameter and has round, solder-coated leads.

## CERAMIC CAPACITORS

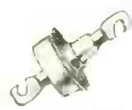
Type DS

Type CE and Type DS ceramic capacitors are available for printed circuit assemblies in all commercial characteristics. These units were developed for operation at above normal ambients and are impervious to moisture.

Type DS resin-enclosed capacitors have a new lug adapted for insertion in a round hole. The lug is so designed as to properly space the capacitor above the printed board.

Type CE

## OTHER MINIATURE COMPONENTS



Type FT

Type FC

Type SO

Type FT feed-thru and Type SO stand-off discoidal capacitors incorporate the maximum performance characteristics together with extreme ruggedness and small size. They are available in standard nominal values from 5 to 1,000 mmf.

Allen-Bradley Type FC Ferri-Cap feed-thru filter is a new, unique device consisting of a discoidal feed-thru capacitor combined with ferrite material to provide internal impedances in series with both ends of the feed-thru electrode.

Investigate these truly superior miniature components.

## COMPONENT PACKAGING


Carton packaging (Standard)

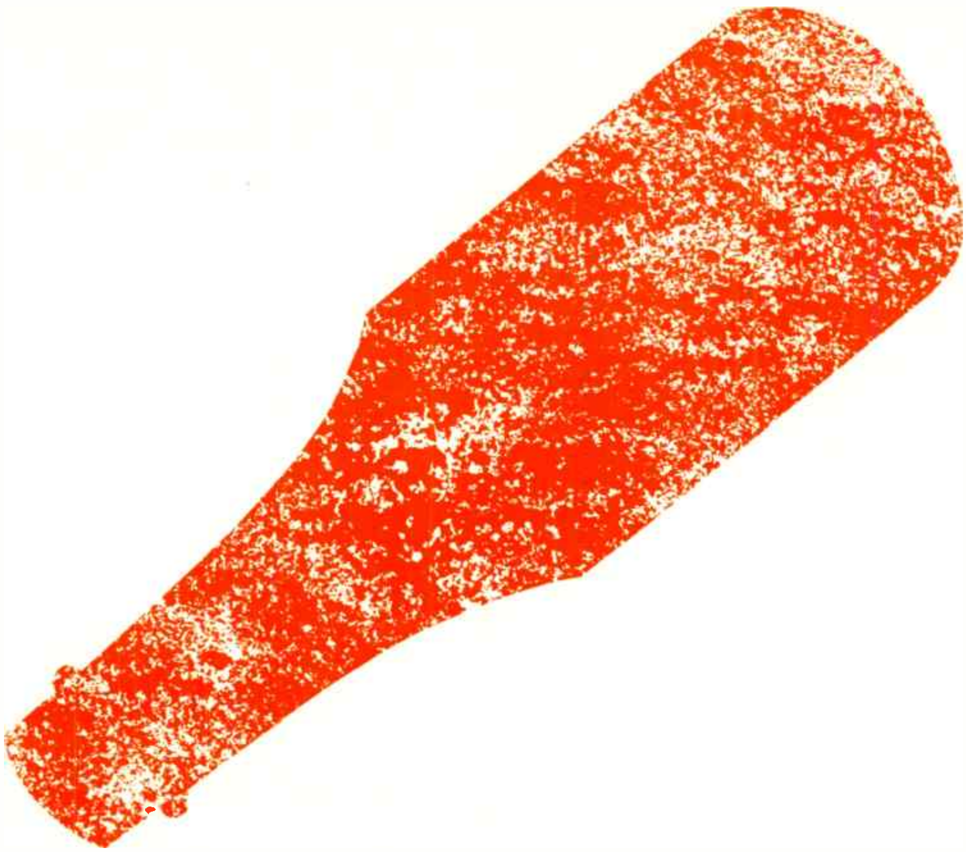
To keep step with the automatic trends of industry, small components, such as the molded fixed resistors and the ceramic capacitors, are now offered by Allen-Bradley in reel packages wherein the components are attached to a pressure-sensitive tape, ideally adapted for automatic assembly or preassembly operations.

Reel packaging (Special)

Allen-Bradley Co.
114 W. Greenfield Ave.
Milwaukee 4, Wis.

In Canada
Allen-Bradley Canada, Ltd.
Galt, Ont.

**ALLEN-BRADLEY**
**RADIO & TELEVISION COMPONENTS**
QUALITY



# no ketchup needed

When a predestined steer meets a dedicated chef ... man, that's steak!

If steer or chef is bad (and double trouble if both) ... ketchup can't help.

Now comes the commercial.

Take the best available materials (sifted by unrelenting research). Season with the same inventiveness used by Dr. A. O. Beckman to develop the precision potentiometer in 1940 (we've never switched brands). Add assembly-line economies without compromising quality. Test and retest in the industry's most complete lab. Inspect a dozen times (too many cooks can't spoil *this* broth). Pack well. Ship on schedule.

Man, that's a HELIPOT precision potentiometer ... no ketchup needed!

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*first in precision potentiometers*
*Helipot Corporation / South Pasadena, California*
*Engineering representatives in principal cities*
*a division of* BECKMAN INSTRUMENTS, INC.

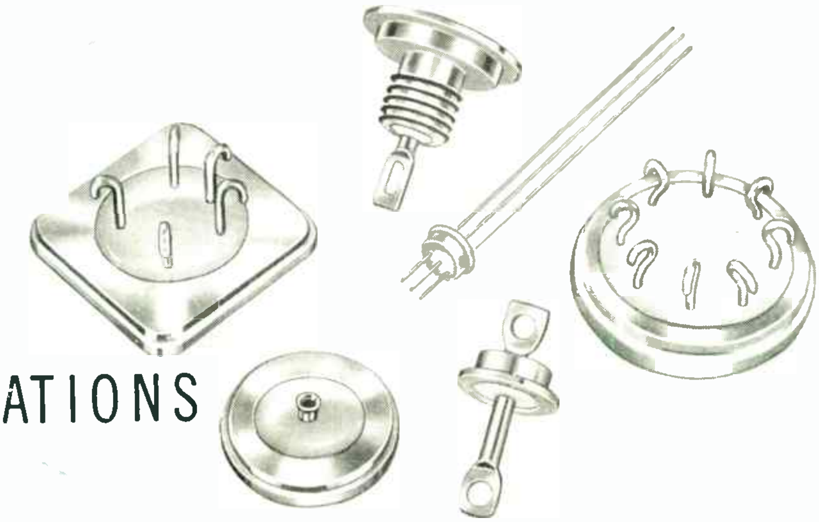
BECKMAN

*Helipot makes precision potentiometers ... linear and non-linear ... in the widest choice of sizes, mounting styles and resistances. Many models are stocked for immediate shipment ... our engineers will gladly adapt standard models to your requirements ... or design entirely new* HELIPOT* *precision potentiometers for you.*

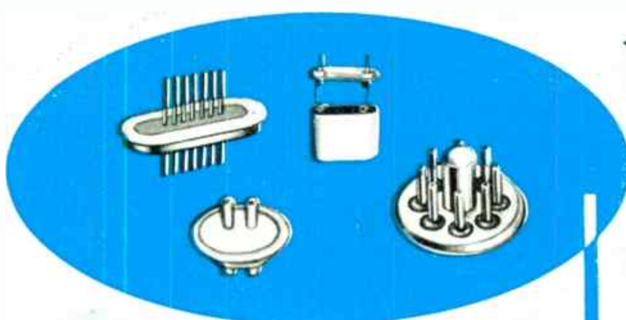
*For information and specifications ... write for data file 702.*

536 *REG. U.S. PAT. OFF.

WHEN SPECIFICATIONS CALL FOR



# GLASS-TO-METAL SEALS



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Ask E-I hermetic seal specialists for a quick, economical solution to your design problems involving glass-to-metal seals. E-I specialization and standard designing means your specifications can be fulfilled, in most cases, by low cost catalog items. E-I offers fast delivery in reasonable quantities on seals developed for practically every type of electronic and electrical termination. Call, write or wire E-I, today!

**COMPRESSION SEALS**
**MULTIPLE HEADERS**
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— offering 8 important advantages including cushioned glass construction, design standardization, high dielectric strength, miniaturization, vacuum tight sealing, vibration resistant, super durability, maximum rigidity, etc.

One **dependable** source for **all** hermetically sealed terminal requirements!

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# INCREASE CIRCUIT RELIABILITY WITH TI-RADELL deposited carbon RESISTORS



**newest line of precision components from Texas Instruments**

For precise resistance values under extreme operating conditions, design with RADELL deposited carbon resistors — now manufactured by Texas Instruments. With resistance tolerance held to ±1%, Texas Instruments RADELL resistors provide exceptional stability plus a wide range of resistance values. Like all TI components, they are manufactured to exacting instrument standards.

Texas Instruments RADELL resistors are mass-produced in three lines and in ½, 1, and 2 watt sizes. Resistance values range from 25 ohms to 30 megohms.

**WRITE** for Bulletin No. DL-C 539 giving detailed specifications of all three lines of Texas Instruments RADELL resistors. Your best source for precision components, TI also manufactures a complete line of subminiature transformers as well as custom capacitors, delay lines, special transformers and other reliable electronic components.

**Hermetically sealed line** — designed for extreme conditions of moisture and temperature. Specially treated ceramic shell effectively seals out moisture and air, resists abusive handling, and assures complete insulation.



**MIL-Line** — designed for the broad field of military applications. Exclusive multi-layer coating provides environmental protection substantially equal to hermetic sealing throughout low and middle ranges of resistance. MIL-Line resistors more than meet MIL-R-10509A specifications.

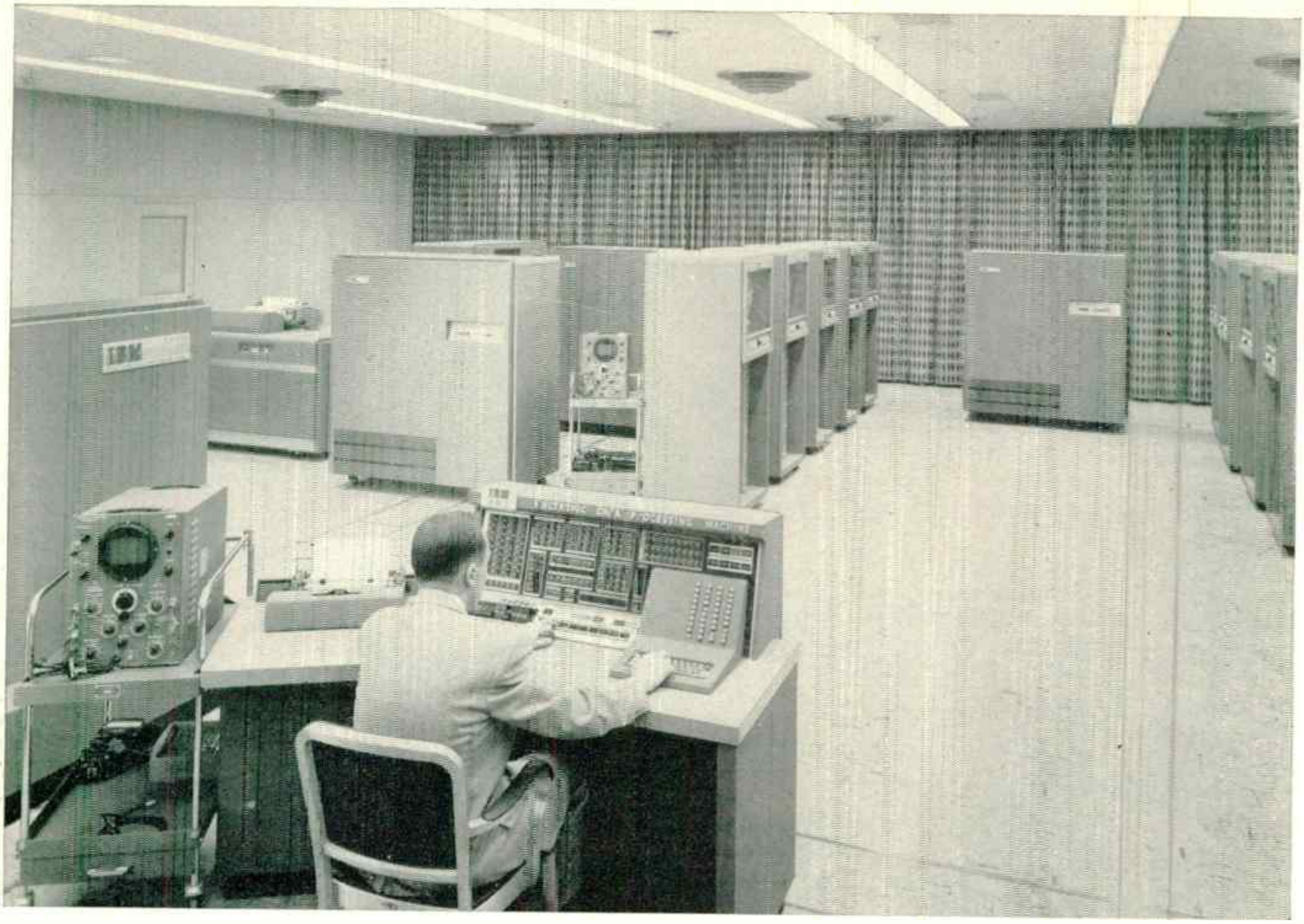


**Industrial-Line** — differs from MIL-Line series only in type of coating. Industrial line resistors provide close tolerances for military, instrument and industrial applications where less extreme humidity conditions are encountered. Typical applications include computers, test equipment, communication and control systems.

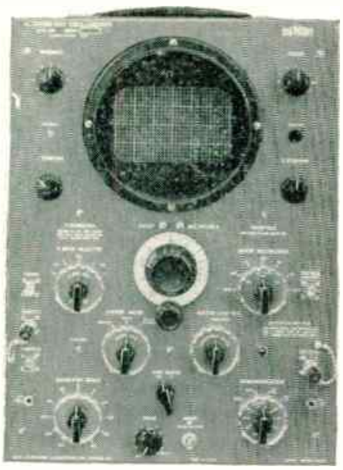
RESISTORS SHOWN ACTUAL SIZE

TWENTY-FIFTH YEAR OF PLANNED GROWTH

TEXAS INSTRUMENTS
INCORPORATED
6000 LEMMON AVENUE DALLAS 9, TEXAS



# IBM selects DU MONT TYPE 329* as test oscillograph for their new type 702 computer



When IBM Corporation, world's largest manufacturer of computer equipment, produced their new Model 702, an essential phase of the project involved selection of a cathode-ray oscillograph to go into the field with each computer as standard test equipment. Requirements were strict.

IBM's approach to the problem was to conduct side-by-side evaluation with other competitive instruments. On the basis of actual performance, they selected the Du Mont Type 329 as their test oscillograph.

What are some of the primary reasons why IBM decided on the Du Mont Type 329? Excellent sensitivity—either d.c. or a.c. coupled. Precisely calibrated sweeps with movable notch magnification—ideal for making accurate measurements. Brightness—adequate for display of very fast pulses. Synchronization simplicity—the Type 329 "locks in" on almost any type of signal. Stability—the trace remains steady as a rock despite power line fluctuations, etc. Reliability in service—calibration adjustment requires no extra test gear and is a simple one-step process. And virtually any tube may be replaced without special selection.

Another factor contributing to the selection of the Type 329 was the well known Du Mont Field Service Organization, which assures that regardless of where in the United States the equipment is used, swift, competent service facilities are in the immediate vicinity.

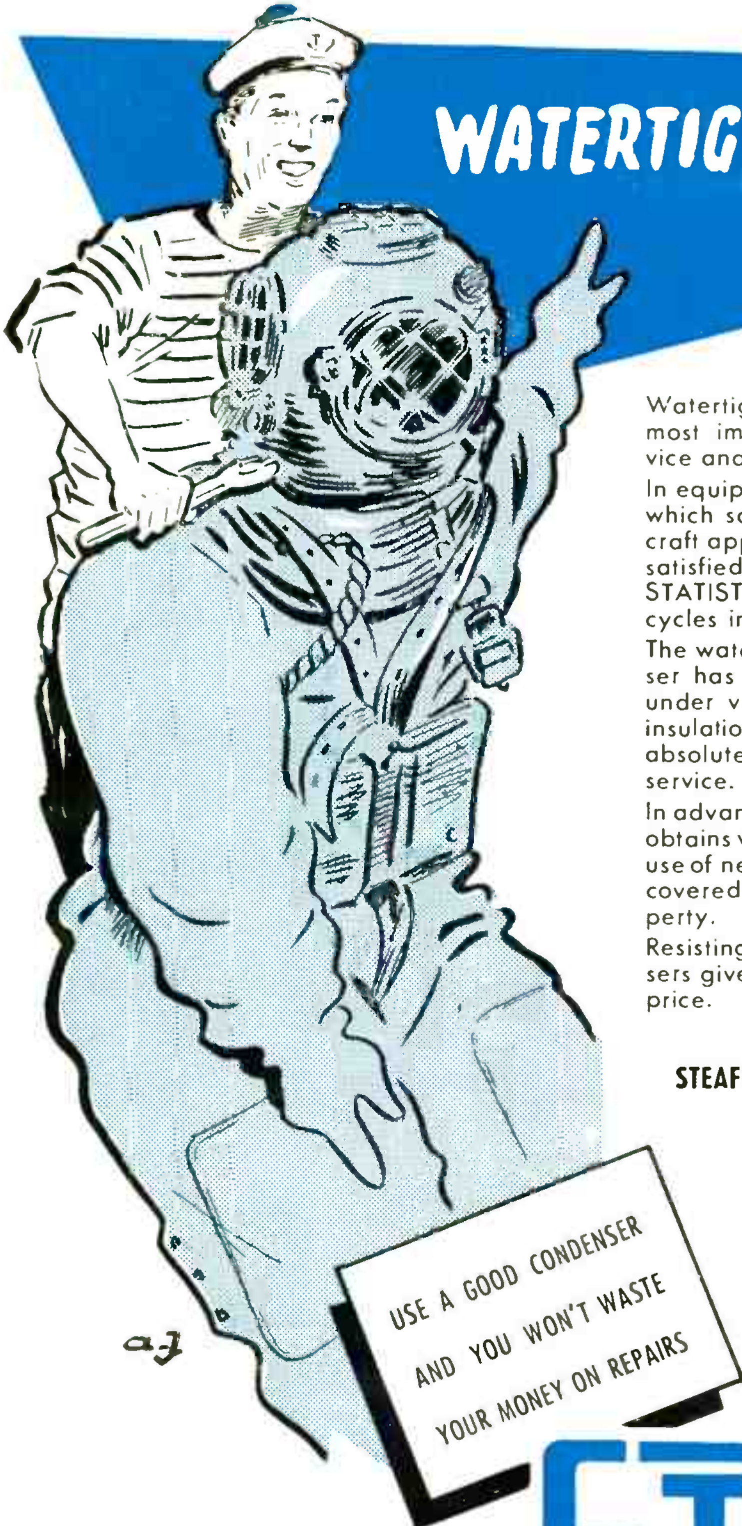
If you have instrumentation requirements, Du Mont facilities are always available for discussion and recommendations. Write us today for complete information on the Type 329, or on any problem you may have relating to cathode-ray instrumentation.

*"Modified slightly for IBM's application."

**DU MONT**

ALLEN B. DU MONT LABORATORIES, INC. • TECHNICAL SALES DEPARTMENT
760 Bloomfield Avenue, Clifton, New Jersey

# WATERTIGHTNESS




Watertightness for a condenser is one of the most important prerequisites for good service and long life.

In equipment intented for long service and on which safety of the users depends (as in aircraft applications for instance) one cannot be satisfied with NEARLY watertight condensers, STATISTICALLY tested by a couple of thermal cycles in moist atmosphere.

The watertightness of EACH STEAFIX condenser has been tested by immersion in water under vacuum, followed by measurement of insulation resistance, thus giving to the user an absolute guarantee of long life and good service.

In advance of all standards now in use STEAFIX obtains watertightness of its condensers by the use of new molding materials and of techniques covered by patents which are its exclusive property.

Resisting time and weather, STEAFIX condensers give you perfect service for the minimum price.

**STEAFIX capacitors resist time and weather**


# STÉAFIX

Société Anonyme au capital de 65.000.000 de Francs

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**TÉL. MONtmartre 02-93**

*the european specialist of the mica capacitors*

PUBLICITÉ SOFEL PARIS

# Tests show two CATHALOYS most versatile cathode materials

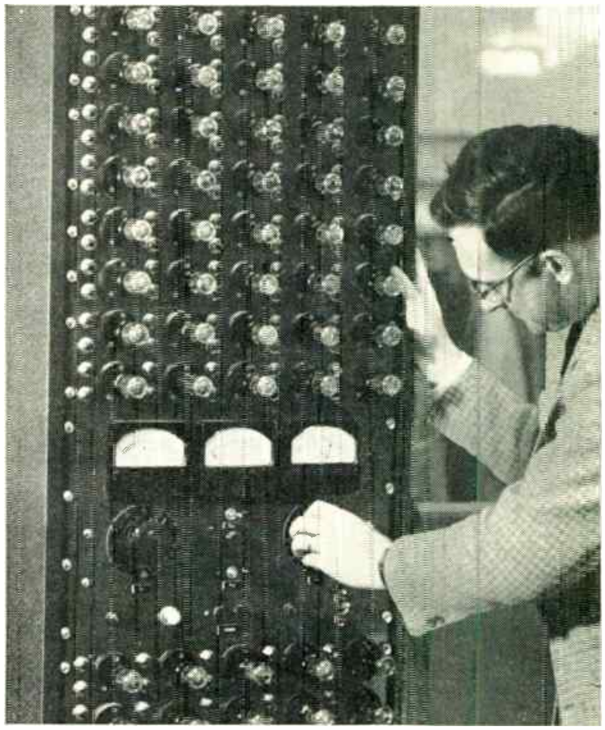
## New alloys from Superior Tube simplify selection, prolong tube life

Now the engineer's job of selecting the right cathode alloy for practically any electron tube can be a simple choice between two new CATHALOYS from Superior Tube.

**CATHALOY A-32*** is an active alloy characterized by rapid activation, high emission level throughout life, absence of interface impedance, and very low sublimation. These remarkable advantages are the result of using aluminum in place of silicon or magnesium as the reducing agent. The addition of a small percentage of tungsten also makes A-32 approximately 50% more shock resistant than cathodes without tungsten. Thus A-32 is suitable for virtually any active alloy application, including ruggedized tubes.

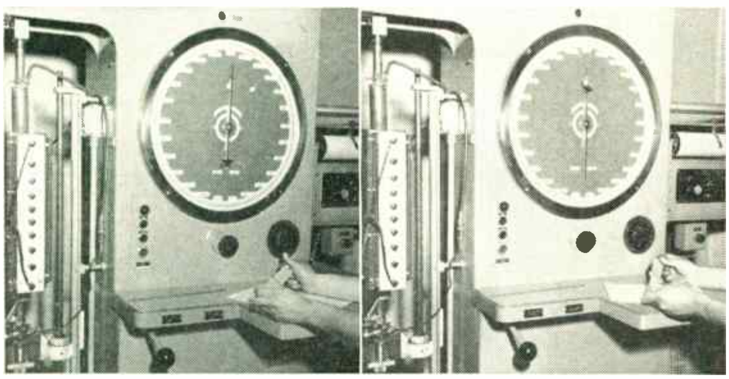
**CATHALOY P-50** is a passive alloy of carefully controlled analysis that is commercially available in WELDRAWN† cathodes as well as Lockseam.‡ It can be made in WELDRAWN form because of its capacity to take much more severe reductions in cold drawing without rupture than other grades of passive alloys. P-50 is identical in composition with the well-known ASTM Grade 21. The important difference is in the method of melting which improves the uniformity and completeness with which deoxidation is accomplished. All heats are tested in Superior Tube's laboratory before being approved for production.

Ask for complete technical reports on both these new CATHALOYS. Write Superior Tube Co., 2506 Germantown Ave., Norristown, Pa.



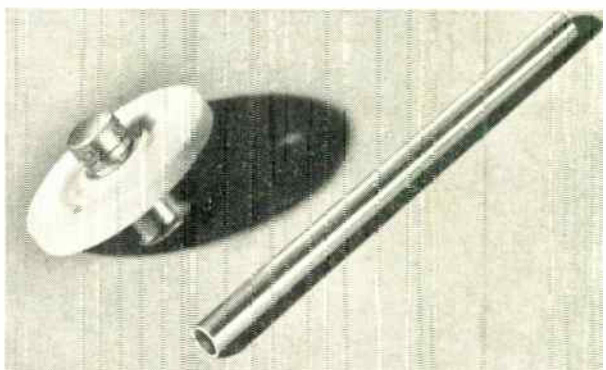
ON TEST. Laboratory photo of test diodes used in Superior Tube's electronic laboratory. Under exhaustive tests, the new Cathaloys display performance characteristics not present in other alloys.

*Patent applied for    †TM Reg. U.S. Pat. Off., Superior Tube    ‡Manufactured under U.S. patents    NOTE. Cathaloy is a trademark of Superior Tube Co., Reg. U.S. Pat. Off.



CATHALOY A-32—3750 psi    Tungsten-free cathode alloy—2500 psi

**50% STRONGER.** High temperature tensile testing machine proves Cathaloy A-32 approximately half again stronger than tungsten-free cathode alloys.



**WELDRAWN PASSIVE ALLOY.** Typical uses for Cathaloy P-50 are in Weldrawn cathode sleeve shown at right and in disc cathode shank at left. Heretofore, passive alloys have not been commercially available for these applications.

**Superior Tube**

**The big name in small tubing**

All analyses .010" to 5/8" OD. Certain analyses in light walls up to 2½" OD.

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**because . . .** AlSiMag is manufactured under strict quality control from the best materials available.

**because . . .** Precision tolerances and superior workmanship are matters of routine with AlSiMag.

**because . . .** Thoroughly trained Research and Engineering Staffs are available, without charge, to help with customer's design and application problems.

**because . . .** AlSiMag represents the net result of over 53 years of specialized experience in the technical ceramics field.

**because . . .** With AlSiMag, you can be sure of receiving as many parts as you want —when you want them. Four large plants house an abundance of the most up-to-date mixing equipment, high-capacity presses, die and tool shops, huge kilns, skilled manpower.

Your own experience will add many more reasons. specify AlSiMag on your next order for precision pressed or extruded ceramics . . . see how easy your buying job can become!

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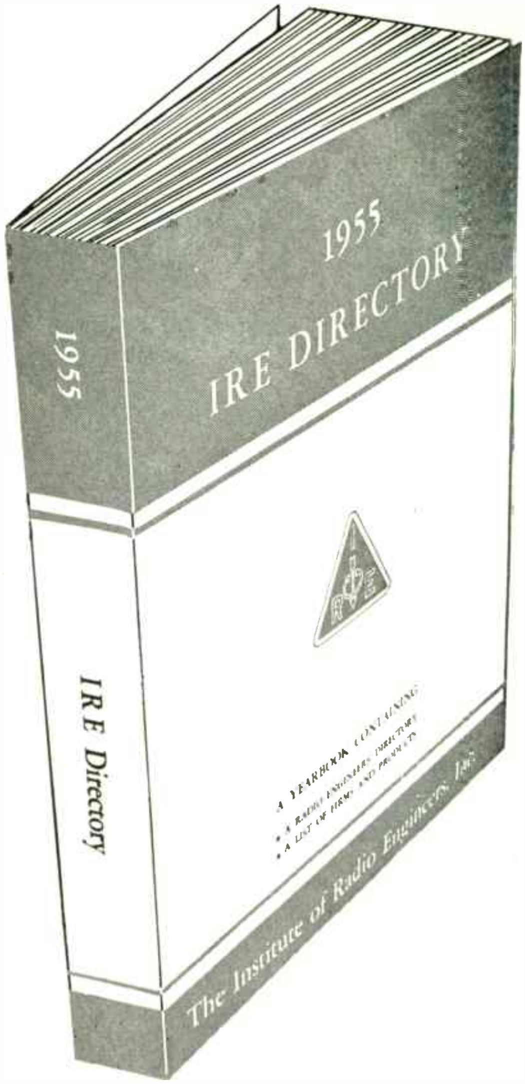
# The most valuable reference work in the world




- for the electronics engineer
- for buyers of component parts
- for users of electronic equipment
- for anyone who must

# FIND FACTS FAST

The IRE Directory contains full information on 3000 firms manufacturing products or furnishing services in the radio-electronic field, brought up-to-date every year.

Arranged efficiently and logically, the way an engineer thinks, for simple location of any product or service—

675 specific products and services, arranged under
99 major headings in
4 great groups

- Communications
- Components
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NEW.... SIGNIFICANT....

# Uniline®

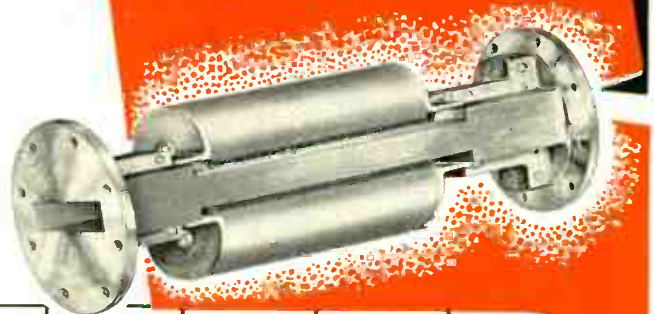
# S band SERIES

## OF MICROWAVE LOAD ISOLATORS

The important load isolation functions of Unilines are now available to designers and users of microwave systems and test equipment operating in the S-Band.

New, inherently rugged mechanical designs permit safe operation at substantial peak and average powers. Cascade Research has designed these units to optimize isolation-to-insertion loss-ratio. Special techniques have made possible a reduction in size and weight of the integral permanent magnets. As in all Unilines, no external power source is required.

These Unilines can also be furnished as an integrated part of a microwave circuit and may include such elements as directional couplers, hybrid junctions, twists, bends, mixers and crystal holders. Unilines with even greater power ratings are now under development.



| MODEL | 28-32A | 28-32B | 28-32C | 28-32D | SL-132 | SL-133 | H-28-32 |
|---|---|---|---|---|---|---|---|
| FREQUENCY RANGE | 2.8-3.2 KMC | 2.8-3.2 KMC | 2.8-3.2 KMC | 2.8-3.2 KMC | 2.0-2.4 KMC | 2.0-2.4 KMC | 2.8-3.2 KMC |
| FORWARD ATTENUATION (TYPICAL) | 1.6 DB | 1.2 DB | 0.8 DB | 0.4 DB | 1.0 DB | 1.0 DB | 0.4 DB |
| REVERSE ATTENUATION (TYPICAL) | 27 DB | 20 DB | 13 DB | 6 DB | 20 DB | 20 DB | 10 DB |
| V.S.W.R. (TYPICAL) | 1.2 | 1.2 | 1.2 | 1.2 | 1.2 | 1.2 | 1.15 |
| AVER. TRANSMITTER POWER INTO V.S.W.R. OF 2:1 | 50 WATTS | 50 WATTS | 50 WATTS | 50 WATTS | 50 WATTS | 50 WATTS | 150 WATTS |
| PEAK POWER INTO V.S.W.R. OF 2:1 | 150 KW | 150 KW | 150 KW | 150 KW | 200 KW | 200 KW | 150 KW |
| WAVEGUIDE SIZE | RG-48/U | RG-48/U | RG-48/U | RG-48/U | | RG-105/U | RG-48/U |
| WAVEGUIDE FLANGES | UG-53/U | UG-53/U | UG-53/U | UG-53/U | COAX TYPE N OUTPUT | UG-437/AU | UG-53/U |

## APPLICATIONS

LOAD ISOLATION: Will provide substantial isolation between source and load with negligible loss in transmitted power.

ELIMINATES LONG LINE EFFECTS present where antennas are located remotely.

REDUCES PHASE DISTORTION by eliminating multiple reflections.

PROVIDES SMOOTHER TUNING, greatly reduces mismatch presented to a magnetron by the antenna, eliminates tendency of magnetrons to lock on some frequencies or fail to operate on others.

SMOOTHS OUT POWER VARIATIONS WITH TUNING: Magnetron power output to Uniline is maintained at maximum despite changes in line parameters due to changes in frequency.

KLYSTRON, OSCILLATOR-AMPLIFIER BUFFER: Addition of Uniline between oscillator and amplifier can markedly reduce possible pulling and power inconstancy.

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## IRE People

**Richard P. Gaunt** (M'53) has joined the Lockheed Missile Systems Division as a member of the research staff. Mr. Gaunt has been associated for many years with the CalTech Jet Propulsion Laboratory where he was a Senior Research Engineer in charge of a missile systems group.

R. P. GAUNT

He has worked in the design and development of the guidance system for the Corporal missile concentrating in the design of special-purpose analog computers and differential analyzers.

Mr. Gaunt was born in Mount Vernon, N. Y. and received the B.S. in Physics from Brown University in 1948. His hobby is music and at the present time he is studying modern jazz piano. He is a member of the Beta Theta Pi fraternity, the IRE and Sigma Xi.

❖

Stuart L. Bailey, President of Jansky & Bailey, Inc., Washington, D. C. recently announced the election of **Delmer C. Ports** (A'38-SM'45), Chief Engineer, to the position of Vice-President.

D. C. PORTS

Mr. Ports received the B.S. degree in Electrical Engineering at the George Washington University and the M.S. degree from Ohio State University.

As an engineer for Jansky & Bailey, which he joined in 1936, Mr. Ports was first engaged in antenna design and installation projects and in wave propagation studies. He also worked in the development of high frequency measuring techniques and equipment.

With the advent of World War II, he took part in a number of research projects, sponsored by the Office of Scientific Research and Development, related to antenna characteristics, propagation phenomena and communication systems. He was in charge of programs to develop techniques and measure characteristics of high frequency antennas and supervised research which experimentally isolated some of the factors affecting propagation in the high frequency and very high frequency regions.

During this period Jansky & Bailey established a laboratory to measure communication systems in use by the military. Mr. Ports assumed the responsibility for this work.

He is a member of the American Insti-

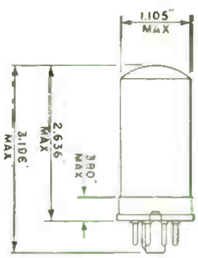
*(Continued on page 52A)*

Does a

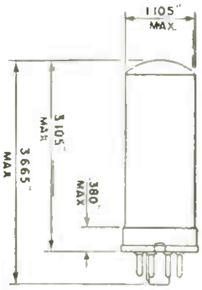
# Billion Operations

## —with no maintenance—

## interest YOU?



Type HG Relay



Type HGP Relay

### New CLARE Mercury-Wetted Contact Relays meet ALL requirements of today's high-speed switching devices

• If you design high speed switching machines or devices which demand accuracy and dependability of the highest order, you should know ALL about the new CLARE Mercury-Wetted Contact Relays.

CLARE Type HG and HGP Relays offer a combination of high speed, high current- and voltage-handling capacity, and extraordinary uniformity of performance over very long periods.

The relays consist of a magnetic switch, hermetically sealed in a high pressure hydrogen atmosphere in a glass capsule, and a coil, enclosed in a steel vacuum-tube-type envelope which has a standard medium-sized octal base. Platinum contact surfaces are continually wetted with mercury by means of a capillary connection to a mercury reservoir below the contacts. Type HGP Relays can be factory-adjusted to provide either biased or polarized operating characteristics.

*For complete information on the new CLARE Type HG and HGP Mercury-Wetted Contact Relays, contact your nearest CLARE representative or address C. P. Clare & Co., 3101 Pratt Blvd., Chicago 45, Illinois.*

**Send for CLARE Sales Engineering Bulletin No. 120**

**CLARE RELAYS**

**FIRST in the industrial field**

• Cutaway view of HG Relay showing how wick action keeps the mercury at the contacting surfaces continuously replenished.

**CLARE Type HG and HGP Relays are built to meet the exacting requirements for—**

| | |
|---|---|
| Computing machines | Servo-mechanisms |
| Sorting machines | Relay amplifiers |
| Tabulating machines | High-speed keying relays |
| Totalizers | Signaling devices |

All kinds of high-speed switching devices

**Outstanding features of CLARE Type HG and HGP Relays**

## ELECTRICAL FEATURES

**LONG LIFE:** Conservative life expectancy of over a billion operations when operated within ratings.

**HIGH SPEED:** Give consistent performance at speeds up to 60 operations per second.

**HIGH CURRENT**—and voltage-handling capacity (up to 5 amperes, and up to 500 volts).

**UNIFORMITY:** Operating time varies by only about 0.1 millisecond under constant drive conditions.

**NO CONTACT BOUNCE**

## MECHANICAL FEATURES

- Small chassis space required
- Convenient plug-in mounting
- Environment-free
- Tamperproof
- High sensitivity
- Maintenance-free
- No contact wear
- Adjustment cannot change

## IRE People

*(Continued from page 50A)*

tute of Electrical Engineers, and an associate member of the Acoustical Society of America. He serves on a number of technical and professional committees including the Committee on Antennas and Waveguides, Standards Committee, Joint IRE-AIEE Committee on High Frequency Measurements, Commission Six of International Scientific Radio Union, and Administrative Committee of IRE Professional Group on Antennas and Propagation. Mr. Ports is a member of Sigma Xi, Eta Kappa Nu, and Sigma Tau.

❖

**F. Stanley Atchison** (A'53–SM'53) has been appointed Technical Director of the U. S. Naval Ordnance Laboratory, Corona, California. He succeeds Ralph A. Lamm who is joining the staff of Bendix Aviation Corporation.

F. S. Atchison

Dr. Atchison was born in Missouri and received the Ph.D. in physics from the State University of Iowa in 1942. Joining the staff of the National Bureau of Standards in 1942 as a physicist in the Ordnance Development Division, he worked on proximity fuzes and went to the Mariana Islands as technical advisor to the Army Air Force during their first use of proximity fuzed bombs at the invasion of Iwo Jima. Dr. Atchison was subsequently put in charge of the Missile Intelligence Section, where he was concerned with the design of electronic systems for guided missile control. He moved to California in 1951 when the National Bureau of Standards established its Corona Laboratories, and directed a special research project on developments for electronic computers. Last year he was appointed head of the Physical Science Department, which is engaged in research in physics, chemistry, and electronics.

Dr. Atchison is president of the Sigma Xi Club at the University of California at Riverside and a member of the American Physical Society.

❖

The TelAutograph Corporation has recently appointed **R. G. Leitner** (SM'53) to the post of Chief Engineer. In his new position, Leitner will direct the development of TelAutograph communication systems and also the company's expanded program into automation, nucleonics, and electronic instrumentation.

Mr. Leitner formerly served as Chief Research Engineer for Packard Bell Company, and prior to that time in a similar

*(Continued on page 54A)*

BENEFITS NEVER BEFORE POSSIBLE

- no annoying ear-popping
- simple controls — automatic, reliable
- far less maintenance
  No sensor contacts or filters to clean.
- completely electrical
  No complex tubing.
- lower cost
- lighter weight
  As much as 60% lighter!
- easily retrofitted in any existing aircraft
- particularly adaptable to future high performance air transports

**STANDARD...**

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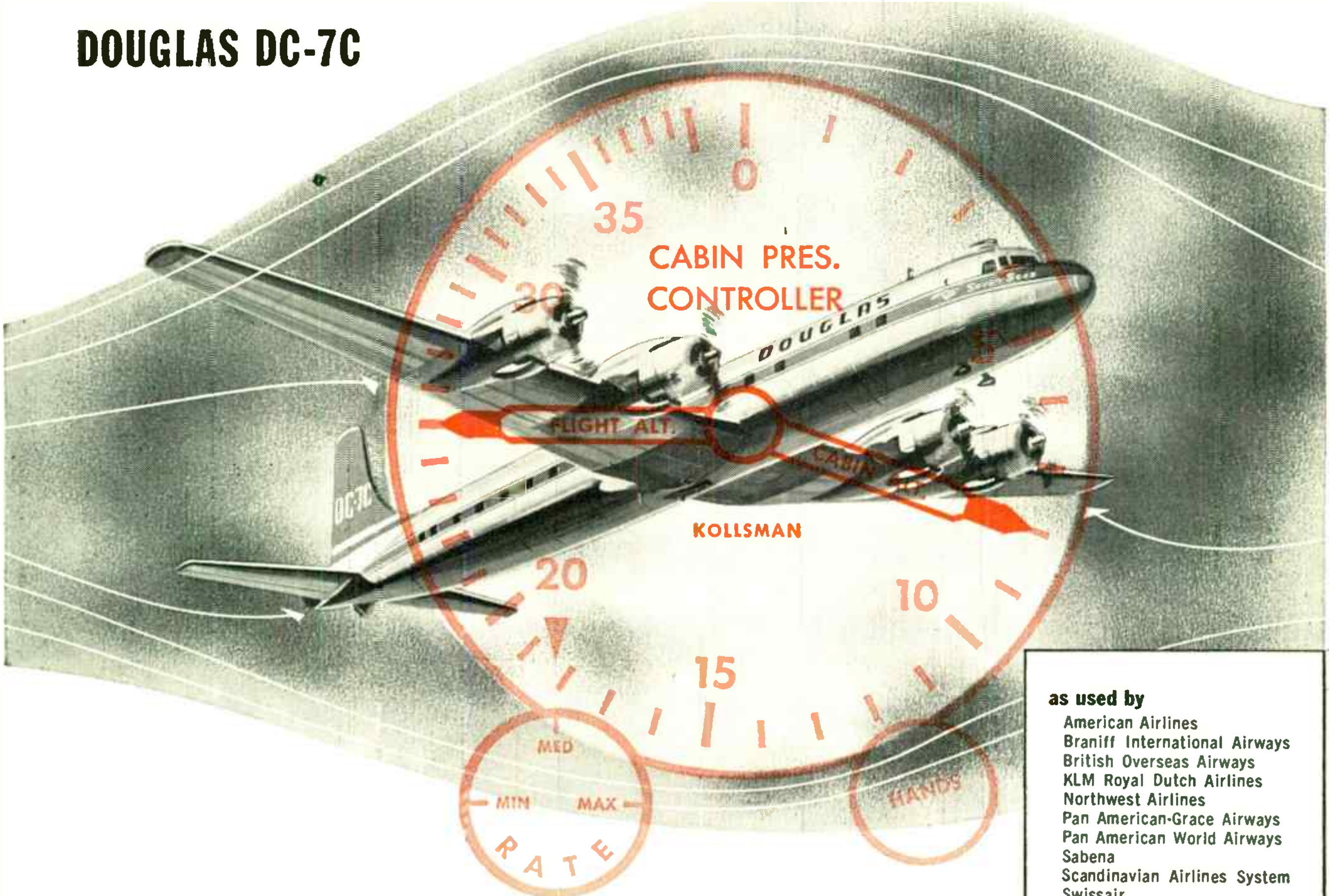
# DOUGLAS DC-7C

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KLM Royal Dutch Airlines
Northwest Airlines
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Pan American World Airways
Sabena
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## new KOLLSMAN KS-54 cabin pressure control system

DOUGLAS chose the new Kollsman KS-54 Cabin Pressure Control System for their new DC-7C's because of the many decided advantages it offers over the other existing systems.

**LIVING-ROOM COMFORT IN THE CABIN** . . . There is no annoying ear-popping because cabin pressure is held practically constant under cruising conditions. Even when cabin pressure is changing, the rate of change is so smoothly controlled that the actual change of pressure is unnoticable.

**PEACE OF MIND IN THE COCKPIT** . . . When the controls are set, the system is fully automatic and thoroughly reliable — especially so because of the simplicity of the Kollsman design.

**NO WORRY IN THE MAINTENANCE SHOP** . . . The components are simple and rugged, proven dependable and require a minimum of maintenance. There are no sensor contacts or filters to clean, no complex tubing to worry about.

**WRITE** for special folder giving full technical details on the new Kollsman KS-54 System, or ask to have a sales engineer visit you.

**kollsman** INSTRUMENT CORPORATION

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## IRE People

*(Continued from page 52A)*

position at Lear, Incorporated. He was chairman of the 1952 Western Electronic Show and Convention and has been a Director of the West Coast Electronic Manufacturers Association as well as Vice-Chairman of the Los Angeles Council.

❖

About 200 of his associates joined in a luncheon this spring at the Naval Research Laboratory to honor **Samuel D. Summers** (SM'50), consultant to the Electronics Division, who is retiring from Civil Service.

A native of Hornbeak, Tenn., Mr. Summers attended Memphis State College; he received the Bachelor of Science in Electrical Engineering from the Tri-State College in Angola, Indiana. He did graduate work at the Massachusetts Institute of Technology and the University of Michigan and received the Master's degree in e.e. from the University of Maryland.

He was an instructor in electrical engineering at Tri-State College in 1923 and later became a professor at the school. For three years, he was head of the Electrical Engineering Department. From 1925 to 1931, Mr. Summers was an electrical engineer with the Commonwealth & Southern Corporation at Jackson, Michigan, where he designed electrical power stations and transmission lines.

During World War II, Mr. Summers served as an electronics officer in the Navy, and was assigned to the Naval Research Laboratory as a member of the Aviation Electronics Service Unit. He joined NRL as an advisor on problems relating to naval electrical and electronics applications in 1945.

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

## Membership

**The following transfers and admissions were approved and are now effective:**

### Transfer to Senior Member

Basore, B. L., 2405 Parsifal, N. E., Albuquerque, N. Mex.
Bindner, J. T., 550 Federal Office Bldg., Seattle, Wash.
Boehm, W. R., Norfolk Naval Shipyard, Portsmouth, Va.
Brown, J., 4804 Lackawanna St., College Park, Md.
Bull, W. A., 16465–21 Ave., S.W., Seattle 66, Wash.
Cahn, S. L., 7712 Danvers St., Downey, Calif.
Chandler, J. F., 7309 W. Touhy Ave., Chicago 31, Ill.
Coan, E. M., 45 Overlook Rd., Cedar Grove, N. J.

*(Continued on page 56A)*


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(Continued on page 58A)

# HUGHES

*subminiature*

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A greater range of diode characteristics—this is what the vigorous and continuing Hughes program of research and development means to you. For instance, you can spell out your requirements for germanium diodes in terms of your particular circuit application. Frequently, you will find that there is a Hughes type in the extensive line that matches those requirements with just the right characteristics. If not, a special type, tested to meet your exact requirements, can be supplied readily.

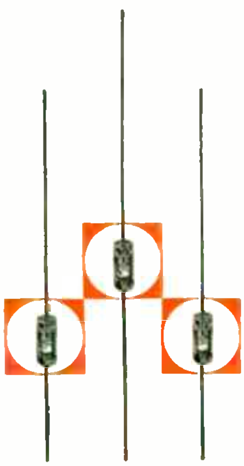
*Listed below are a few of the more popular types, arranged for quick and easy selection, according to forward and reverse characteristics. More detailed specifications are given in pertinent data sheets.*

The Hughes line of semiconductors is being steadily expanded. New germanium and silicon devices, including transistors and power rectifiers, now under development, are being readied for commercial production. Watch for their release. Meanwhile, whenever your equipment design calls for subminiature germanium diodes, be sure to specify Hughes. With extraordinary records of failure-free service, they are *first of all* . . . for RELIABILITY!

| WORKING INVERSE VOLTAGE | | FORWARD CURRENT (Milliamperes) | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|
| | | 3-5 | 5 | 10 | 20 | 50 | 100 | 150 | 200 |
| 30 | 200μA @ – 20V | | | | | | HD 2152 | | |
| 40 | 10μA @ – 10V | 1N128* | | | | | | | |
| 60 | 50μA @ – 50V<br>100μA @ – 50V<br>500μA @ – 50V<br>850μA @ – 50V | | 1N116<br>1N90<br>1N126* | 1N117<br>1N95 | 1N118<br>1N96 | HD 2167<br>HD 2166<br>HD 2155 | HD 2173<br>HD 2174<br>HD 2162 | | HD 2160<br>HD 2171<br>HD 2172 |
| 80 | 50μA @ – 50V<br>100μA @ – 50V<br>125μA @ – 50V**<br>250μA @ – 50V**<br>250μA @ – 50V<br>500μA @ – 50V | 1N67A<br>1N89 | 1N191**<br>1N192**<br>1N198* | 1N99<br>1N97 | 1N100<br>1N98 | HD 2151<br>HD 2168<br>HD 2169 | HD 2150<br>HD 2163<br>HD 2175 | | HD 2158<br>HD 2157<br>HD 2159 |
| 100 | 180μA @ – 90V<br>500μA @ – 100V<br>625μA @ – 100V<br>300μA @ – 50V<br>50μA @ – 50V | 1N68A<br>1N127*<br>HD 2051 | | | | HD 2170 | HD 2165 | HD 2154 | HD 2161 |
| 150 | 500μA @ – 150V | | 1N55B | | | | | | |

*JAN Types. 1N198 only high-temperature tested at 75°C.
**Computer Types. Special recovery tests. 1N191 and 1N192 tested for back current at 55°C.

All Hughes diodes are presently packaged in the famous one-piece, fusion-sealed glass envelope, impervious to moisture and to external contamination. Maximum dimensions, standard glass envelope:
Length, 0.265 inch; Diameter, 0.105 inch.



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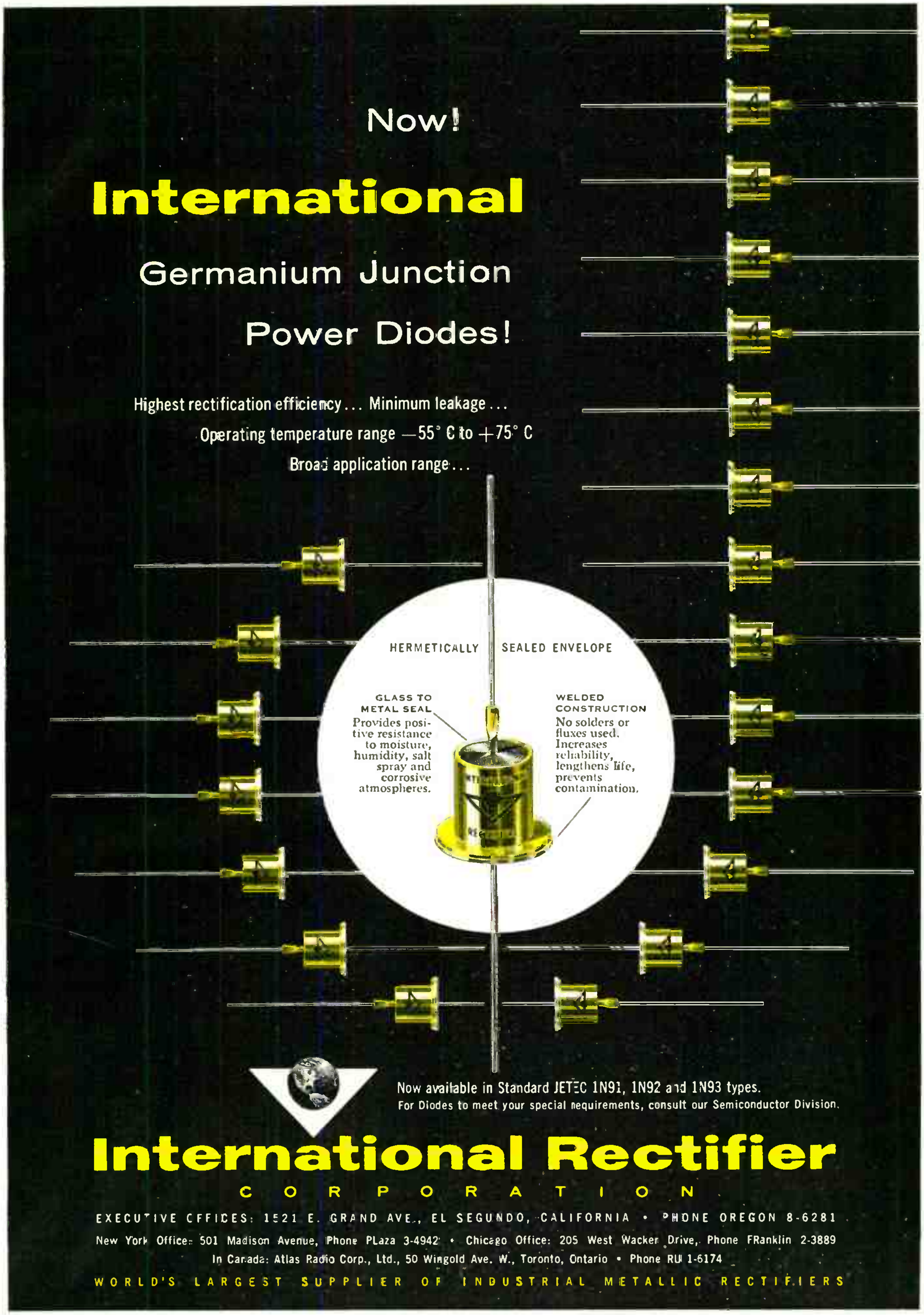
(*Continued on page 60A*)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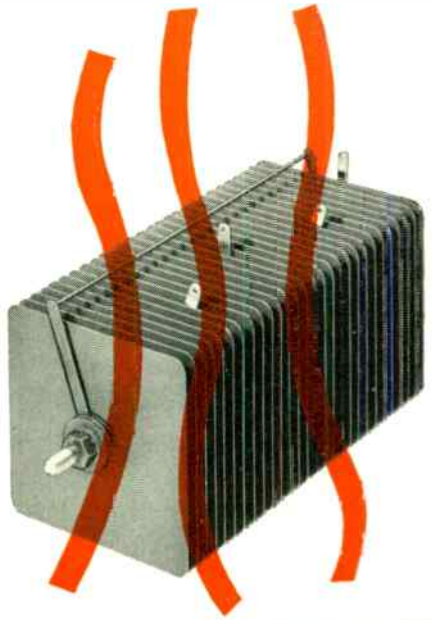
(Continued on page 62A)

latest result of **RADIO RECEPTOR** research



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| AMBIENT TEMPERATURE | POTENTIAL LIFE WITHOUT DERATING |
|---|---|
| 125° C | 500-1000 hours |
| 100° C | 1500-2500 hours |
| 85° C | 2500-4000 hours |

| CELL TYPE | CELL SIZE (Inches) | CONTINUOUS DC AMPERES SINGLE PHASE | | MAX. REVERSE RMS VOLTS PER CELL |
|---|---|---|---|---|
| | | Half Wave | Bridge or Center Tap | |
| X | ¼ (Diam.) | .005 | .010 | 24 |
| Y | ½ x ½ | .025 | .050 | 24 |
| J | 11/16 x 11/16 | .065 | .130 | 24 |
| M | 1 x 1 | .075 | .150 | 24 |
| P | 1 3/16 x 1 3/16 | .150 | .30 | 24 |
| Q | 1½ x 1½ | .30 | .60 | 24 |
| S | 2 x 2 | .50 | 1.0 | 24 |
| U | 3 x 3 | 1.2 | 2.4 | 24 |
| W | 4 x 4 | 2.25 | 4.5 | 24 |
| T | 4¼ x 6 | 3.35 | 6.7 | 24 |
| H | 5 x 6 | 4.0 | 8.0 | 24 |

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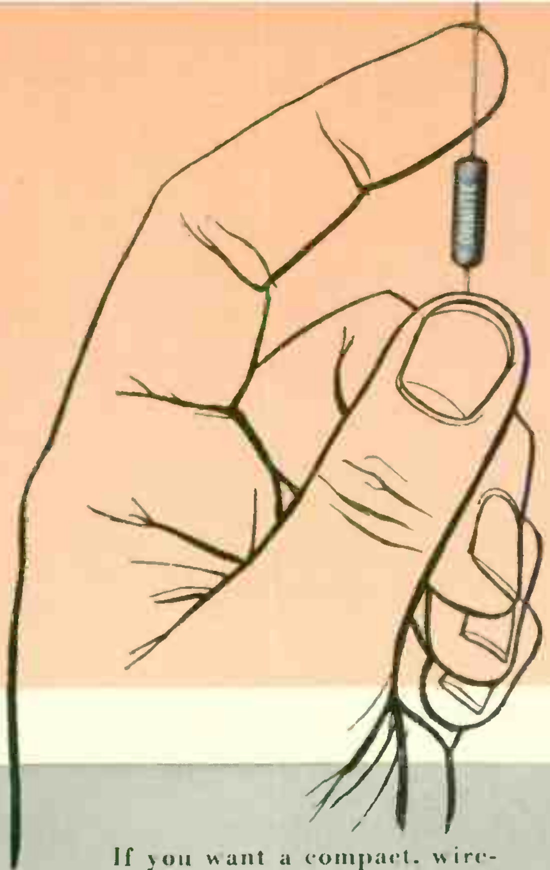

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**18 SIZES — 3 to 15 WATTS***
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Write for Bulletin 147

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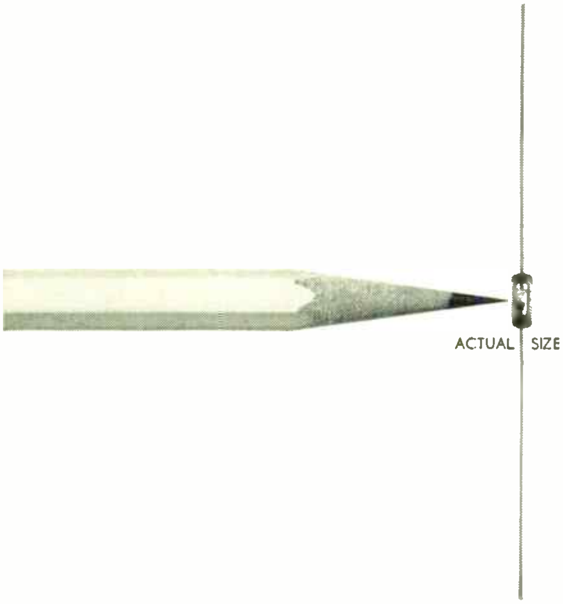
*(Continued on page 66A)*

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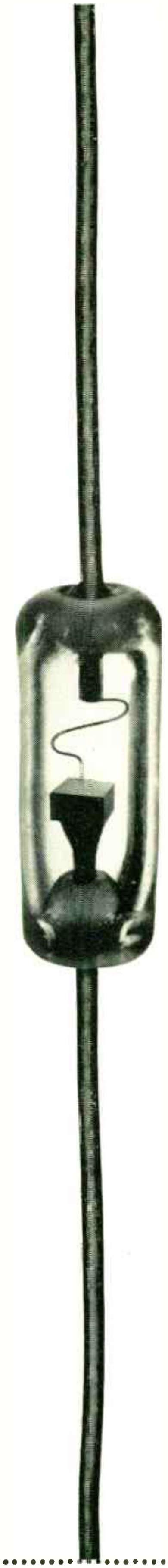


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(Continued on page 70A)

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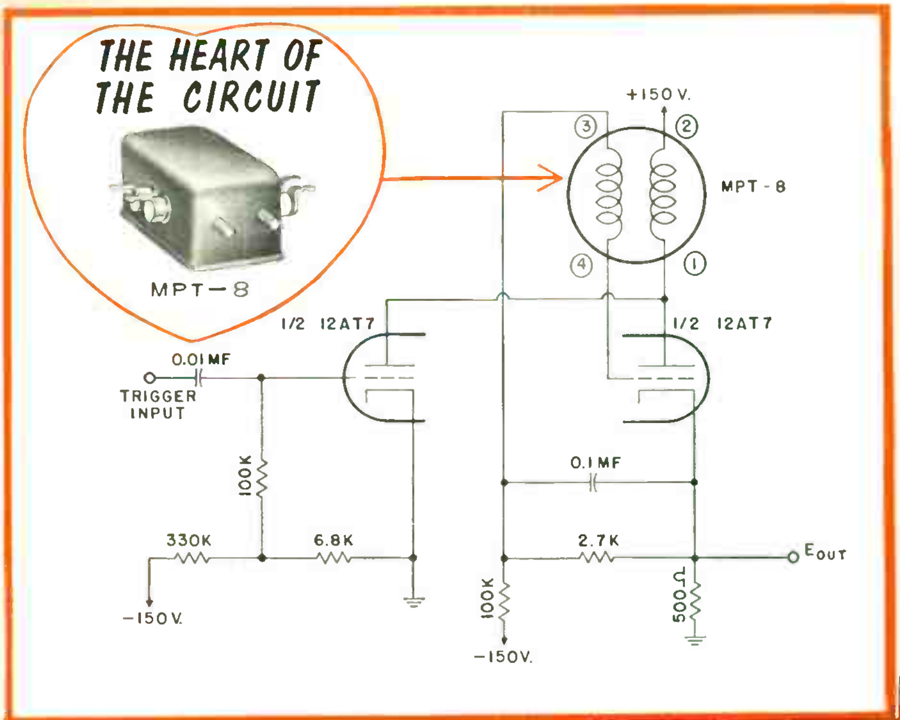


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| MPT-11 | Blocking oscillator, interstage coupling or low power output. | 1.0/1.0/1.0 | 1.0-5.0 | .002 | 2.0 | 500 | DM-01 |
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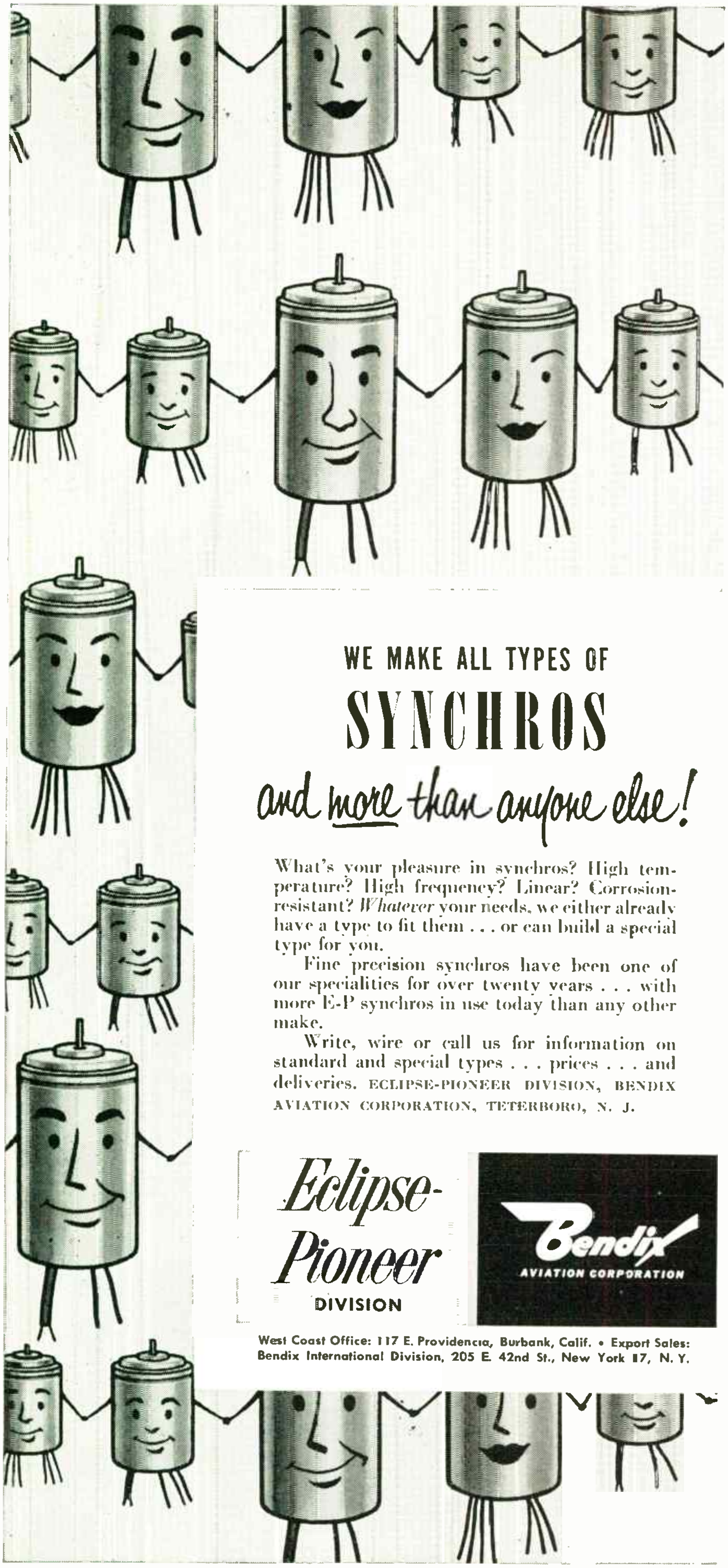


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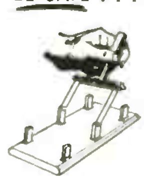

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(Continued on page 78A)

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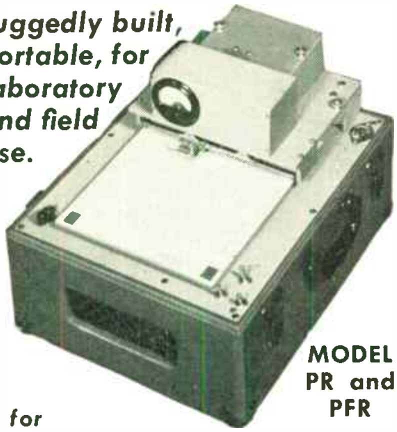

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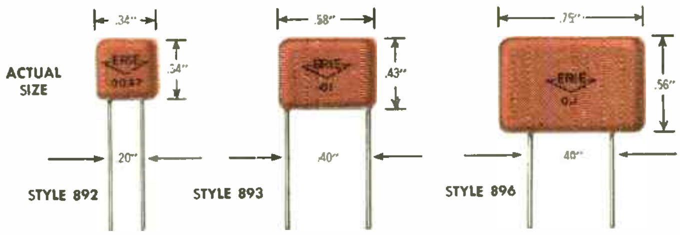


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| 892 | .0033 ± 20% | 896 | .022 ± 20% |
| 892 | .0047 ± 20% | 896 | .033 ± 20% |
| 893 | .0068 ± 20% | 896 | .047 } + 80% |
| 893 | .01 ± 20% | 896 | .068 } − 20% |
| | | 896 | 0.1 } |

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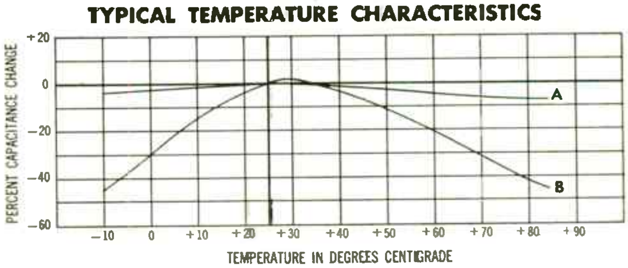


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# PROCEEDINGS OF THE IRE

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## CONTENTS







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# Howard Vollum

Director, 1955

Howard Vollum was born on May 31, 1913, in Portland, Oregon. He attended Columbia University in Portland from 1931 to 1933, transferring to Reed College in 1934. In 1936 he received the B.A. degree in physics from the latter school.

Upon graduation from college, he spent the next few years servicing and installing home, auto, and aircraft radios and constructing electronic devices. From 1940 to 1941 he was Supervisor of Radio Project, NYA, in Portland. Mr. Vollum served as an officer in the U. S. Army Signal Corps from 1942 to 1946. His first two years in service he spent at ADRDE, in Malvern and Christ Church, England, working on coast artillery fire control radar. He was awarded the Legion of Merit for this work. For the next two years he was stationed at the Evans Signal Laboratory in Belmar, New Jersey, in charge of a subsection concerned with the use of radar by ground forces. As a result of this contribution, the Oak Leaf Cluster was added to his award.

In 1946 Mr. Vollum helped to found Tektronix, Inc., of which he is now President. Mr. Vollum is known for his work on the development of the cathode-ray oscilloscope. In recognition of his achievements, Portland University awarded him an honorary Sc.D. degree in 1953.

Mr. Vollum became a Senior Member of the IRE in 1950, and received the Fellow Award in 1955, "for his contribution to the development and manufacture of electronic laboratory instruments." In 1954 he was Chairman of the Portland Section.

# Changes in the IRE Dues Structure

J. D. Ryder, President

Because of the breadth of the electronics field, including as it does most aspects of electrical engineering and many areas in physics, the IRE has become one of the large and well-respected professional engineering societies of the world. As such, W. R. Hewlett, our president in 1954, believed it undesirable that our organization should have so many qualified professional members in the Associate grade of membership, and therefore unable to vote and participate fully in IRE activities. At the March, 1955, meeting of the Board of Directors, certain changes were made in the membership and dues structure of the IRE, which it is hoped will channel new members more directly into membership grades properly representative of their professional qualifications.

For many years the IRE dues pattern has provided for annual dues of $10 for the first five years of Associate membership, all other dues being $15 per year. This arrangement has furnished in the past, a financial incentive to initial application at the Associate, or non-professional level, and normal human inertia, or reluctance to fill out and file transfer blanks, has kept many professionally-qualified members there. It would seem more desirable, however, that a potential member's professional qualifications, and not a favorable dues condition, should determine his initial grade of IRE membership, and this the new dues structure aims to provide.

While other engineering societies are finding the going difficult and are raising dues, the IRE remains in strong financial condition, and it seemed inadvisable and unnecessary to adopt the obvious possibility of eliminating the $10 initial Associate dues level. Accordingly, the Board of Directors has adopted the reverse policy and has ruled that for all new members elected after July 1, 1955, regardless of grade of membership, the annual dues shall be $10 for the first three years of membership, after which the dues rise to the present maximum level of $15 annually. This will represent a dues reduction for new elections as Members or Senior Members. It is hoped thereby that new applicants will enter in the grades for which they are qualified, and thus will strengthen the IRE through increase of qualified professional membership.

Several other changes have been made in the bylaws governing membership. Age limits have been eliminated in all grades, the Board feeling that anyone who has achieved the required professional standing should be permitted the privileges of that standing, irrespective of age. It has also been pointed out that the Associate grade requires merely "an interest in radio," whereas the graduate of an accredited four-year school who has specialized in radio, electrical engineering, or allied studies certainly has progressed beyond mere interest in radio. This fact has been recognized through bylaw changes allowing such graduates to enter as, or transfer to Member level, the lower professional grade, by granting of three years of professional credit for the four school years instead of the previous two year credit. IRE Student Members graduating from such curricula will henceforth transfer to Member instead of Associate grade, upon expiration of Student Member status on graduation.

It should be noted that the changes in dues structure are to affect only new members elected *after* July 1, 1955, and do not apply to members admitted before that time. Thus Associates elected prior to that date will continue their five years at the rate of $10, whereas Associates, Members, or Senior Members elected after July 1, 1955, will have three years at the new rate. To newly elected Members and Senior Members the revised plan represents a slight reduction in dues.

It seems desirable to reemphasize the feeling of the Board that the strength of the IRE depends on sufficient numbers of professionally qualified members, and that it is of great importance for every member to be in the highest membership grade for which he is professionally qualified. To this end. Section Membership Chairmen stand ready to aid. It is also suggested that every Associate and Member study page 5 of the 1954 Directory to determine if he has the qualifications necessary to a higher membership grade. Such transfer can be of much value to each individual and to the IRE in furthering its professional responsibilities.

# Frequency Aging of High-Frequency Plated Crystal Units*

A. W. WARNER†, MEMBER, IRE

The following paper was procured and recommended for publication in the PROCEEDINGS by the Professional Group on Ultrasonics Engineering.—*The Editor*

*Summary*—**The frequency aging of high-frequency crystal units is explained in terms of residual contamination, which may be a partial molecular layer. Experimental data are given on methods designed to reduce frequency aging to a minimum.**

THE ALLOWABLE change in frequency of a crystal unit with time is often specified in parts in $10^6$ per month, and for the more precise frequency standards is specified in parts in $10^8$ per week. These tolerances are several orders of magnitude beyond those required of other standards used in communications measurements. To maintain and improve such tolerances requires the application of a high degree of skill and engineering ingenuity. It is the purpose of this article to show the relationship betweeen various fabrication methods and frequency aging, and to present typical frequency aging data on plated AT crystal units in metal and glass enclosures.

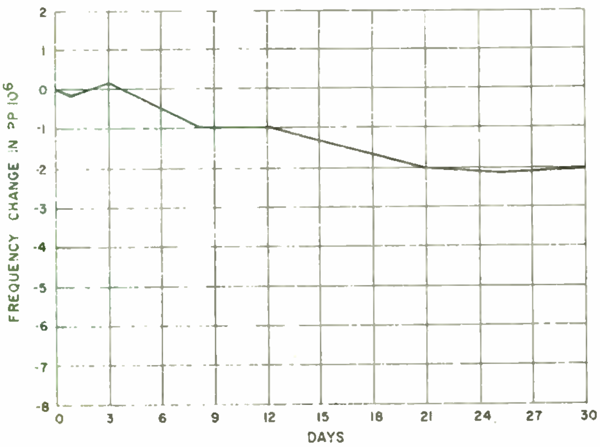


Fig. 1—Aging record at 70 degrees C. of a CR32 type crystal unit' 44 megacycles, fifth overtone, enclosed in an HC6 metal holder.

The resonant frequency of a crystal unit is determined by its mechanically vibrating section, consisting of the quartz plate, electrodes deposited on its two major faces, and to some extent the mounting or support wires.[1] At sufficiently high frequencies, or with specially contoured quartz plates,[2] the mounting can be effectively divorced from the vibrating system, leaving a part of the crystal plate and its associated metal electrodes as the frequency determining elements. Any change in the elastic constants or the mass of these elements will cause a change in frequency.

The aging of typical AT crystal units can be best explained by the transfer to or from the quartz plate of material other than quartz or electrode metal. This is not to say that aging cannot be caused by loss of quartz from the surface and by migration within the metal electrode, but with present-day methods of etching quartz and the use of noble metals for electrodes, aging from this source is much less prevalent than from contaminants.

Fig. 1 shows the aging record at 70 degrees C. of a 44-megacycle, fifth-overtone, CR32-type crystal unit enclosed in the commonly used metal HC6 holder. The aging is two parts per million per month for the first month of operation. This aging rate is about normal for

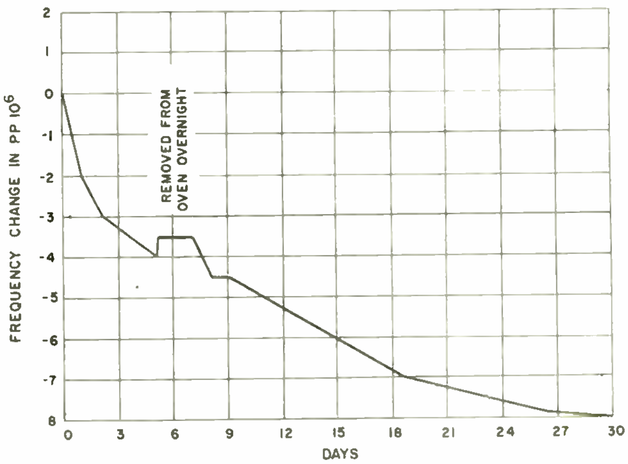


Fig. 2—Aging record at 70 degree C. of a contaminated metal enclosed crystal unit, 15-mc, third-overtone.

the CR32 type of crystal unit. The downward direction of the aging curve is attributed to the slow transfer of contamination to the crystal plate from the inner walls of the metal enclosure. This assumption seems reasonable in view of the fact that the quartz crystal plate and its gold electrode can be cleaned more effectively than can the metal enclosure.

Fig. 2 shows the aging record of a similar crystal, except that additional contamination, probably soldering flux, is present. This is surmised from the fact that the unit failed to pass a test for low vapor content. The aging rate is four times as bad and, since the direction of frequency change is reversed upon exposure to room

* Original manuscript received by the IRE, April 6, 1955.
† Bell Telephone Laboratories, Inc., Whippany, N. J.
[1] R. A. Sykes, "High-frequency plated quartz crystal units," PROC. IRE, vol. 36, pp. 4–7; January, 1948.
[2] A. W. Warner, "High-frequency crystal units for primary frequency standards," PROC. IRE, vol. 40, pp. 1030–1033; September, 1952.

temperature, it is evident that equilibrium is a function of temperature. Although this crystal unit might eventually age at an acceptably low rate, any change in temperature, such as an oven shut down, would start a new aging cycle.

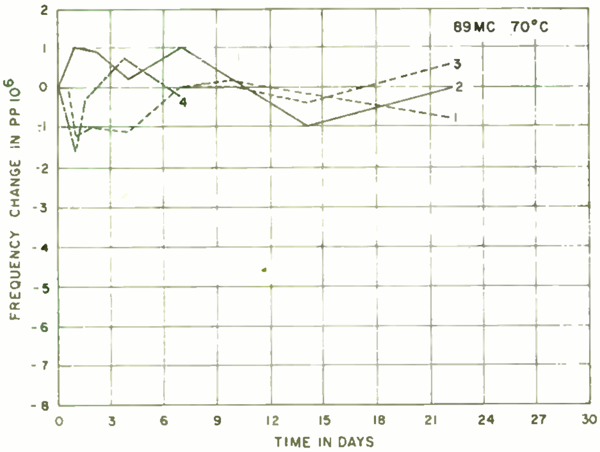


Fig. 3—Aging record of metal enclosed, 89-mc, seventh-overtone crystal units at 70 degrees C.

Fig. 3 shows the aging record of a group of three 89-mc, seventh-overtone crystal units in metal holders at 70 degrees C. At this frequency the quartz plate is less than half as thick as that of Fig. 2, yet no aging is apparent within the error of measurement, $\pm 1$ part per million. These units were baked for 24 hours prior to assembly. Fig. 4 shows the aging record of two third-overtone crystal units made in the laboratory, with great care taken in the cleaning and sealing methods. The aging is about 1 ppm for the extended period of four months, five to ten times better than the crystal in Fig. 1.

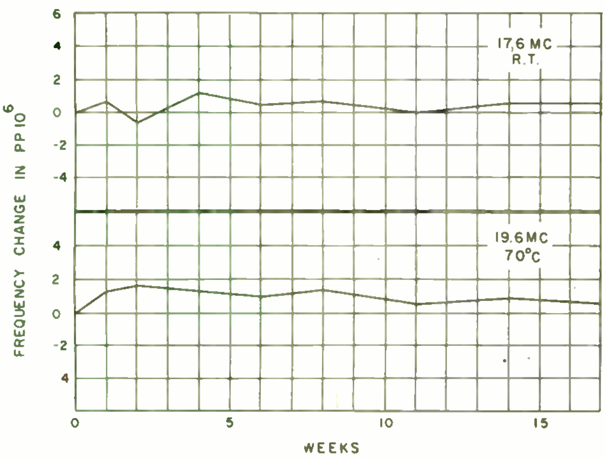


Fig. 4—Aging record of two third-overtone laboratory-made crystal units, metal enclosed.

In view of the foregoing experiments, it was reasoned (1) that a glass enclosure, with a smooth surface which can be easily degassed and cleaned, should be superior to a metal enclosure, and (2) that by the use of glass, aging rates low enough for primary frequency standard use might be achieved. To demonstrate this a group of crystal plates having a large frequency determining dimension, 10 times that of the crystal unit in Fig. 1 were tried, first in evacuated HC6 metal enclosures and then in glass bulbs. The frequency measuring equipment was improved to measure parts in $10^8$ rather than $10^6$. Using a 25X expanded frequency scale, it can be seen in Figs. 5 and 6 that in the metal enclosure the frequency aging is slightly erratic and downward, and in the glass enclosure it is uniform and upward. The conclusions reached from this experiment were (1) that the

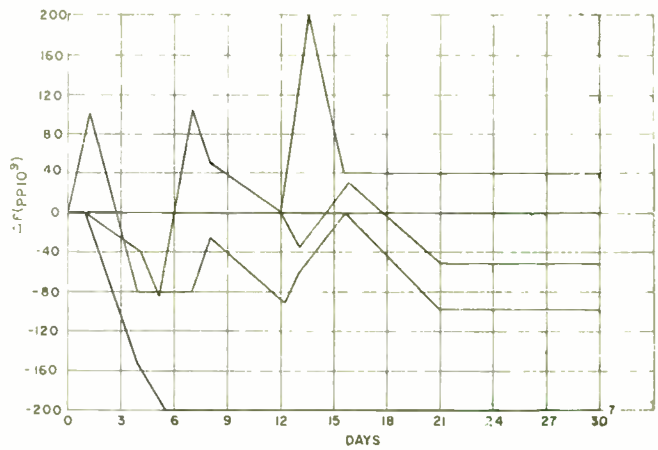


Fig. 5—Aging record of 5-mc, fifth-overtone crystal units in HC6 metal enclosures, at 60 degrees C.

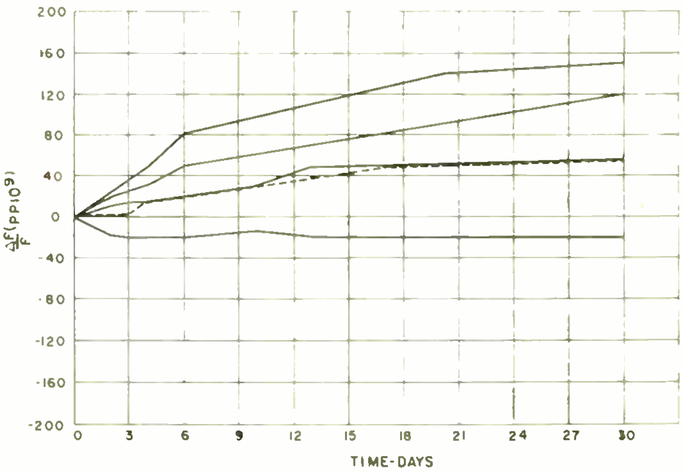


Fig. 6—Aging record of 5-mc, fifth-overtone crystal units in T5-1/2 glass bulbs, at 60 degrees C.

glass enclosure was contaminated less than the crystal plate surface, which is apparently losing mass to its surroundings, and (2) that for further improvement of frequency aging the crystal plate surface should now be improved.

The next experiments were performed on polished quartz plates. It was reasoned that a smooth surface would not only have less surface area, but also could be more easily cleaned. The contaminants are not imbedded in tiny crevices and can be removed by relatively short exposures to cleaning agents. Fig. 7 shows the aging record of the polished crystal plates in glass enclosures. These are twice as good as the unpolished plates and are uniform, but the indications are still that the crystal plate is not entirely free of contamination.

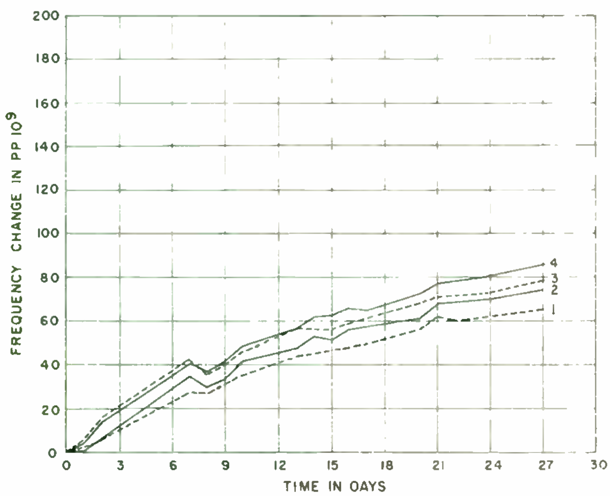


Fig. 7—Aging record of polished 5-mc, fifth-overtone crysta units in glass bulbs, at 60 degrees C.

This was investigated further in an aging test designed to show the relative amounts of contamination removed by various processes. Referring to Fig. 8, crystal units number 5 and 6 were made as in the previous experiment. Crystal unit number 4 is a similar

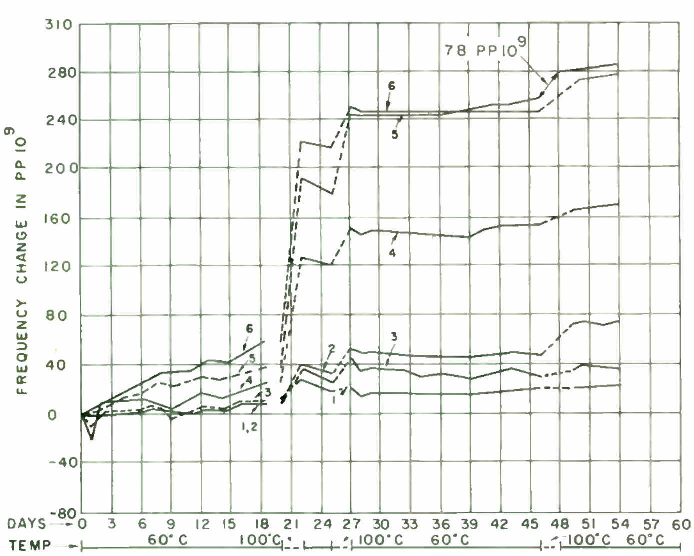


Fig. 8—Aging record of various polished 5-mc, fifth-overtone crystal units in glass bulbs at 60 degrees C. with a 24-hour 100 degrees C. bake at intervals.

unit that had been through the previous aging test. Crystal units 1, 2, and 3, were made in the same manner as units 5 and 6, except that they were vacuum baked in a special apparatus just prior to final sealing. During the aging test the crystals were removed periodically and exposed to a temperature of 100 degrees C. for 24 hours. It will be seen that the vacuum baking technique had the effect of removing about 80 per cent of the remaining contamination, making the crystal unit much less susceptible to frequency changes due to temperature interruptions.

Fig. 9 shows the aging record of a group of five crystal units made in accordance with the above principles; that is, smooth surfaces and maximum removal of contamination. The average aging of the five units is nearly zero, the maximum excursion 10 parts in a billion in 30 days.

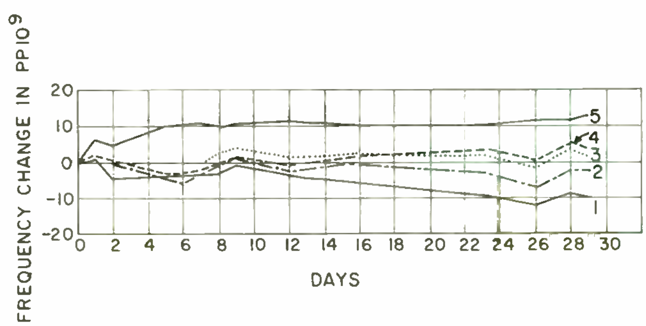


Fig. 9—Aging record at 75 degrees C. of crystal units baked 20 hours at 140 degrees C.

From these data one can conclude empirically that the mechanism of frequency aging of high-frequency crystal units is one of transfer of mass to and from the crystal plate. The rate of transfer and the degree of permanence after transfer will be a function of the vapor pressure of the contaminant and the degree of adherence, which may be molecular, chemical, or mechanical, between the contaminant and the electrode material. For this reason, it is not likely that any degree of permanence would be achieved by stabilizing a contaminated crystal after sealing, as by heat cycling and the like. Also, it is probably not possible to predict an aging curve, since the equilibrium reached at any given operating condition is not likely to repeat itself.

The relationship between contaminant and frequency change may be clearer if one calculates the mass involved. A change in frequency of 1 part in $10^9$ in the crystal units of Fig. 9 requires a change in mass at the surface of $2\times10^{-4}$ micrograms per square centimeter. If this mass were, for example, an oil film having a density of 1, it would be only two hundredths of an angstrom thick. Since a molecular layer is at least 4Å thick, this represents only a partial molecular layer.

It should be recognized, however, that a high degree of frequency stability is obtainable from crystal units mounted in the widely used HC6 metal enclosure, i.e., 1 part in $10^6$ for the first month of operation with continuing improvement as long as operating conditions are unchanged. Where frequency stability greater than this is required, as in primary frequency standards, the cleaner glass-enclosed crystal units are preferable, despite their larger size.

# Some Gyrator and Impedance Inverter Circuits*

B. P. BOGERT†

*Summary*—**The use of feedback amplifiers to provide impedance inversion is considered. If the proper types of feedback connections are used, and the input and output impedances of the amplifier and its gain are chosen suitably, the input impedance of the circuit will be approximately proportional to the reciprocal of the load impedance. The approximation may be improved by the insertion of negative impedance elements to compensate for the residual impedances in the inverter. When this is done, the circuit exhibits gyrator properties. Experimental verification of one of the circuits, using high quality stable amplifiers, gave a range of impedance inversion of two decades.**

IT IS THE purpose of this paper to discuss some feedback amplifier circuits which act as impedance inverters and as gyrators. The gyrator[1] may be considered to be a two terminal-pair circuit which has the property that the phase shift for transmission in one direction differs by 180 degrees from that for transmission in the other direction, over a broadband of frequency.[2] This is a consequence of the fundamental property of a gyrator, which is that the transfer impedance $z_{12}$ be $R$, $z_{21}$ be $-R$, and that $z_{11}$ and $z_{22}$ be zero. Another property which follows from the above is that of impedance inversion. If, as a starting point, we consider circuits capable of impedance inversion, it is possible to arrive at some active circuits for gyrators.

We consider the relation between the input impedance and the load impedance of amplifiers with external feedback connections of specified types. By using the proper feedback connections, and by suitable choice of amplifier input and output impedances and gain, the circuit approximates an impedance inverter. The residual impedances may be corrected by addition of negative impedance elements, and the resulting circuits possess the properties of a gyrator. Since the gyrator circuits so obtained employ active elements, their over-all stability is an important consideration.

Consider first the fourpole shown in Fig. 1(a). The input impedance $Z_{in}$ when an impedance $Z_L$ terminates the output is given by the expression

$$Z_{in} = z_{11} - \frac{z_{12}z_{21}}{z_{22} + Z_L}. \tag{1}$$

In order to make an impedance inverter, we must have $z_{11} = z_{22} = 0$, and $z_{12}z_{21}$ negative real. If we cannot strictly realize the first condition, we should have, in the impedance range of interest,

$$z_{22} \ll Z_L \tag{2}$$

and

$$z_{11} \ll \frac{z_{12}z_{21}}{Z_L}. \tag{3}$$

Consider now the circuit of Fig. 1(b). It consists of an amplifier having an input impedance $Z_1$, a passive output impedance $Z_2$ and open circuit gain $k$. The output is fed back into the input using a series connection at the output end and a parallel connection in the input circuit.

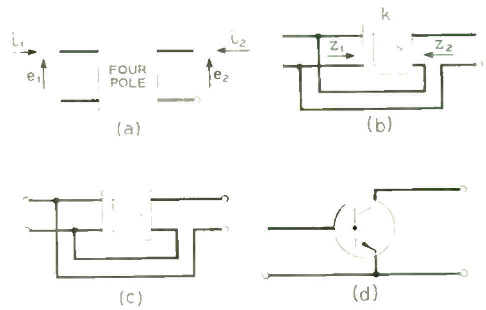


Fig. 1—(a) Basic fourpole with current and voltage conventions. (b) Impedance inverter circuit. (c) Alternative form of circuit of Fig. 1(b). (d) Grounded emitter transistor stage.

If the output circuit is terminated in an impedance $Z_L$, the input impedance $Z_{in}$ is:

$$Z_{in} = \frac{Z_1(Z_2 + Z_L)}{Z_1(k+1) + Z_2 + Z_L}. \tag{4}$$

If we make

$$Z_1(k+1) + Z_2 = 0, \tag{5}$$

and

$$Z_2 \gg Z_L \tag{6}$$

then

$$Z_{in} \cong \frac{Z_1Z_2}{Z_L} = \frac{Z^2}{Z_L}, \tag{7}$$

which has the desired property. This circuit is easily recognized as a form of feedback amplifier,[3] but its action as an impedance inverter does not appear to be particularly well known.

An alternative circuit is obtained by interchanging the feedback leads and replacing $k$ by $-k$. This circuit is shown in Fig. 1(c), which makes it evident that we are dealing with the ordinary reactance tube circuit, except that for operation as an impedance inverter, the phase shift in the amplifier must be zero when $Z_1$ and $Z_2$ are resistances. Thus we are led to consider the use of a

* Original manuscript received by the IRE, December 1, 1954; revised manuscript received, March 10, 1955.

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

[1] B. D. H. Tellegen, "The gyrator, a new electric network element," *Philips Res. Rep.*, vol. 3, pp. 81–101; April, 1948.

[2] W. Nonnenmacher and F. Schreiber, "Der zweidrahtverstärker als gyrator and als vierpol zur herstellung ungewöhnlicher scheinwiderstände," *Frequenz*, vol. 8, pp. 201–204; July, 1954.

[3] R. B. Blackman, "Effect of feedback on impedance," *Bell Sys. Tech. Jour.*, vol. 22, p. 276 (Fig. 4); October, 1943.

transistor, using the circuit shown in Fig. 1(d). This circuit is essentially the grounded emitter transistor circuit, and the input impedance of such a stage is given in terms of the transistor constants as:[4]

$$R_i = r_e + r_b + \frac{r_e(r_m - r_e)}{r_e + r_c - r_m + R_L}. \tag{8}$$

and, looking from the output end:

$$R_0 = r_e + r_c - r_m + \frac{r_e(r_m - r_e)}{r_e + r_b + R_g}. \tag{9}$$

If we employed a transistor having the idealized properties $r_b \to 0$; $r_c - r_m \to 0$; $r_e \to 0$ and $r_m \to \infty$ such that $r_e r_m = R^2$, then a good impedance inverter could be made by using a grounded emitter transistor stage.

It is convenient to regard these circuits as feedback amplifiers, and to characterize them in terms of the feedback connections at the input and output. The circuit we have been discussing is parallel connected at the input and series connected at the output. It will be denoted by PS. Let us see how other circuits behave as regards impedance transformations. We consider, in order, the series-series (SS), series-parallel (SP), parallel-series (PS), and parallel-parallel (PP). Using the symbols and conventions shown in Fig. 2 we have for the impedance transformations:

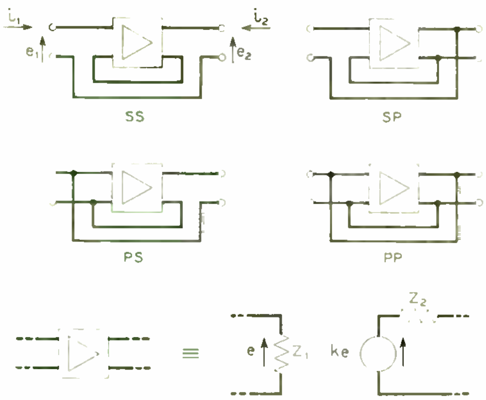


Fig. 2—General forms of feedback amplifier circuits discussed. SS = series-series; SP = series-parallel; PS = parallel series; PP = parallel-parallel. First letter refers to input feedback connection, second letter to output feedback connection. Meaning of amplifier symbol shown.

$$\text{SS}: Z_{in} = (1-k)Z_1 + Z_2 + Z_L = \alpha + Z_L \tag{10}$$

$$\text{SP}: Z_{in} = \frac{Z_1Z_2 + Z_L[Z_1(1-k) + Z_2]}{Z_2 + Z_L} = \frac{Z_1Z_2 + \alpha Z_L}{Z_2 + Z_L} \tag{11}$$

$$\text{PS}: Z_{in} = \frac{Z_1(Z_2 + Z_L)}{Z_1(1-k) + Z_2 + Z_L} = \frac{Z_1(Z_2 + Z_L)}{\alpha + Z_L} \tag{12}$$

$$\text{PP}: Z_{in} = \frac{Z_1Z_2Z_L}{Z_1Z_2 + Z_L[Z_1(1-k) + Z_2]} = \frac{Z_1Z_2Z_L}{Z_1Z_2 + \alpha Z_L} \tag{13}$$

where $\alpha = Z_1(1-k) + Z_2$, and is assumed real.

Inspection of (10) and (13) show that it is impossible to obtain impedance inversion with the SS and PP feedback connections. We have already discussed the PS connection so it remains to consider the SP circuit. For good inversion we need

$$Z_2 \ll Z_L, \qquad \alpha = 0 \tag{14}$$

in which case

$$Z_{in} = \frac{Z_1Z_2}{Z_L}. \tag{15}$$

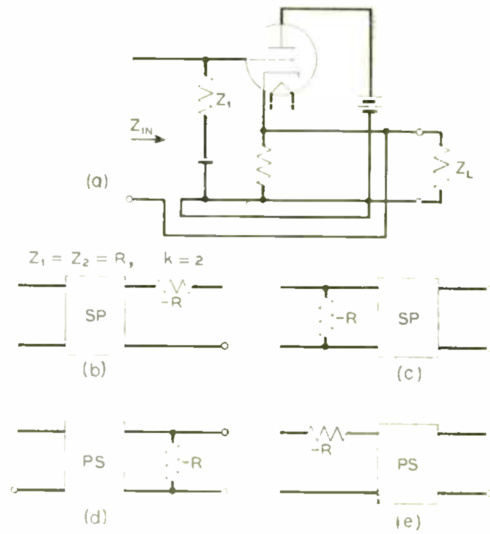


Fig. 3—(a) Ideal cathode follower type impedance inverter. (b) Gyrator formed from SP inverter circuit with addition of series resistance $-R$ in output lead. (c) Gyrator formed from SP inverter circuit with addition of shunt resistance $-R$ across input. (d) Gyrator formed from PS inverter circuit with addition of shunt resistance $-R$ across output. (e) Gyrator formed from PS inverter circuit with addition of series resistance $-R$ in input lead.

A possible circuit is shown in Fig. 3(a), which involves the use of an ideal cathode follower.[5] The impedance to be inverted is placed from cathode-to-ground and the inverted impedance appears from cathode-to-grid.[6]

For the SP and PS cases we have for the fourpole impedance matrix:

$$\text{SP}: \|z\| = \begin{Vmatrix} \alpha & -Z_2 \\ Z_1 - \alpha & Z_2 \end{Vmatrix} \tag{16}$$

$$\text{PS}: \|z\| = \begin{Vmatrix} Z_1 & -Z_1 \\ Z_2 - \alpha & \alpha \end{Vmatrix}. \tag{17}$$

The admittance matrix in each case is

$$\text{SP}: \|y\| = \begin{Vmatrix} Y_1 & Y_1 \\ \alpha' - Y_2 & \alpha' \end{Vmatrix} \tag{18}$$

[4] R. L. Wallace, Jr. and W. J. Pietenpol, "Some circuit properties and applications of n-p-n transistors," *Bell Sys. Tech. Jour.*, vol. 30, pp. 530–563 [eqs. (34) and (37)]; July, 1951.

[5] J. Shekel, "The gyrator as a 3-terminal element," Proc. IRE, vol. 41, pp. 1014–1016; August, 1953.

[6] If the circuit of Fig. 3(a) is redrawn to bring into evidence the combination of the internal feedback of the cathode follower stage and the external SP feedback under discussion, a differently appearing circuit results, which consists of a grounded cathode stage with an impedance $Z_1$ connected between the grid and plate. Although the circuit now appears to belong to the PP family, this is not strictly the case, since the external feedback path has other than zero series impedance. Zero series impedance in feedback paths is a somewhat arbitrary limitation for the types of feedback circuits under discussion.

$$\text{PS: } \|y\| = \begin{Vmatrix} \alpha' & Y_2 \\ \alpha' - Y_1 & Y_2 \end{Vmatrix}, \tag{19}$$

where

$$Y_1 = 1/Z_2,\ Y_2 = 1/Z_2, \quad \text{and} \quad \alpha' = Y_2(1-k) + Y_1. \tag{20}$$

Examination of (10) through (13) shows that the expression $\alpha = Z_1(1-k)+Z_2$ enters into each equation in a rather fundamental way. It is not difficult in a practical experimental circuit to satisfy the condition $\alpha=0$ (which implies $\alpha'=0$). It must be noted if $\alpha \leqq 0$ the circuit is potentially unstable,[7] so that generally, $\alpha$ must be small but greater than 0. If we neglect $\alpha$ we have, from (11) and (12)

$$\text{SP: } Z_{in} = \frac{Z_1 Z_2}{Z_2 + Z_L} \tag{21}$$

and

$$\text{PS: } Z_{in} = \frac{Z_1(Z_2 + Z_L)}{Z_L}, \tag{22}$$

while the matrices (16) and (18) become

$$\text{SP: } \|z\| = \begin{Vmatrix} 0 & -Z_2 \\ Z_1 & Z_2 \end{Vmatrix} \text{ and } \|y\| = \begin{Vmatrix} Y_1 & Y_1 \\ -Y_2 & 0 \end{Vmatrix}, \tag{23}$$

and (17) and (19) become

$$\text{PS: } \|z\| = \begin{Vmatrix} Z_1 & -Z_1 \\ Z_2 & 0 \end{Vmatrix} \text{ and } \|y\| = \begin{Vmatrix} 0 & Y_2 \\ -Y_1 & Y_2 \end{Vmatrix}. \tag{24}$$

If we consider the SP $z$-matrix, we see that if an impedance $-Z_2$ is placed in series in the output circuit, we have a perfect impedance inverter, and if $Z_1=Z_2=R$, and $k=2$ (so that $\alpha=0$), we have

$$\|z\| = \begin{Vmatrix} 0 & -R \\ R & 0 \end{Vmatrix}, \tag{25}$$

which defines a gyrator. In addition, we can make a gyrator if an admittance $-Y_1$ is shunted across the input terminals, with $Y_1 = Y_2 = 1/R$, so that

$$\|y\| = \begin{Vmatrix} 0 & 1/R \\ -1/R & 0 \end{Vmatrix}, \tag{26}$$

which is the admittance matrix corresponding to (25). These circuits are shown in Figs. 3(b) and 3(c). In the same way, a gyrator can be made using the PS circuit by either shunting $-Y_2$ across the output terminals or by inserting $-Z_1$ in series with an input terminal, as shown in Figs. 3(d) and 3(e).

It is interesting to note that in both SP and PS circuits, to correct for the residual $z_{11}$ or $z_{22}$ (which ever occurs), the series negative impedance is placed on the "P" or parallel feedback side, and the shunt negative admittance is placed on the "S" or series feedback side.

[7] A "potentially unstable" circuit is here defined to be one which, by proper passive termination, may be made to oscillate.

Thus, when $Z_1=Z_2=R$, and $k=2$ ($\alpha=0$), the SP and PS circuits can be regarded as identical except for a reversal in amplifier direction.

Since these circuits involve active elements, problems of stability are important. If we suppose that the terminations are passive, but otherwise arbitrary, the circuits must remain stable. In the first place, the condition that $\alpha=0$ is the boundary between stable and potentially unstable operation, as has been mentioned. In addition the negative resistances used to make the gyrators must be of the proper stability type. The series negative resistances must be open-circuit stable and the shunt conductances must be short-circuit stable.[8] The range of impedance seen by the series negative resistance runs from $R$ to $\infty$, while the range of admittance seen by the shunt negative conductance runs from 0 to $1/R$.

Experimental confirmation of one of these circuits was made, using the circuit of Fig. 3(e). The amplifiers used were of a design having accurately controlled input and output impedances (600 ohms) and were very stable. To obtain the negative resistance of $-600$ ohms, an SS circuit was used with $Z_1=Z_2=600$ ohms, $Z_L=0$, and $k=3$ [see (10)]. The theoretical gain setting corresponding to $k=3$ is 9.54 db and the actual one used was 9.8 db.

A measure of the stability of the amplifiers used is given by the fact that the negative resistance obtained by the SS circuit referred to varied by less than 5 ohms from day to day. This resistance variation corresponds to a gain variation of approximately 0.04 decibel.

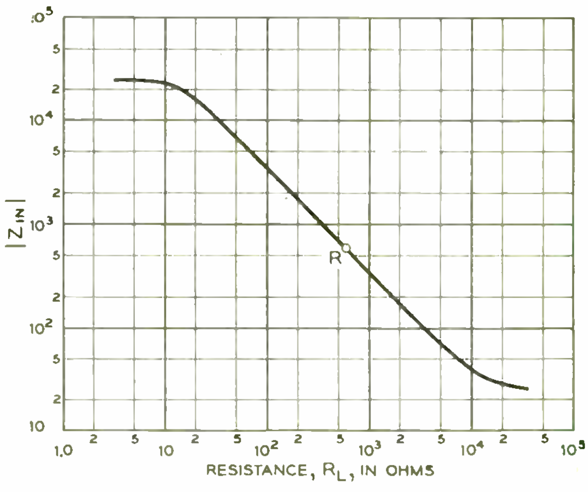


Fig. 4—Measurements of $|Z_{in}|$ vs $R_L$ of an impedance inverter circuit of PS type with series resistance $-R$ in input lead [Fig. 3(e)] $R=600$ ohms.

Measurements of $Z_{in}$ versus $R_L$ at 400 cps were made using a Technology Instrument Company Type 310A $Z$-Angle Meter, the results of which are shown in Fig. 4. The inversion was very good over a two-decade range $40 \leqq R_L \leqq 4{,}000$ ohms. The angle of $Z_{in}$ was less than 10 degrees for $80 \leqq R_L \leqq 3{,}000$ ohms.

[8] H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., New York, N. Y., p. 189 ff.; 1945.

When $R_L$ was placed at the input, and $Z_{in}$ measured at the output, the impedance was inverted, but the inversion range was reduced.

A 0.572 mfd condenser was used as $Z_L$ and the impedance $Z_{in}$ at 400 cps was measured as $535\angle 90$ degrees, whereas the computed impedance $\omega C(600)^2 = 517$ ohms.

Computations were made of the fourpole impedances using the $Z_{in}$ versus $R_L$ data discussed above. These impedances are

$$z_{11} = -2.8 + j24.8 \text{ ohms}$$

$$z_{12} = +598$$

$$z_{21} = -598$$

$$z_{22} = -4.1 + j13.8,$$

which indicate that the experimental circuit approaches an ideal gyrator quite closely.

The applications of these circuits to practical use is limited by the stability and accuracy requirements of the amplifiers employed. The closer the circuit approaches an ideal impedance inverter, the closer it approaches potential instability. For this reason, these circuits might prove useful in those cases in which the residual impedances of the inverter current would not require compensation by additional negative impedance circuits, or the requirements on $\alpha$ would not be too stringent.

# A Bridge for Measuring Audio-Frequency Transistor Parameters*

B. F. C. COOPER†, MEMBER, IRE

***Summary***—**A bridge is described which measures the small-signal parameters of point contact and junction transistors at a frequency of 1 kc. The impedance parameters of point-contact transistors are measured for the grounded-base connection, while a set of parameters representing a compromise between the impedance parameters and the *h* parameters is measured for junction transistors operating in either the grounded-base or grounded-emitter connections. This set includes the short-circuit input impedance $h_{11}$, the short circuit current amplification factor $h_{21}$, a paralleled resistance $r_{22}$ and capacitance $C_{22}$ representing the open-circuit output impedance, and two feedback resistances $r_{12}$ and $r_{12}'$ which in the particular case of the grounded-base connection represent the "low-frequency" base resistance $r_b$ and the "high-frequency" base resistance $r_b'$ respectively. It is also shown that the $\alpha$ cut-off frequency of junction transistors can be calculated with good accuracy from the bridge measurements.**

## Introduction

Equipment for measuring small-signal parameters is an important adjunct to development work on transistors and transistor circuits, and such equipment has been described by a number of authors.[1-6] These authors have generally favored the use of impedance parameters when dealing with point-contact transistors and either the admittance or the *h* parameters when dealing with junction transistors. The bridge which is described in the present paper has been designed, in accordance with the usual practice, to measure the impedance parameters of point-contact transistors. However, the set of parameters measured for junction transistors represents a compromise between the impedance and the *h* parameters, and is discussed more fully in the next section. The bridge which operates at a frequency of 1 kc measures all resistive and reactive parameter components which play a significant part in the audio-frequency operation of transistors. Parameters are measured for the grounded-base connection of point-contact transistors and for the grounded-base or grounded-emitter connection of junction transistors. The other possible connections have not been considered because point-contact transistors are normally used only in the grounded-base connection for small-signal work, and grounded-collector measurements on a junction transistor do not appear to yield any more useful information than can be gotten from the other two connections. It has also been found possible to deduce a fairly accurate figure for the $\alpha$ cut-off frequency of alloyed-junction transistors from the measured phase lag of the parameter $\alpha_{cb}$ at 1 kc.

## Parameter Systems

Transistor small-signal parameters can be specified in a number of ways, depending on which of the four variables, input voltage ($v_1$), input current ($i_1$), output voltage ($v_2$), and output current ($i_2$), are chosen as independent variables. The combinations of $i_1$ with $i_2$, $i_1$ with

* Original manuscript received by the IRE, January 28, 1955.
† Radiophysics Laboratory, City Road, Chippendale, N.S.W., Australia.
[1] K. Lehovec, "Testing transistors," *Electronics*, vol. 22, pp. 88–89; June, 1949.
[2] G. Knight, Jr., R. A. Johnson, and R. B. Holt, "Measurement of the small-signal parameters of transistors," Proc. IRE, vol. 41, pp. 983–989; August, 1953.
[3] L. J. Giacoletto, "Equipment for measuring junction transistor admittance parameters over a wide frequency range," *RCA Rev.*, vol. 14, pp. 269–296; June, 1953.
[4] A. G. Bousquet, "Transistor measurements with the vacuum tube bridge," *Gen. Rad. Exper.*, vol. 27; March, 1953.
[5] D. Dorman, "A bridge transistor tester," *Radio and Television News*, vol. 51, pp. 5–7, 34; February, 1954.
[6] R. F. Shea, "Principles of Transistor Circuits," John Wiley and Sons, Inc., New York, N.Y., ch. 22; 1953.

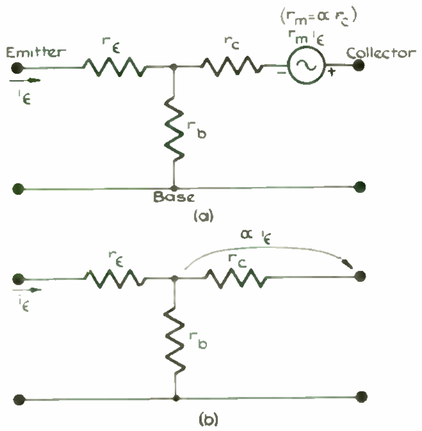


Fig. 1—Grounded-base equivalent $T$-circuits for a point-contact transistor.

$v_2$, and $v_1$ with $v_2$ as independent variables have all received extensive treatment in transistor literature, but in this paper attention will be restricted to the first two combinations which lead, respectively, to the impedance parameters and the $h$ parameters.

### *Impedance Parameters*

The well-known relationships which yield the impedance parameters are[7]

$$v_1 = z_{11}i_1 + z_{12}i_2. \tag{1}$$

$$v_2 = z_{21}i_1 + z_{22}i_2. \tag{2}$$

Eqs. (1) and (2) lead to the definitions

$$z_{11} = v_1/i_1 \mid i_2 = 0 \tag{3}$$

$$z_{12} = v_1/i_2 \mid i_1 = 0 \tag{4}$$

$$z_{21} = v_2/i_1 \mid i_2 = 0 \tag{5}$$

$$z_{22} = v_2/i_2 \mid i_1 = 0. \tag{6}$$

It is also convenient to introduce the current gain parameter $\alpha_{21}$ at this stage, although strictly speaking it belongs among the $h$ parameters. Here we write

$$\alpha_{21} = -i_2/i_1 \mid v_2 = 0, \tag{7}$$

and it may also be seen that

$$\alpha_{21} = z_{21}/z_{22}. \tag{8}$$

In order to give the impedance parameters a physical basis it is customary to express them in terms of the parameters of the transistor equivalent $T$-circuit which for a point contact transistor can take either of the forms shown in Fig. 1(a) and (b). For a junction transistor operating in the grounded-base connection, either of the forms shown in Fig. 2(a) and (b) are applicable; while Fig. 3(a) and (b) apply to the grounded-emitter connection. In the audio-frequency range the parameters of point contact transistors are essentially resistive, and collector capacitance can be ignored. However, collector capacitance is included in the junction transistor circuits because its effect can become significant at frequencies of a few hundred cycles per second. A further important feature of these circuits is the division of the base resistance into two components $r_b'$ and $r_b''$, where the component $r_b'$ results from ohmic resistance in the base region, and $r_b''$ is an equivalent feedback resistance which results from "space-charge layer widening" as described by Early.[8] $r_b'$ is usually termed the "high-fre-

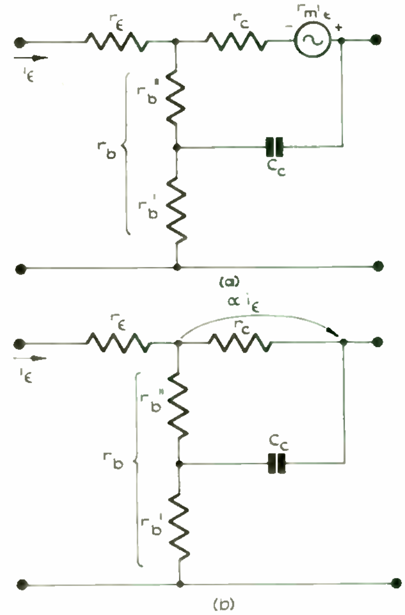


Fig. 2—Grounded-base equivalent $T$-circuits for a junction transistor.

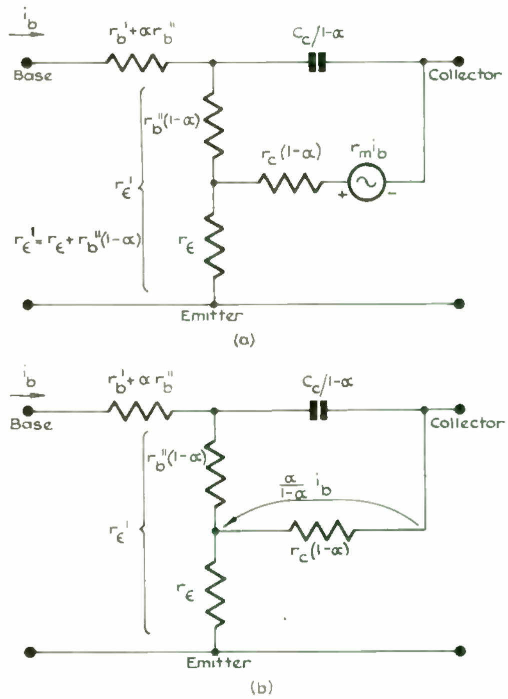


Fig. 3—Grounded-emitter equivalent $T$-circuits for a junction transistor.

[7] The positive direction of current flow will be taken as inwards for a positive applied voltage.

[8] J. M. Early, "Effects of space-charge layer widening in junction transistors," Proc. IRE, vol. 40, pp. 1401–1406; November, 1952.

quency base resistance" since it dominates the feedback at high frequencies when most of the collector signal current flows through $C_c$. In the grounded emitter equivalent circuit it is common practice to omit component of feedback resistance shown as $r_b''(1-\alpha)$ in Figs. 3(a) and 3(b). However, it is included here because its presence can be detected with sensitive measuring equipment of the type described in this paper. Owing to the relatively rapid frequency variation of the factor $(1-\alpha)$ the parameters $\alpha/(1-\alpha)$, $r_c(1-\alpha)$, $C_c/(1-\alpha)$ and $r_b''(1-\alpha)$ become frequency-dependent at a much lower frequency than other parameters not involving the factor $(1-\alpha)$. However, with average transistors the values measured at 1 kc are usually applicable up to high audio frequencies.

For a point-contact transistor or for a grounded-base junction transistor operating at frequencies so low that $C_c$ can be ignored, the impedance parameters are essentially resistive and are given by the simple relationships

$$z_{11} = r_{11} = r_e + r_b \tag{9}$$

$$z_{12} = r_{12} = r_b \tag{10}$$

$$z_{21} = r_{21} = r_m + r_b \approx r_m \tag{11}$$

$$z_{22} = r_{22} = r_c + r_b \approx r_c. \tag{12}$$

It may also be seen that

$$a_{21} = \alpha_{ce} = r_{21}/r_{22} \approx \alpha. \tag{13}$$

On the other hand, when collector capacitance is taken into account the impedance parameters become complex quantities whose resistive and reactive components are all frequency dependent. This makes the impedance parameters cumbersome to apply and it is desirable to seek some form of simplification. For the purposes of this paper, simplification has been achieved by using the equivalent circuit of Fig. 4 which employs a notation suitable for either the grounded-base or grounded-emitter connection.

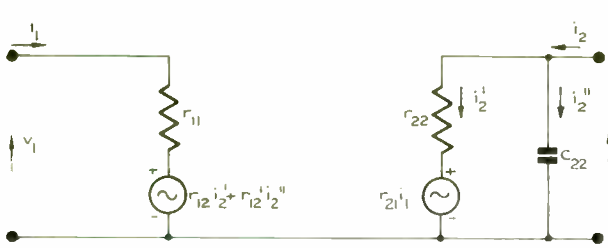


Fig. 4—Four-pole equivalent circuit using modified impedance parameters.

This circuit is derived by considering first the operation at very low frequencies where only the parameters $r_{11}$, $r_{12}$, $r_{21}$, and $r_{22}$ play a part. It is then observed that so long as $1/\omega C_c \gg r_b'$ or $(1-\alpha)/\omega C_c \gg r_e'$ the collector capacitance can be taken into account by adding a capacitance $C_{22}$ across $r_{22}$ and by introducing an additional component of feedback voltage $r_{12}'i_2''$ resulting from the flow of quadrative current through a resistance $r_{12}'$. Here $r_{12}' = r_b'$ for the grounded-base connection and $r_{12}' = r_e'$ for the grounded emitter connection.

Summarizing the parameter values applicable to Fig. 4, we may write for the grounded base connection (adding the appropriate subscript $b$)

$$r_{11b} = r_e + r_b \tag{14}$$

$$r_{12b} = r_b \tag{15}$$

$$r_{12b}' = r_b' \tag{16}$$

$$r_{21b} = r_m + r_b \approx r_m \tag{17}$$

$$r_{22b} = r_c + r_b \approx r_c \tag{18}$$

$$C_{22b} = C_c. \tag{19}$$

For the grounded emitter connection

$$r_{11e} = r_e + r_b \tag{20}$$

$$r_{12e} = r_e \tag{21}$$

$$r_{12e}' = r_e' \tag{22}$$

$$r_{21e} = r_m + r_e \approx r_m \tag{23}$$

$$r_{22e} = r_c(1-\alpha) + r_e \approx r_c(1-\alpha) \tag{24}$$

$$C_{22b} = C_c/(1-\alpha). \tag{25}$$

This simplification of junction-transistor impedance parameters is achieved at the expense of dealing separately with the in-phase and quadrature components of the electrode currents and voltages, but such procedures are standard in bridge measurements.

### *h Parameters*

The use of $i_1$ and $v_2$ as independent variables leads to the relationships

$$v_1 = h_{11}i_1 + h_{12}v_2 \tag{26}$$

$$i_2 = h_{21}i_1 + h_{22}v_2. \tag{27}$$

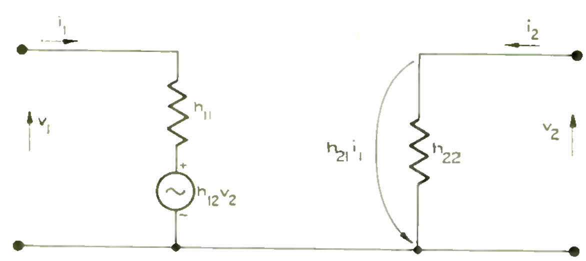


Fig. 5—Four-pole equivalent circuit using $h$ parameters.

These relationships correspond to the equivalent circuit of Fig. 5, and it may also be seen that

$$h_{11} = v_1/i_1 \mid v_2 = 0 \tag{28}$$

$$h_{12} = v_1/v_2 \mid i_1 = 0 \tag{29}$$

$$h_{21} = i_2/i_1 \mid v_2 = 0 \tag{30}$$

$$h_{22} = i_2/v_2 \mid i_1 = 0. \tag{31}$$

The $h$ parameters have the advantage of including the important current gain factor $h_{21}$ $(= -\alpha_{21})$ explicitly, and when dealing with junction transistors they have the further advantage as compared with the impedance parameters of not requiring the measuring apparatus to provide an open-circuit collector termination $i_{2=0}$. This latter requirement is a difficult although not insurmountable one, in view of the high collector resistance of junction transistors.

The parameters $h_{11}$ and $h_{21}$ are virtually unaffected by the presence of collector capacitance since they are measured with the collector shorted. For the grounded-base connection it may be shown that

$$h_{11b} = r_e + r_b(1 - \alpha) \tag{32}$$

$$h_{21b} = - a_{ce} \approx -\alpha. \tag{33}$$

For the grounded-emitter connection

$$h_{11e} = r_b + r_e/(1 - \alpha) \tag{34}$$

$$h_{21e} = - \alpha_{cb} \approx \alpha/1 - \alpha. \tag{35}$$

Thus these parameters are virtually pure resistances except in so far as they are affected by the frequency variation of $\alpha$.

The parameters $h_{12}$ and $h_{22}$ can be conveniently expressed in terms of the parameters $r_{12}$, $r_{12}'$, $r_{22}$, and $C_{22}$, viz.

$$h_{12} = r_{12}/r_{22} + j\omega C_{22}r_{12}' \tag{36}$$

$$h_{22} = 1/r_{22} + j\omega C_{22}. \tag{37}$$

In view of the general suitability of the $h$ parameters for application to junction transistors, the utility of the special treatment of impedance parameters shown in Fig. 4 may be questioned. However, the author is of the opinion that for general laboratory use it is more convenient to deal with the resistance $r_{22}$ than the conductance $1/r_{22}$. Similarly the parameters $r_{12}$ and $r_{12}'$ are more readily visualized than the complex voltage ratio $h_{12}$. Consequently a compromise proposal has been adopted here for junction transistors in which the parameters $h_{11}$ and $h_{21}$ are directly measured, but instead of $h_{12}$ and $h_{22}$ the quantities $r_{12}$, $r_{12}'$, $r_{22}$, and $C_{22}$ are measured. Where the complete set of $h$ parameters is needed, as for instance in network calculations, the values of $h_{12}$ and $h_{22}$ can be easily calculated from (36) and (37).

The bridge if necessary could be modified to read $h_{22}$ and $h_{12}$ directly, but so far the need has not seemed great enough to warrant the additional circuit complications. For point-contact transistors the resistance parameters and the $\alpha$ (i.e., $h_{21}$) parameter are normally measured although $h_{11}$ can be measured, if necessary, so long as its value is positive.

### *Estimation of $\alpha$ Cut-off Frequency*

It may be shown that the frequency variation of $\alpha$ (i.e., $\alpha_{ce}$) can be described by the relationship

$$\alpha = \frac{\alpha_0}{1 + j\kappa\omega/\omega_\alpha}, \quad \omega \ll \omega_\alpha, \tag{38}$$

where $\alpha_0$ = low frequency value of $\alpha$, $\omega_\alpha$ = cut-off angular frequency, i.e., the frequency at which $|\alpha|\alpha_0| = 0.707$, and $\kappa$ is a constant whose value is determined by the nature of the physical process giving rise to the $\alpha$ cutoff. Usually the cutoff is determined by the dispersion in transit time of minority carriers crossing the base, in which case $\kappa$ has a theoretical value of 1.21.

At low frequencies, (38) may be rewritten with good accuracy as

$$\alpha \approx \alpha_0(1 - j\kappa\omega/\omega_\alpha), \qquad \omega \ll \omega_\alpha. \tag{39}$$

This relationship in principle offers the possibility of calculating $\omega_\alpha$ by measuring the phase lag $\tan^{-1} \kappa\omega/\omega_\alpha$ at a known low frequency $\omega$. In the case of the present bridge operating at 1 kc, the phase angle is too small to measure with any degree of accuracy. However, the phase angle of $\alpha_{cb}$ is much larger than that of $\alpha_{ce}$, as may be seen from the following analysis:

$$\begin{aligned}\alpha_{cb} &= \frac{\alpha}{1-\alpha}\cdot \\ &= \frac{a_0/(1 + j\kappa\omega/\omega_\alpha)}{1 - \alpha_0/(1 + j\kappa\omega/\omega_\alpha)}, \\ &= \frac{\alpha_0}{1 - \alpha_0 + j\kappa\omega/\omega_\alpha} && (40) \\ &\approx \frac{\alpha_0}{1-\alpha_0}\left\{1 - j\kappa\omega/\omega_\alpha(1-\alpha_0)\right\}, \omega \ll \omega_\alpha(1-\alpha_0). && (41)\end{aligned}$$

This shows that $\alpha_{cb}$ has a phase lag of $\tan^{-1} \kappa\omega/\omega_\alpha(1-\alpha_0)$, which is very much greater than that of $\alpha_{ce}$, owing to the presence of the factor $1/1-\alpha_0$.

It has been found possible to measure this phase lag with good accuracy at 1 kc, and the values of $\omega_\alpha$ calculated from the phase lag with $\kappa = 1.21$ are found to agree fairly closely with the directly-measured values of $\omega_\alpha$.

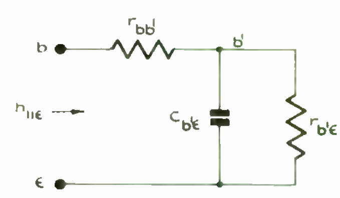


Fig. 6—Equivalent circuit for $h_{11e}$.

Mueller and Pankove[9] using equipment developed by Giacoletto[8] have described an alternative method of calculating $\omega_\alpha$ from bridge measurements on junction transistors. They show that the parameter $h_{11e}$ can be represented by the network of Fig. 6, where the resistance $r_{bb'}$ originates in the base layer ohmic resistance, and the capacitance $C_{b'e}$ and the resistance $r_{b'e}$ result from diffusion of minority carriers in the base. The value

[9] C. W. Mueller and J. L. Pankove, "A *p-n-p* triode alloy junction transistor for radio-frequency amplification," Proc. IRE, vol. 42, pp. 386–391; February, 1954.

of $C_{b'\epsilon}$ is given as $qI\epsilon/kT\omega_\alpha$, where $q/kT=39$ volt$^{-1}$ at room temperature. Hence when $C_{b'\epsilon}$ is known $\omega_\alpha$ can be calculated.

The bridge described here is capable of measuring an equivalent capacitative component of $h_{11\epsilon}$ referred to the base terminal, and the value of $C_{b'\epsilon}$ could then be calculated if $r_{bb'}$ were known. $r_{bb'}$ tends to have a value close to that of $r_b'$, but it would appear that there can be appreciable differences between these two parameters in the case of alloyed-junction transistors. As shown by Giacoletto, the values of $r_{bb'}$, $C_{b'\epsilon}$ and $r_{b'\epsilon}$ can only be determined accurately by a multifrequency test set, and it must therefore be concluded that with the present bridge only a rough value of $\omega_\alpha$ can be deduced from $h_{11\epsilon}$ measurements. This has been verified by actual measurements.

## Principle of Measurement

The bridge circuit is derived from the one used in the General Radio vacuum tube bridge and described by Tuttle.[10] The adopted principle makes the measurement of $h_{21}$ and $h_{11}$ interdependent. Similarly the measurement of $r_{22}$ and $C_{22}$ is linked with the measurement of $r_{12}$ and $r_{12}'$, while the measurement of $r_{11}$ is linked with the measurement of $r_{21}$.

### *Measurement of $h_{21}$ and $h_{11}$*

Fig. 7 shows the basic circuit for measuring these parameters. In this and the following circuits the voltages $e_1$, $e_2$, and $e_3$ are small 1,000 cps voltages derived from low-impedance attenuators connected to a number of secondary windings on an input transformer, the input voltages to the attenuators being carefully equalized. The voltages $e_1$ and $e_2$ are exactly in phase with each other but the voltage $e_3$ has an adjustable phase angle. Accordingly we may write

$$e_1 = k_1e_0 \tag{42}$$

$$e_2 = k_2e_0 \tag{43}$$

$$e_3 = k_3(1 + j\beta)e_0, \tag{44}$$

where $k_1$, $k_2$, and $k_3$ are attenuation factors, $e_0$ is the attenuator input voltage, and $\tan^{-1}\beta$ is the phase angle of $e_3$. For reasons which will become apparent later, it is convenient to make $k_2$ and $k_3$ continuously adjustable while $k_1$ is adjustable by factors of ten, i.e., it takes the values 1.0, 0.1, 0.01 etc. It will be noticed that the voltage $e_2$ appears twice in Fig. 7. This is arranged by the use of a 1:1 transformer.

The resistance $R_1$ is made at least a hundred times greater than the input impedance so that the input current can be taken as $i_1=e_1/R_1$ within 1 per cent accuracy. Emitter or base bias current is obtained by connecting a suitably by-passed dc supply to the "low" end of $e_1$. Collector bias is applied through the primary of the detector transformer $T_1$. The latter is followed by a high-gain detector amplifier.

The polarity of the voltages shown in Fig. 7 is suitable for positive values of $h_{21}$, i.e., for the grounded emitter connection. For the grounded base connection, $e_2$ must be reversed.

For a detector null to be obtained ($i_D=0$) it is seen that

$$i_2 = i_R + i_Q. \tag{45}$$

At the same time $v_2=0$ so that

$$i_2 = h_{21}i_1 = h_{21}e_1/R_1. \tag{46}$$

Also $i_R=e_2/R_2$ and $i_Q=-j\omega Ce_2$, whence

$$h_{21} = \frac{R_1e_2}{R_2e_1}(1 - j\omega CR_2) \tag{47}$$

$$= \frac{R_1k_2}{R_2k_1}(1 - j\omega CR_2). \tag{48}$$

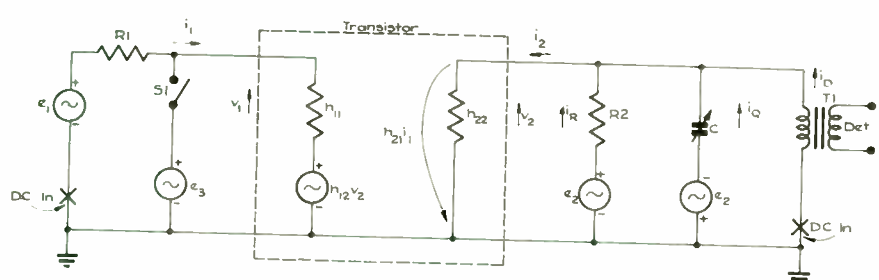


Fig. 7—Basic circuit for measuring $h_{21}$ and $h_{11}$.

Thus the real part of $h_{21}$ can be read directly from the setting of the $k_2$ attenuator multiplied by the factor $R_1/R_2k_1$ which can be arranged to be a power of ten. For the grounded base connection the real part of $h_{21}$, i.e., $\alpha_{ce}$, is the only significant term.

In the grounded emitter case it was shown in (41) that $h_{21}$, i.e., $\alpha_{cb}$, has a significant phase lag equal to $\tan^{-1} \kappa\omega/\omega_\alpha(1-\alpha_0)$. Hence equating this to the phase lag of (48) yields

$$\kappa\omega/\omega_\alpha(1 - \alpha_0) = \omega CR_2 \tag{49}$$

whence

$$\omega_\alpha = \kappa/(1 - \alpha_0)CR_2. \tag{50}$$

[10] W. N. Tuttle, "Dynamic measurement of electron tube coefficients," Proc. IRE, vol. 21, pp. 844–857; June, 1933.

Once a null balance has been established for the $h_{21}$ measurement the transistor input impedance is by definition $h_{11}$. A voltage $h_{11}i_1$ is then developed at the transistor input terminal and this can be measured by closing $S_1$ and adjusting $e_3$ to restore the null balance. When this is done we have

$$e_3 = h_{11}i_1 \tag{51}$$

or

$$h_{11} = R_1 e_3/e_1 \tag{52}$$

$$= R_1 k_3(1 + j\beta)/k_1. \tag{53}$$

Thus the real part of $h_{11}$ is given directly by the attenuating factor $k_3$ multiplied by the factor $R_1/k_1$ which can be arranged to be a power of ten. For the grounded base connection, $h_{11b}$ is essentially resistive and $\beta$ is quite small.[11] As mentioned earlier, $h_{11e}$ has a substantial capacitative component so that negative values of $\beta$ must be provided to obtain a balance.

### *Measurement of $r_{22}$, $C_{22}$, $r_{12}$ and $r_{12}'$*

The method of measurement is illustrated in Fig. 8. It will be appreciated that for point-contact transistors a simpler analysis applies in which $C_{22}$ and $r_{12}'$ are ignored.

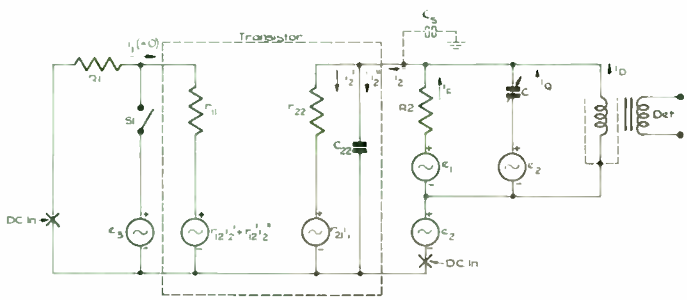


Fig. 8—Basic circuit for measuring $r_{22}$, $C_{22}$, $r_{12}$, and $r_{12}'$.

Here the resistance $R_1$, through which the input bias current is fed, is made large enough to represent an effective ac open circuit at the input terminal. When a null is achieved with this circuit the collector voltage must be equal to $e_2$ and $i_D = 0$ so that

$$i_R + i_Q = i_2' + i_2'', \tag{54}$$

whence

$$e_1/R_2 + j\omega C e_2 = e_2/r_{22} + j\omega C_{22} e_2. \tag{55}$$

Equating in-phase and quadrature components in this equation yields

$$r_{22} = R_2 e_2/e_1 = R_2 k_2/k_1 \tag{56}$$

and

$$C_{22} = C. \tag{57}$$

[11] Theoretically $h_{11b}$ can be either slightly inductive or slightly capacitative.

This shows that $r_{22}$ can be read from the $k_2$ attenuator multiplied by the factor $R_2/k_1$. In addition, $C_{22}$ is obtained by calibrating the variable capacitor $C$.

The foregoing analysis has neglected the effect of stray capacitance. It is found in practice that only the stray capacitance $C_s$ shown dotted in Fig. 8 has any significant effect. This capacitance appears at a high impedance point and, in effect, augments the collector capacitance. Stray capacitances at other points are shunted across low impedances and can be ignored. The magnitude of $C_s$ can be determined by placing a resistance of known self-capacitance (determined, say, by a $Q$-meter) between the collector terminal and ground. The setting of the quadrature balancing capacitor $C$ when the bridge is in balance then gives a measure of $C_s$ which can be deducted from subsequent measurements with a transistor in circuit. It is to be noted that $C_s$ does not influence the $\alpha$ measuring circuit since the collector ac voltage is then zero.

In measuring $r_{12}$ and $r_{12}'$ a null is first established in respect of $r_{22}$ and $C_{22}$. $S_1$ is then closed and $e_3$ is adjusted in magnitude and phase to restore the null. In this case

$$e_3 = r_{12}i_2' + r_{12}'i_2'' \tag{58}$$

$$= r_{12}e_1/R_2 + j\omega C_{22}e_2 r_{12}', \tag{59}$$

whence

$$k_3(1 + j\beta) = r_{12}k_1/R_2 + j\omega C_{22}k_2 r_{12}'. \tag{60}$$

Equating in-phase and quadrature components yields

$$r_{12} = R_2 k_3/k_1, \tag{61}$$

$$r_{12}' = \beta k_3/\omega C_{22}k_2, \tag{62}$$

or, alternatively,

$$r_{12}'/r_{12} = \beta/\omega C_{22}r_{22}. \tag{63}$$

In this case $r_{12}$ can be read directly from the $k_3$ attenuator multiplied by the factor $R_2/k_1$. On the other hand, the value of $r_{12}'$ is not directly indicated but must be calculated from (62) or (63).

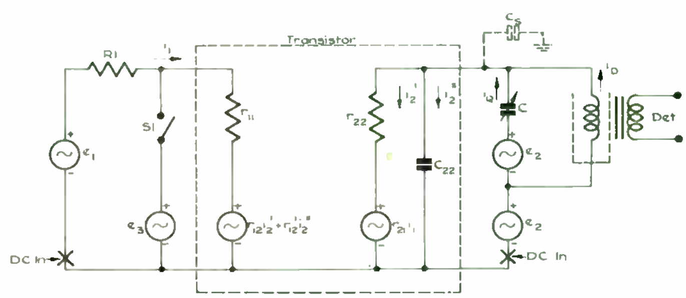


Fig. 9—Basic circuit for measuring $r_{21}$ and $r_{11}$.

### *Measurement of $r_{21}$ and $r_{11}$*

The basic circuit for measuring these parameters is shown in Fig. 9 and is illustrated for the sake of generality with regard to a junction transistor, although such measurements are usually only performed on point-contact transistors. Once again, $R_1$ is made very much greater than $r_{11}$ so that $i_1 = e_1/R_1$. When a null is established on the collector side, the in-phase component of collector current is reduced to zero; $i_2' = 0$. Hence

$$r_{21}i_1 = e_2 \tag{64}$$

and

$$i_2'' = j\omega C_{22}e_2 = j\omega Ce_2, \tag{65}$$

whence

$$r_{21} = e_2R_1/e_1 = k_2R_1/k_1 \tag{66}$$

and

$$C_{22} = C. \tag{67}$$

Thus $r_{21}$ is read from the bridge in very much the same way as $r_{22}$, and the capacitance balancing conditions, including the effects of $C_s$, are identical with those discussed in the previous section.

In measuring $r_{11}$, the null is first established on the collector side, $S_1$ is closed and $e_3$ is adjusted to restore the null balance. Since $i_2' = 0$ in this case,

$$e_3 = r_{11}i_1 + r_{12}'i_2'', \tag{68}$$

or

$$k_3(1 + j\beta)e_0 = r_{11}k_1e_0/R_1 + j\omega C_{22}\kappa_2e_0r_{12}'. \tag{69}$$

Whence, equating the in-phase terms,

$$r_{11} = \kappa_3R_1/\kappa_1. \tag{70}$$

Thus $r_{11}$ is read with the same scale factors as $r_{12}$.

It may also be seen that the quadrature terms of (69) offer the possibility of measuring $r_{12}'$, but since this has already been done it need not be considered any further.

In all three basic bridge circuits there is an element of indirectness in the measurements carried out on the input side of the transistor, since they depend on a prior balance being performed on the output side. However this method has the advantage that in Fig. 9 the residual in-phase component of the collector current can be made as low as $10^{-10}$ amps, i.e., comparable with the transistor noise current. This is equivalent to maintaining an effective ac impedance of thousands of megohms at the collector, a value which would be impossible to achieve with more conventional methods. Furthermore, the method of Fig. 8 allows accurately known currents as low as $10^{-8}$ amps to be passed into the collector terminal for the purpose of measuring $r_{12}$ and $r_{12}'$. Other more direct methods of carrying out measurements on the input side of the transistor would require the detector to be transferred to that side. The present arrangement which keeps the detector on the output side results in a minimum of switching and allows the power gain of the transistor to be used to improve the balance sensitivity.

## Complete Circuit

The complete circuit embodying the basic circuits of Figs. 7, 8 and 9 is shown in Fig. 10 (opposite). Alternating current at 1 kc is applied to the bridge attenuators through a transformer $T_1$ having a 20:1 step-down ratio between the primary and each secondary. The primary is driven by a cathode-follower in order to keep the secondary output impedance down to a few ohms. The secondary voltage is adjustable up to a maximum value of about 1 volt. This transformer should be designed to have equal secondary leakage reactances so that the secondary voltages are exactly in phase. In the experimental transformer used here the middle secondary was found to have a slightly higher leakage reactance than the outer two, and tne loading arrangement shown in Fig. 10 which places approximately 1,100 ohms on the middle secondary and approximately 100 ohms on the outer secondaries was found necessary in order to bring the secondary currents exactly into phase with each other. Finally the small trimming resistors $R_5$ and $R_{10}$ were added to make the full-scale attenuator voltages equal within $\pm 0.1$ per cent. These adjustments were carried out by performing an $\alpha$ measurement with the emitter and collector terminals shorted together. In this case if $R_1$ and $R_2$ are carefully equalized and the quadrature balancing capacitor is disconnected, $e_1$ and $e_2$ must be exactly equal in amplitude and phase in order to obtain a null. The bridge should then, of course, register an $\alpha$ of unity. A check on $e_3$ can then be made by connecting it temporarily in place of $e_1$.

The attenuator controlling $e_1$ takes the form of a simple voltage divider which yields values of $k_1$ equal to 1.0, 0.1, 0.01, or 0.001. Its output impedance is 100 ohms or less depending on the setting used. The attenuator controlling $e_2$ has a coarse and fine control which give a setting accuracy of $\pm 0.1$ per cent of the full-scale value. Nine 10-ohm resistors $R_{3A \ldots I}$ and the 10.5-ohm potentiometer $R_4$ are used here, the value of 10.5 ohms being adopted to give a small overlap between the coarse steps. For the sake of smoothness $R_4$ has been constructed in the form of a single-turn slide wire potentiometer. A switch $S7$ is provided to reverse the polarity of $e_2$ wherever required.

The method of collector capacitance balancing shown in Fig. 10 represents an improvement over the simple scheme described earlier and follows the technique adopted in the General Radio bridge. It will be seen that two voltages $\pm e_2$ are generated with the aid of the center-tapped autotransformer $T_2$[12] and are applied to the stators of the differential variable capacitor $C_1$ whose rotor is connected to the collector terminal. With this arrangement the effective balancing capacitance can take positive or negative values and can be varied smoothly through zero effective value. This avoids the limitations which would be imposed by the minimum capacitance of a single-ended variable capacitor. Often it is found that the built-in capacitor is not large enough to secure a balance, and in this case an external

[12] This transformer should have interleaved windings to minimize its leakage reactance. The best construction might possibly employ a bifilar winding.

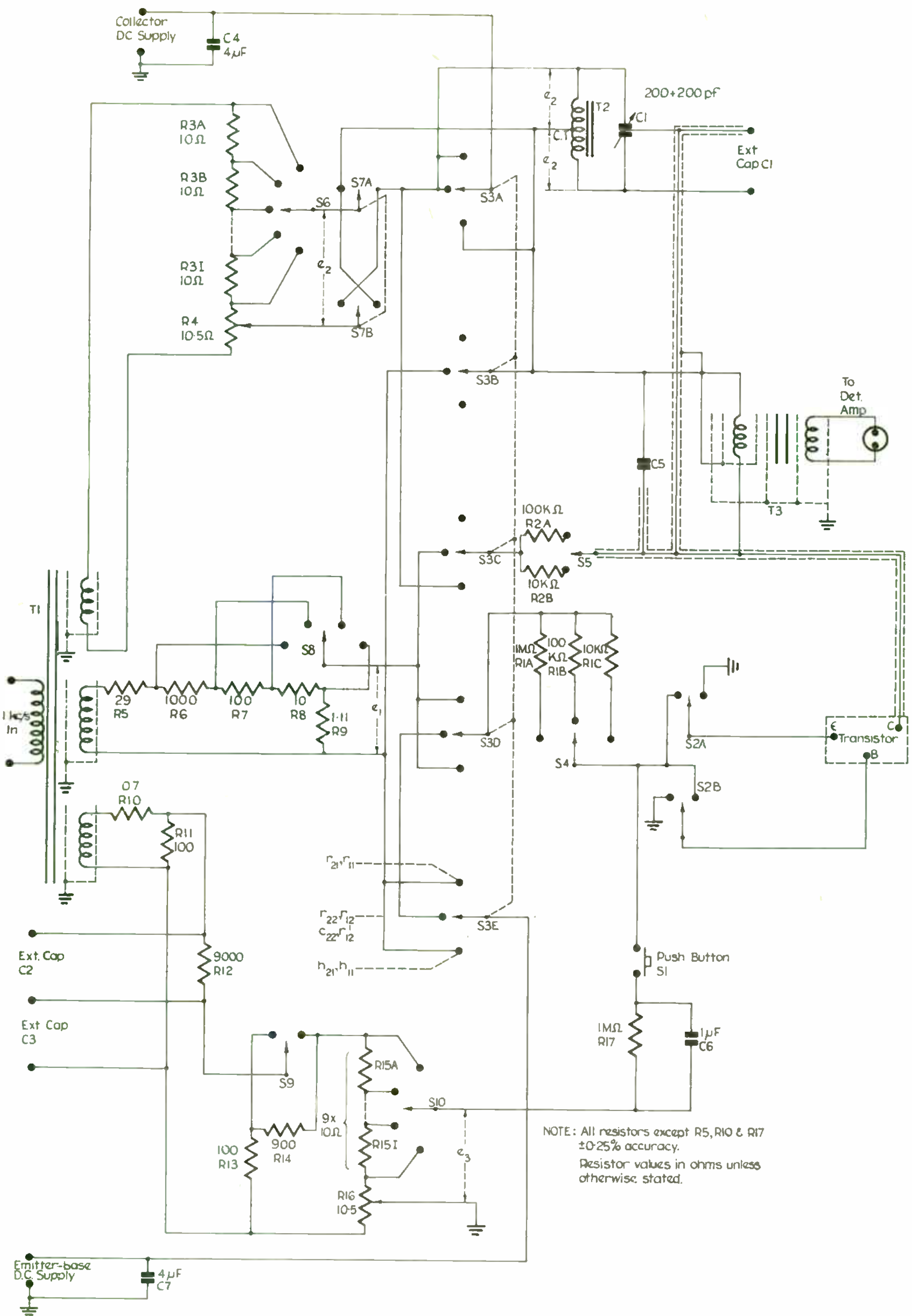


Fig. 10—Complete circuit of transistor bridge.

variable capacitor is connected to the terminals marked "Ext. Cap. $C_1$." Alternatively the external capacitor can be used to obtain greater precision than the internal capacitor is capable of providing.

The attenuator controlling $e_3$ includes the coarse and fine potentiometer system $R_{15A} \cdots R_{15I}$ and $R_{16}$ which are identical with $R_{3A} \cdots R_{3I}$ and $R_4$. These are preceded by a network including $R_{12}$, $R_{13}$, $R_{14}$ and $S_9$ which is arranged to provide two ranges of $\kappa_3$, namely 0 to 0.01 and 0 to 0.001. When the phase angle of $e_3$ requires adjusting, an external decade capacitor is introduced in the position marked $C_2$ or $C_3$. $C_2$ is used when the phase angle of $e_3$ must be advanced, as when measuring $r_{12}'$, and $C_3$ is used when measuring $h_{11e}$. Analysis shows that the magnitude of the quantity $\beta$ appearing in (44) is given to good approximation by relationship

$$\beta = \omega C_2 R_{12} \qquad (|\beta| < 1). \quad (71)$$

In this case (63) can be rearranged to the form,

$$r_{12}'/r_{12} = R_{12}C_2/r_{22}C_{22}. \quad (72)$$

Further analysis shows that for $|\beta| > 1$ (which occasionally occurs) the value of $r_{12}$, as read from the bridge, should be increased by a factor $(1+.01\beta^2)$. This is because the presence of $C_2$ causes the in-phase component of $e_3$ to increase slightly. Notice that this arrangement for shifting the phase angle of $e_3$ is an improvisation which might well be replaced by some other method giving a more direct indication of the value of $r_{12}'$.

The voltage $e_3$ is coupled to the transistor input terminal through the push button switch $S_1$ and a 1 $\mu F$ capacitor $C_6$ which is shunted by a 1-megohm resistor in order to prevent the accidental accumulation of charge on $C_6$.

Three values are provided for the resistance $R_1$, namely 1 megohm, 100 kilohms and 10 kilohms. The first value is used when working with small input bias currents, when at the same time the transistor input impedance tends to be high. The second value is used for moderate values of bias current, i.e., up to 3 ma with a 300-volt dc supply. The third value is used for occasional measurements at higher bias currents. With this value of $R_1$ it may be necessary to make corrections to the measured parameter value because the transistor input impedance may be greater than $R_1/100$.

For the resistance $R_2$ alternative values of 100 $\kappa\Omega$ or 10 $\kappa\Omega$ are provided, the former value belng most frequently used. The 10,000-ohm value sometimes permits a higher signal current level to be achieved with a consequent sharpening of the null point.

The detector transformer is specially wound with a double-shielded primary, the inner shields being connected to the "low" end of the primary to secure a guard-ring effect. The primary is parallel resonated by the capacitor $C_5$ at 1 kc to improve the detection sensitivity and at the same time to provide some discrimination against hum pickup and transistor noise. By airgapping the transformer core, the effects of dc saturation are minimized and the $Q$ at 1 kc is improved. Induced voltages resulting from stray fields are cut down below the limits of detection by proper orientation of the transformer and by enclosing it in a mumetal shield.

Fig. 11 is a photograph showing the layout of controls on the panel of the bridge.

### Performance Data

With the component values shown in Fig. 10 values of $\alpha$ up to 1,000 and values of $r_{22}$ or $r_{21}$ up to 100 megohms can be measured. The range of values of $h_{11}$, $r_{11}$, and $r_{12}$ which can be measured depends on the type of transistor and the connection being used but is adequate for all normal purposes. Experience has shown, however, that it would be advantageous to have an additional attenuating position on $S_9$ yielding a 0–0.0001 range of values of $\kappa_3$. This would improve the accuracy when measuring small values of $r_{12}$ in association with large values of $r_{22}$ (i.e., $h_{12}$ very small).

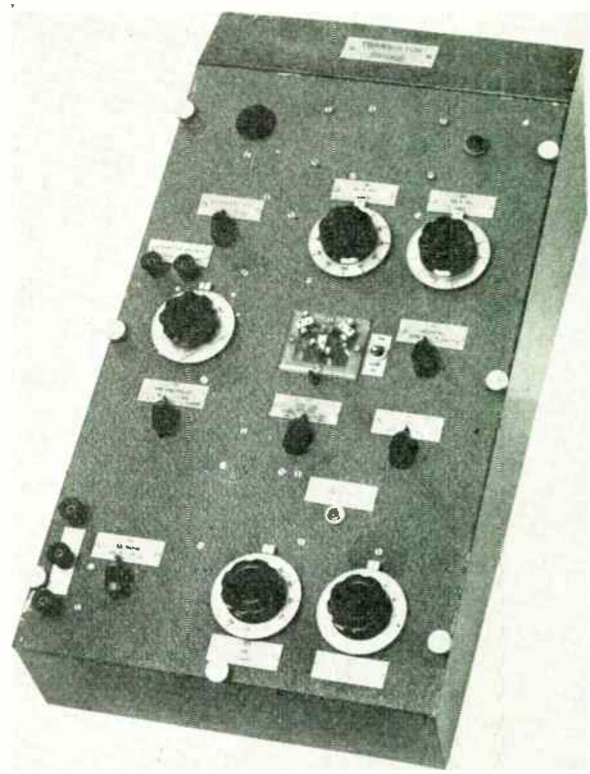

Fig. 11—Photograph of transistor bridge.

In checking the accuracy of the bridge, certain tests can be carried out such as an $\alpha$ measurement with the emitter and collector terminals shorted which, as mentioned earlier, should give an indicated $\alpha$ of unity. A comprehensive check on $r_{22}$ and $C_{22}$ measurements can be carried out with the aid of precision resistors and capacitors, and a similar check can be devised for $r_{11}$ and $h_{11}$ measurements by connecting a precision resistor between the emitter and base terminals and using a vacuum tube to supply an amplifying link between the emitter and collector terminals. These tests show that where the transistor noise level is low enough to yield a sharp null point the resistance measurements should be accurate within 1 or 2 per cent except where the parameters take extremely high or low values. $C_{22}$ measurements are most accurate when associated with high values of $r_{22}$, in which case an accuracy of the order of $\pm 2$ per cent $\pm 1$ $pF$ is obtainable. High values of $r_{22}$ need to be corrected for the inherent leakage resistance of the bridge which can be measured with the collector terminal open-circuited.

The bridge is capable of measuring parameters over a wide range of biasing conditions extending down to as little as 0.1 volt collector bias and 1 $\mu a$ emitter current.

Under these conditions special precautions must be taken to ensure that signal voltages and currents are small in comparison with bias values without at the same time allowing the signal to be obscured by noise. The best compromise is achieved by increasing the signal level to the point at which the null setting just begins to shift owing to curvature of the transistor characteristics.

The correlation between the $\alpha$ cut-off frequency as calculated from the bridge measurements and the directly measured value has been found to be best with alloyed-junction transistors. In this regard a test was carried out with 24 alloyed-junction transistors, mostly type OC70, OC71, and some CK722's. The value of $\kappa$ calculated from (50) as an empirical constant to harmonize the measured $\alpha$ cut-off frequency with the 1-kc phase measurements worked out as $\kappa = 1.17$ with a standard deviation of $\pm 0.05$. The difference between the measured value of 1.17 and the theoretical value of 1.21 is not considered significant in view of the spread of measurements and the possibility of small systematic errors.

In the case of grown-junction transistors, of which relatively few have been available to the author, apparent values of $\kappa$ ranging from 0.7 to 1.4 have been determined; indicating that the factors determining the cut-off frequency of these transistors are more complex than those applying to alloyed-junction transistors. On the whole it is felt that, in the absence of an "$\alpha$ sweeper" or other convenient $\alpha$ cut-off measuring equipment, the feature of being able to calculate from the bridge measurements a fairly accurate $\alpha$ cut-off frequency for alloyed-junction transistors should prove quite useful.

Values of $r_b'$ measured on the bridge are found to fall within $\pm 10$ per cent of the values measured directly at high frequencies and in this regard it may be of interest to note that with grown-junction transistors, values of $r_b'$ greater than $r_b$ are often measured. Here the value of $r_b''$ is negative, an effect which Early[13] has shown to be due to "base-resistance modulation." In alloyed-junction transistors $r_b$ is always several times greater than $r_b'$ for moderate values of emitter current. A corollary of these remarks in regard to $r_b$ and $r_b'$ is that $r_e'$ is greater than $r_e$ except when $r_b''$ is negative.


## Acknowledgment

Thanks are due to Mr. C. D. Howarth, who attended to the constructional details of the bridge.


[13] J. M. Early, "Design theory of junction transistors," *Bell Sys. Tech. Jour.*, vol. 32, pp. 1271–1312; November, 1953.

# Skin Resistance of a Transmission-Line Conductor of Polygon Cross Section*

HAROLD A. WHEELER†, FELLOW, IRE

*Summary*—**If a conductor cross section is any straight-sided polygon that can be circumscribed on a circle, it is found to have the same skin resistance as a conductor whose cross section is this circle. For example, a square wire has the same resistance as a round wire of the same radius, though the square perimeter is $4/\pi$ times as great. This "polygon rule" is derived from the "incremental-inductance rule" of the skin effect, published in 1942. It applies equally to inner or outer conductors, though the current distribution is very different. It applies to some unusual shapes that are difficult to compute by any other method.‡**

THE SKIN EFFECT is the well-known phenomenon of high-frequency currents concentrating just under the exposed surfaces of a conductor. The effective resistance of a cylindrical conductor may be computed by integrating the power losses of such currents associated with its surfaces. This has been done for cylindrical conductors of some simple cross sections, and more laboriously for various polygon cross sections.

The theorem to be presented gives a very simple evaluation of this resistance for a certain category of polygon cross sections. It applies to all polygons that can be circumscribed on a circle. Two conductors of the same material have the same skin resistance if their cross sections are respectively a circle and a polygon of such size and shape that it can be circumscribed on this circle. This rule applies to inner or outer conductors bounding a high-frequency magnetic field. A special case of the former is an antenna wire. The conditions of validity will be specified in more detail.

It is essential to distinguish between the skin resistance and the inductive reactance of a conductor, because they are obtained by different methods of computation. Conformal mapping can be utilized to obtain directly, for different shapes, the ratio of sizes that will have the same reactance. To this operation must be added the integration of power dissipation over the

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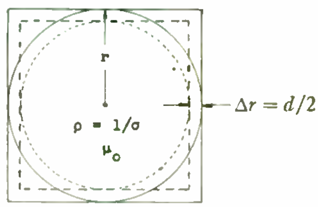
‡ This theorem was discovered by the author shortly after publication of his 1942 paper on the skin effect (Bibliographical reference 2). He has presented it at various meetings, including a seminar at New York University, New York, N. Y., on March 29, 1950, and a staff meeting at Wheeler Laboratories on December 12, 1951.

surface, to obtain the ratio of sizes that will have the same resistance. The present theorem is an implicit solution for the latter in a variety of cases.

For example, a square wire has the same skin resistance as a circular wire of the same diameter but only $\pi/4$ as great a perimeter. This is illustrated in Fig. 1, as a basis for stating the theorem and its proof.




Skin Depth: $d = \dfrac{1}{\sqrt{\pi f \mu_0 \sigma}}$

Surface Resistivity: $R_s = \rho/d = \sqrt{\pi f \mu_0 \rho}$

Skin Resistance: $R = R_s l/2\pi r$

Fig. 1—Cross-sectional circle and circumscribed polygon (square) having equal resistance.

The underlying principle is the "incremental-inductance rule" previously stated by the author.[2,3] Its basis will be restated briefly.

In the idealized skin effect, the current appears to be distributed uniformly in the skin depth ($d$) just under the surface of the conductor. The average depth of the current is then half the skin depth ($d/2$).

In a perfect conductor, current would travel on the surface, since there would be zero penetration. This is the usual assumption in inductance formulas for high frequencies; coaxial line is the most common example.

The depth of penetration effects a proportional increment of inductance, just as if the conductor surface receded by an amount equal to the half-depth ($d/2$). It happens that the reactance of this increment of inductance is equal to the skin resistance, so the latter may be evaluated by computing the former. This is the "incremental-inductance rule."

Referring to Fig. 1, there are shown two cross sections of the same radius, a circle and a circumscribed square. These are given as examples of all the polygons that can be circumscribed on a circle. We wish to compare these two shapes with respect to the increment of inductance caused by reducing the radius ($r$) by a relatively small amount ($\Delta r \ll r$). In each case, this change of size is accompanied by no change of shape.

Here we rely on the well known principle that the change of inductance depends only on the ratio of change of size, if the shape remains the same. The simplest example of this principle is the coaxial line, whose inductance depends only on the shape (determined by the ratio of radii) and not on the size. In Fig. 1, either cross section may represent the inner conductor of a coaxial line whose outer conductor is represented by a concentric circle of much greater radius. If then the radii of both inner and outer conductors are reduced in the same relative amount ($\Delta r/r$), the inductance remains the same. This proves that the change of the inductance of either inner conductor is the same as that of the outer conductor; hence the circle and the circumscribed polygon have the same change of inductance.

From the preceding relationship and the incremental-inductance rule, it is deduced that *the circle and the circumscribed polygon have the same skin resistance.* This theorem is designated, the "polygon rule."

The actual skin resistance may be evaluated by the well-known formulas in Fig. 1; the symbols are defined as follows (MKS rationalized units):

$r$ = radius of circular cross section of conductor
$\Delta r$ = effective reduction of radius by penetration
$\rho$ = resistivity of conductor
$\sigma$ = conductivity of conductor
$\mu_0$ = magnetivity of free space (also in the conductor, assumed nonmagnetic)
$f$ = frequency
$l$ = length of conductor
$d$ = skin depth
$R_s$ = surface resistance of a square (equal length and width)
$R$ = skin resistance of a conductor (of specified radius and length).

As an example of the polygon rule applied to inner and outer conductors, Fig. 2 shows a coaxial line of square cross section. The squares are circumscribed on the circles shown in dotted lines. By the same reasoning, either square has the same skin resistance as its inscribed circle. (With respect to reactance, the pair of squares gives less than the pair of circles, and occupies more space.) Either square may be used with the other circle, since the polygon rule applies independently to each of the two conductors.

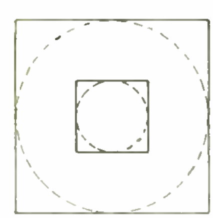


Fig. 2—Concentric circles, and circumscribed squares having equal resistance (but less reactance in greater space).

In this and further cases, the equality of resistance for the circle and the circumscribed polygon is realized to the extent that the following conditions are approximated. (The first of these is the usual assumption for the simple formulas of the skin effect.)

## Conditions

1. The skin depth is much less than $\frac{1}{4}$ the thickness of any substantial part of the conductor. (At the angles of a polygon, this condition is met if the skin depth is a very small fraction of the radius of the inscribed circle, as is true in many applications.)

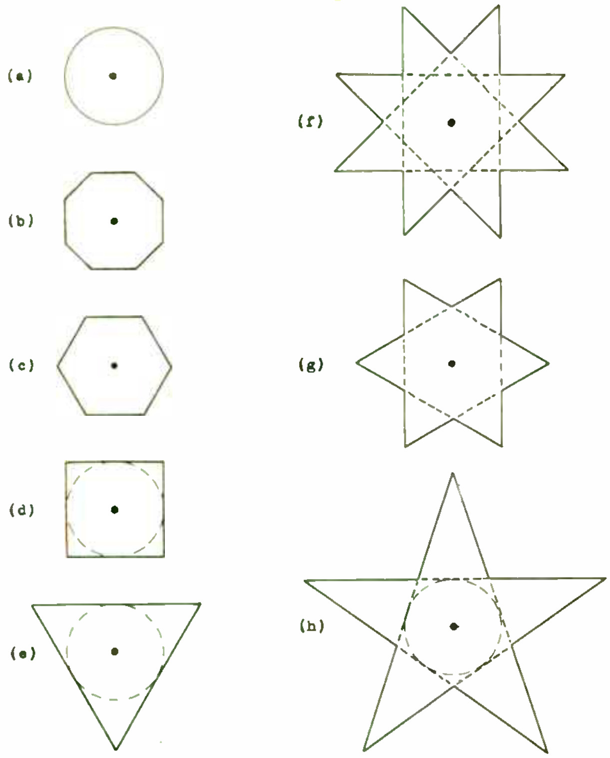



Fig. 3—Regular polygons circumscribed on equal circles, examples of inner or outer conductors.

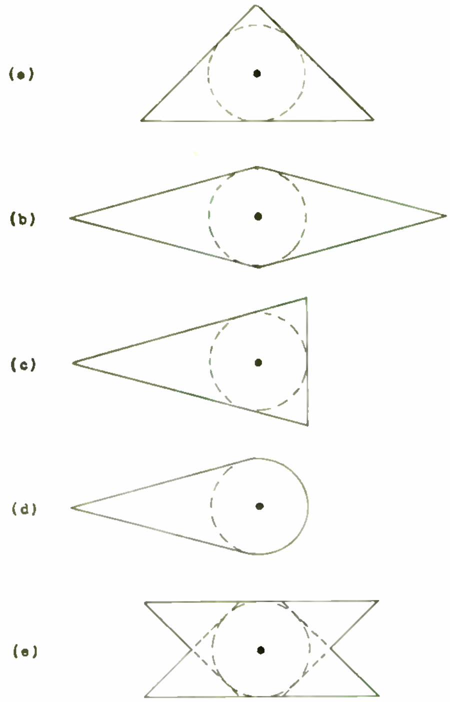



Fig. 4—Elongated polygons circumscribed on equal circles, examples of inner or outer conductors.

2. Outside of an "inner" conductor (or isolated wire) any other conductors affecting the field pattern are spaced at a distance sufficient to provide that they have a negligible effect on the current distribution.
3. Inside of an "outer" conductor, the inner conductor is near the center of the inscribed circle and is small enough to provide that the current distribution on the outer conductor is substantially the same as would be obtained with a fine wire in the center.

Commenting on the second and third conditions applied to a coaxial line, the polygon rule requires that the ratio of radii be great enough to prevent either conductor causing appreciable distortion of the field at the other conductor. The ratio of radii may be closer to unity as either or both of the conductors assumes a higher order of symmetry or regularity.

Figs. 3 and 4 show examples of regular polygons and elongated polygons circumscribed on equal circles and therefore having the same values of skin resistance. Every one is closed and hence may serve as inner or outer conductor; if the latter, the associated inner conductor is centered in the circle.

Fig. 5 shows examples of right polygons that have some sides open. Every open conductor is assumed to extend outward as far as the field is appreciable. Therefore these are suitable only for outer conductors.

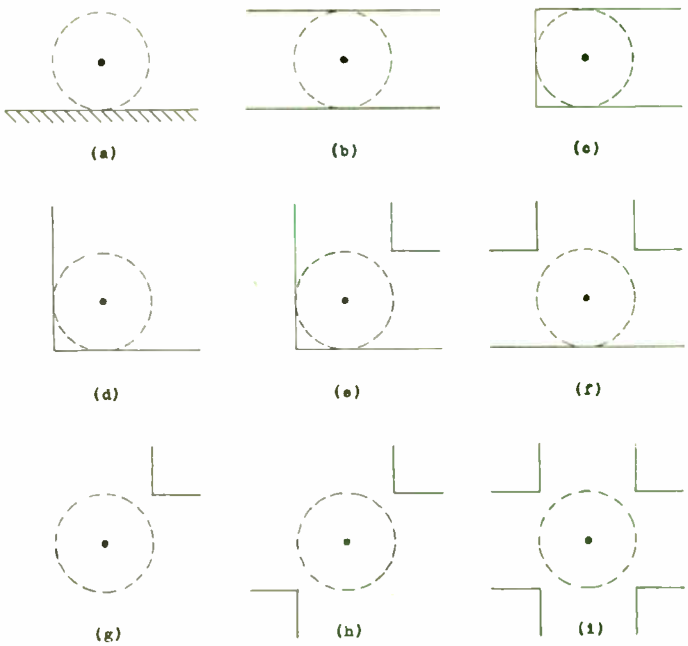



Fig. 5—Right polygons with some sides open, examples of outer conductors.

The basic requirement of this category of polygons is simple. They comprise every cross-sectional contour that retains the same shape if the entire surface recedes by a specified small amount. It appears that this category includes all polygons formed of straight lines tangent to a circle, and excludes all other shapes. The circle is one limiting case. Every one of these contours presents the same skin resistance if circumscribed on the same size of circle.

With respect to all inner conductors, the contours with more acute outer angles make less effective use of their perimeters, because there is more extreme concentration of current near the outer angles. By analogy with electric potential gradient, the current density is theoretically infinite at any angle whose outer side is exposed to the magnetic field. In spite of this fact, it is noted that the square suffers very little, effectively utilizing $\pi/4$ of its perimeter. An opposite extreme is the elongated rhombus, Fig. 4(b).

Fig. 4(e) as an inner conductor might be expected to approximate the behavior of a rectangle. On the contrary, it has much less effective utilization of its contour because of the acute angles.

Referring to the more extreme shapes in Figs. 3 and 4, their equality of resistance as inner and outer conductors is remarkable and unexpected, because the current distribution is radically different in these alternative functions.

Every example of the polygon rule is an evaluation of a certain definite integral. Some of the more unusual cases may be integrals that cannot be evaluated by any procedure known to mathematicians. A long table could be prepared on the basis of this one rule.

The polygon rule offers a fascinating variety of examples based on a single theorem. It has some practical utility in computing or estimating the skin resistance of inner and outer conductors of various polygon cross sections. Its greatest value lies in the ideas to be perceived in its examples, particularly the effect of extreme current concentration on acute angles exposed to the field. It is another interesting application of the basic "incremental-inductance rule."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# Active-Error Feedback and Its Application to a Specific Driver Circuit*

J. R. MACDONALD†, SENIOR MEMBER, IRE

***Summary***—**A short discussion of the advantages and disadvantages of active-error feedback in amplifier design is given. Such feedback can yield all the advantages of ordinary negative feedback without gain reduction and is particularly suitable for use in reducing the distortion of individual amplifier stages.**

**Active-error feedback is applied to a cathode follower by amplifying the difference between its input and output voltages, then adding the amplified error to the output. The resulting driver has very low output impedance and low distortion and is especially useful for driving the grid of an output tube far positive. A direct-coupled version of the circuit using ordinary miniature tubes had an output impedance of 5.6 ohms and could supply several hundred milliamperes of positive current. The theory of the circuit agrees with experiment, and the distortion of the driver when driving an output tube grid to the diode line is found to be far less than that of an ordinary cathode-follower driver.**

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## I. Active-Error Feedback

Unlike ordinary negative feedback, where a portion of the available gain of an amplifier is expended in obtaining the benefits of feedback, active-error feedback (AEF) is a type of feedback with which no such direct gain reduction occurs. With AEF, a portion of the output signal from an amplifier or single-amplifier stage is subtracted from the input signal, then the resulting difference amplified in an external circuit whose gain is equivalent to the extra gain necessary with ordinary feedback. If the portion of the output subtracted is nominally equal to the input, the difference is proportional to the error or distortion in the output. This error is then injected back into the original circuit with the proper polarity to reduce the output error. Although the principle of AEF has been used in servomechanisms, it does not seem to have been as fully exploited in amplifier design as it deserves to be.[1] Therefore, it is worthwhile to discuss its advantages and disadvantages in this application in some detail and to present a specific example of this type of feedback.

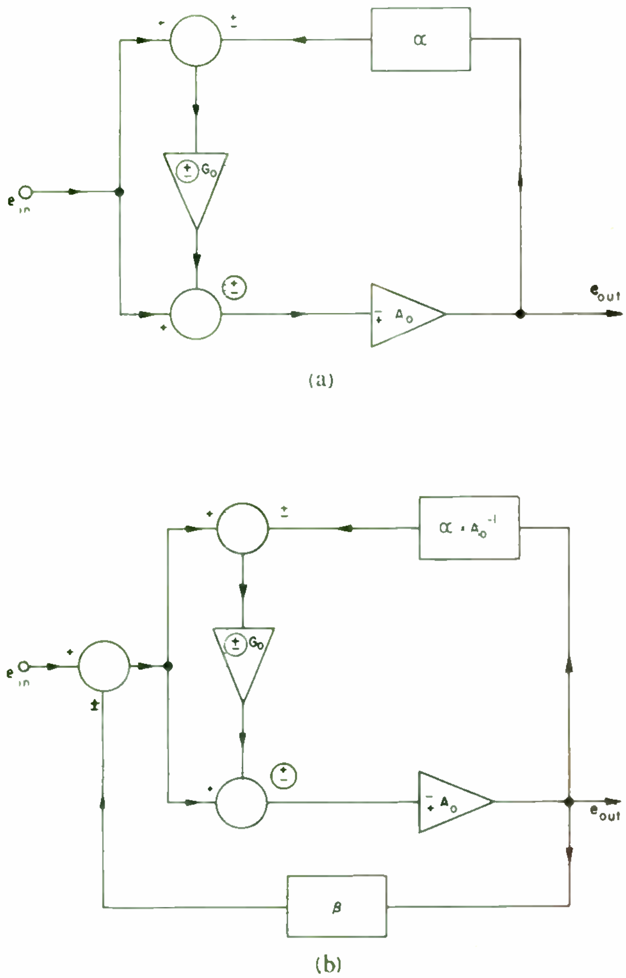
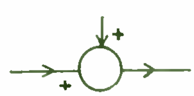


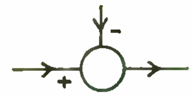

Fig. 1—(a) Block diagram showing connection of active error feedback around the amplifier of gain $A_0$; (b) block diagram showing a method of combining active error feedback and ordinary negative feedback.

Fig. 1(a) shows a block diagram of a general AEF circuit. The circuit, the gain of which is to be stabilized, the distortion and output impedance of which are to be reduced, and the frequency response of which is to be improved, has a mid-frequency numerical gain of $A_0$. We have drawn this block diagram in terms of the positive mid-frequency numerical gains $A_0$ and $G_0$ rather than the complex phasor gains $A(f)$ and $G(f)$ in order to show explicitly the possible signs which may occur in the mid-frequency region. The symbol



indicates addition and

[1] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., New York, pp. 403–404; 1943.



subtraction of the two input voltages. A variable voltage entering the junction at a plus sign goes through unchanged in sign, but a voltage entering at a minus sign has its polarity inverted. The plus-or-minus signs within circles in Fig. 1(a) go together as do those without circles, but the signs of the two sets may be specified independently.

The block diagram shows that the output voltage is multiplied by a factor $\alpha$, the result subtracted from the input voltage, and the resulting error voltage amplified by the factor $G_0$. Since only error voltage is amplified in this branch of the circuit, the amplifier of mid-band gain $G_0$ need handle only fairly small signals and need not itself be distortionless. Finally the amplified error voltage is added to the input in such a phase sense that it reduces the difference between the input and $\alpha$ times the output. It is usually most convenient to make $\alpha$ the pure numeric $A_0^{-1}$. Then the AEF tends to make the output follow the input with no gain reduction.

Analysis of the block diagram yields the following result for the over-all gain $e_{out}/e_{in}$,

$$e_{out}/e_{in} = A_0[1 + G_0]/[1 + \alpha A_0 G_0] = A_0. \qquad (1)$$

The second equation follows on taking $\alpha = A_0^{-1}$. If we continue to take $\alpha = A_0^{-1}$ and generalize (9) for complex phasor gains, we obtain

$$\frac{e_{out}}{e_{in}} = \frac{A(f)[1 + G(f)]}{1 + G(f)A(f)/A_0}. \qquad (2)$$

So long as $G(f)A(f)$ is considerably greater than $A_0$, (2) reduces closely to

$$\frac{e_{out}}{e_{in}} \cong A_0, \qquad (2')$$

the midband gain. We thus see that AEF can considerably extend the flat response of the $A$-circuit provided that the frequency response of the $G$-circuit is initially the wider and that $G_0$ is considerably larger than unity. A straightforward calculation also shows that harmonic components and the output impedance are each reduced by the factor $|1+G(f)A(f)/A_0|$, which will be considerably greater than unity over the frequency range of interest. Finally, (2′) shows that the fundamental-signal gain of the circuit is stabilized by the AEF circuit to the mid-frequency value when $\alpha = A_0^{-1}$. Thus, the circuit yields the usual advantages of negative voltage feedback without the usual decrease of gain. The additional gain required is of course supplied by the active $G$-circuit. As in any feedback circuit, it is necessary, in order to avoid regeneration, that $G(f)A(f)/A_0$ become less than unity before the phase shift of the combination reaches 180 degrees. The usual Nyquist criterion for stability is applicable here with $\beta$ given by $-G(f)/A_0$.

A combination of AEF and negative feedback can be applied to an amplifier as shown in Fig. 1(b). However, analysis of this circuit shows that the combination acts as though the extra gain of the AEF were directly in the normal negative feedback loop. Thus, although the effective negative feedback is increased, the AEF has not appreciably simplified the problem of equalizing the amplifier and feedback paths to avoid regeneration and to achieve unconditional stability. This latter statement needs qualification in one way. Since the gain $G$ is essentially outside the main amplification path, its phase and amplitude may be conveniently controlled without the necessity (which might arise with the same total gain used only with negative feedback) of having to equalize the gain $A$ and possibly thereby reduce the effective feedback at high or low frequencies. In addition, if the entire circuit is to be direct coupled, the splitting of the effective feedback into two paths in the fashion of Fig. 1(b) will usually require a smaller dc supply voltage than would be needed had all the available gain been distributed serially in the direct amplification path. When a large amount of effective negative feedback is required, its realization in a direct-coupled amplifier with reduced supply voltages may be an important economic advantage.

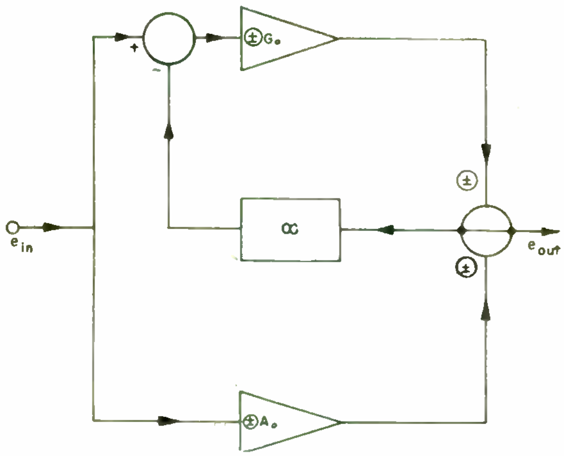



Fig. 2—Block diagram showing alternative connection of active error feedback.

The AEF circuit of Fig. 1(a) may be rearranged to inject the amplified error voltage into the output rather than the input of the amplifier of gain $A(f)$. The resulting circuit, with some of the possible signs indicated, is shown for midband gains in Fig. 2. If $\alpha$ is taken as $A_0^{-1}$ as usual, the complex gain of the circuit is found to be

$$\frac{e_{out}}{e_{in}} = A_0\left[\frac{A(f) + G(f)}{A_0 + G(f)}\right]. \qquad (3)$$

Thus, the gain will be stabilized to the value $A_0$ over a wide frequency range as long as $G(f)$ is appreciably larger in magnitude than $A(f)$. Here it is necessary for stability that $G(f)/A_0$ become less than unity before the phase shift of $G(f)$ reaches 180 degrees.

The output impedances $Z_A$ of amplifier $A$ and $Z_G$ of amplifier $G$ will be connected together across the load in the circuit of Fig. 2. The effective output impedance of the combination (the internal impedance of the composite unit) is readily found to be

$$Z_{i\,eff} = \frac{Z_A Z_G}{Z_G + Z_A[1 + G(f)/A_0]} \cong \frac{Z_G}{1 + G(f)/A_0}, \qquad (4)$$

where the second equation follows when $|G(f)|/A_0 \gg 1$ and when $Z_G$ and $Z_A$ are comparable. These conditions also lead to the gain given by (3).

When $A_0$ is large, it will usually be inconvenient to make $G_0/A_0 \gg 1$. In this case, the AEF circuit of Fig. 1(a) will be more suitable than that of Fig. 2. However, when AEF is applied around an individual stage of relatively low gain, the circuit of Fig. 2 may become preferable. This may be particularly the case when added power or current handling capacity is required, since the $A$ and $G$ amplifier outputs are effectively in parallel and thus need each supply only part of the total required output power or current. Examples are a driver which must supply appreciable undistorted current, or a power output stage. The former will be discussed in more detail in the next section.

The distinction between amplified (or active) negative feedback and AEF should be emphasized. Amplified negative feedback would be obtained if the amplifier $G$ amplified a portion $\alpha$ of the output only. It is only when the error between a portion of the output and the input is amplified that AEF is obtained. It may be noted that amplified negative feedback produces the same reduction in output impedance that AEF does, but that while AEF stabilizes but does not reduce the midband gain, amplified negative feedback reduces it by about the same factor that the output impedance is reduced. It is obvious that while the present discussion has dealt only with AEF involving the output voltage, an AEF circuit could be applied which would make the output current, rather than the output voltage, follow the input voltage (or current).

## II. The Augmented Cathode Follower

For many applications, a circuit having wide dynamic range and low output impedance is desirable. For example, the direct-coupled driver of an output tube which is to be driven into the positive-grid region must have such characteristics. The input resistance of such a tube may be as low as 100 ohms when its grid is driven far positive. Further, this resistance is a strongly nonlinear function of grid voltage. To avoid appreciable distortion, the driver of such a tube must itself have an output impedance considerably below 100 ohms and must, at the same time, be capable of supplying large positive peak grid currents.

An arbitrarily low output impedance can be obtained from an ordinary plate-loaded amplifier by applying sufficient negative voltage feedback around it. However, the load current must flow through the output plate resistor, which is often undesirable, and the change of dc voltage level between the grid and plate of the output tube may complicate the use of such a circuit in a direct-coupled amplifier. Even if the driver tube itself is a cathode follower whose output impedance is reduced by ordinary inverse feedback around previous amplifier stages, these stages will be in the direct amplification path, again complicating its use in a direct-coupled circuit. In the present section, we show how these difficulties may be avoided by applying AEF to a cathode-follower driver. The resulting direct-coupled circuit has both very low output impedance and no appreciable change in dc voltage level between input and output.

Fig. 3 indicates one way of adapting the AEF circuit of Fig. 2 to a cathode follower. We shall call the resulting circuit a parallel augmented cathode-follower driver (PACFD). The type of AEF shown in Fig. 2 is particularly applicable to the cathode follower because the latter's gain is near unity and thus the external gain $G$ need only be greater than unity to be effective in reducing output impedance and distortion. Further, $\alpha$ can be conveniently taken equal to unity.

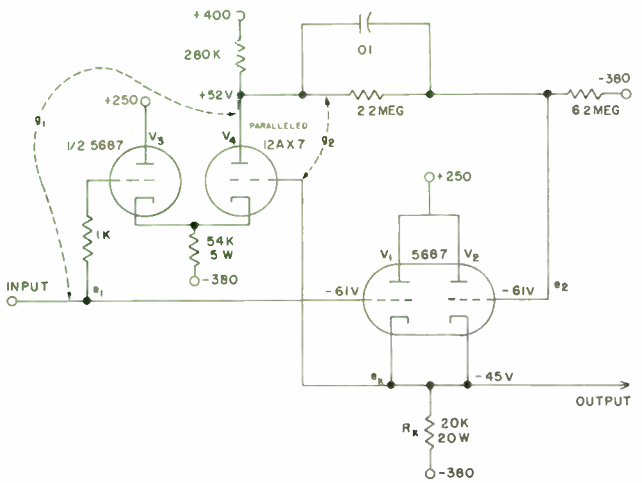



Fig. 3—The parallel augmented cathode-follower driver.

As shown in Fig. 3, the difference between the input $e_1$ and the output $e_k$ of tube $V_1$ is amplified by the differential amplifier[2] consisting of $V_3$ and $V_4$, then applied to the grid of the parallel cathode follower $V_2$ to reduce the error between $e_1$ and $e_k$. In this direct-coupled circuit, it is desirable that $V_3$ be of the same tube type as $V_1$ and $V_2$, in order that operating biases be correct. In an ac coupled version of the circuit both $V_3$ and $V_4$ could be, for example, the halves of a single 12AX7. It is worth mentioning that a cathode follower can be augmented in another way by using the tube half $V_2$ as a cathode follower in series with $V_1$ so that the cathode of $V_2$ is connected to the plate of $V_1$. Then the grid of $V_2$ could be direct-coupled to the plate of $V_4$ without the voltage divider necessary in Fig. 3. We shall designate such a unit a series augmented cathode-follower driver (SACFD). The SACFD is superior to an ordinary cathode-follower driver (CFD) but inferior to a PACFD, as we shall see below. In addition, its dynamic range is limited, for a given supply voltage value, by the necessary voltage division across $V_1$ and $V_2$ in series, which does not occur with the PACFD.

A straightforward analysis of the midband equivalent circuits of the SACFD and PACFD yield the following results for their gains and internal impedances:

$$G_S = [\mu(1+g_1)+\mu^2]/[\mu g_2+(1+\mu)^2+(\mu+2)r_p/R_k], \tag{5}$$

$$r_{iS} = r_p/[\mu g_2/(2+\mu)+(1+\mu)^2/(2+\mu)+r_p/R_k], \tag{6}$$

$$G_P = \mu(1+g_1)/[\mu g_2+2(1+\mu)+r_p/R_k], \tag{7}$$

$$r_{iP} = r_p/[\mu g_2+2(1+\mu)+r_p/R_k]. \tag{8}$$

$G_S$ and $r_{iS}$ refer to the SACFD, $G_P$ and $r_{iP}$ to the PACFD. In the above equations, the arithmetical gains $g_1$ and $g_2$ of the differential amplifier are those indicated on Fig. 1; they are slightly unequal, with $g_2$ the larger. Note that the algebraic gain corresponding to $g_2$ is negative.[2] It is also assumed that the tube halves $V_1$ and $V_2$ have the same characteristics. For most purposes, we shall ignore the small difference between $g_1$ and $g_2$ and designate them both by $g$. The above equations show that if $\mu g$ is sufficiently large and $r_p/R_k$ small, both $G_S$ and $G_P$ will approach unity closely. Further $r_{iS}$ will approach $r_p/(\mu+g)$ and $r_{iP}$ will be approximately $r_p/\mu g$. Note that were amplified negative feedback used in the PACFD (e.g., by grounding the grid of $V_3$ for input signals) instead of AEF, $g_1$ would then be zero, and $G_P$ would be reduced to about $g_2^{-1}$ while $r_{ip}$ would remain unchanged.

For comparison with the above results, the equations pertaining to an ordinary cathode follower are

$$G = \mu/[1 + \mu + r_p/R_k], \tag{9}$$

$$r_i = r_p/[1 + \mu + r_p/R_k]. \tag{10}$$

When $r_p/R_k$ is small and $\mu$ appreciably larger than unity, we see from these results that to good approximation the output impedance of the SACFD is reduced over that of an ordinary cathode follower of the same characteristics as $V_1$ by the factor $(\mu+g)/\mu$ and that of the PACFD is reduced by the factor $g$. The principal reason for the difference is that the error voltage at the plate of $V_4$ is degenerated in the SACFD by a factor of about $\mu$ when applied to the plate of $V_1$ and so is less effective in reducing the output error than is that of the PACFD. Such degeneration is instrumental in reducing the dynamic range of the SACFD even further. Since the PACFD makes superior use of the same tubes required in the SACFD, we shall concentrate on the former in the rest of this work.

[2] G. E. Valley, Jr. and H. Wallman, "Vacuum Tube Amplifiers," Rad. Lab. Ser., McGraw-Hill Book Co., Inc., New York, vol. 18, pp. 442–443; 1948.

It may be noted that the double cathode follower[3] achieves, with two tubes in series, about the same small-signal gain and output impedance as the PACFD. The top input tube is plate loaded and its cathode connected to the plate of the bottom tube. The bottom tube is itself driven from the plate of the top tube. Neither the SACFD nor the double cathode follower are comparable to the PACFD as drivers, however. In the SACFD, the driving current must pass through both the upper series tube and the lower cathode-follower tube. In the double cathode follower, it must pass through both the load resistor $R_L$, which should be appreciably greater than $r_p$, and through the upper tube. In the PACFD, the driving current is supplied by both the cathode-followers $V_1$ and $V_2$ of Fig. 3, essentially in parallel. The dynamic range and current handling capacity of the PACFD are thus much superior to those of the other two circuits.

### III. Comparison of Theory and Experiment

The circuit of Fig. 3 was constructed with the parameter values and tubes shown. It was found that its no-load gain was 0.986. Next, the output voltage was measured as a function of total load resistance $R_L$ (the parallel combination of $R_k$ and any added load) for a fixed input voltage. The measurements were carried out at $10^4$ cps using a 30 $\mu f$ oil capacitor in series with a variable load resistance; only at the lowest load resistances was the capacitative reactance of importance.

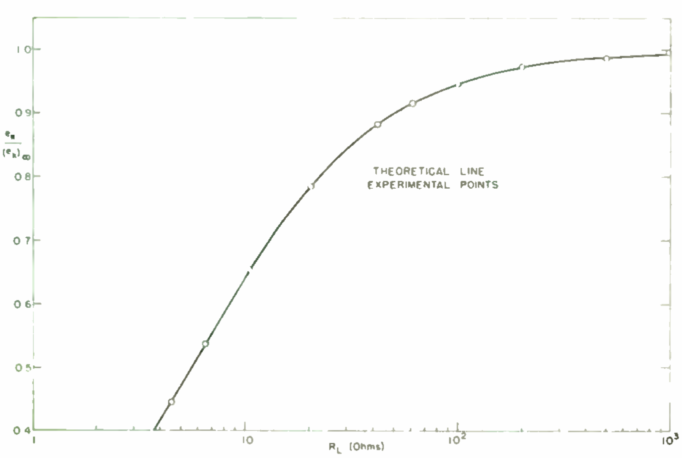



Fig. 4—Dependence on load resistance of the normalized output voltage of the PACFD.

Fig. 4 shows the load dependence of the output voltage $e_k$ normalized with respect to that without load $(e_k)_\infty$. The theoretical line of this figure was calculated using (7) with $R_k$ replaced by $R_L$. The values $g=70$, $\mu=16$ and $r_p=6.45$ kilohms were employed; these values are in reasonable agreement with published curves. Fig. 4 shows that these values are indeed a good choice, and that theory and experiment are in agreement. In addition, the internal impedance, defined as the added load necessary to make $e_k/(e_k)_\infty=0.5$, is shown to be 5.6 ohms. For comparison, the internal impedances of the SACFD and CFD using the same tubes were found to be of the order of 70 and 370 ohms, respectively. The above definition of $r_{iP}$ leads to the same result for this quantity as that given in (8), which was calculated on the basis of a grounded input and a measuring signal applied to the output. Alternatively, if $r_{iP}$ is again determined by loading the output but defined as the added load required to make $e_k/e_1=0.5$, the expression for $r_{iP}$ becomes

$$r_{iP} = r_P/[\mu(2g_1 - g_2) - 2 - r_p/R_k]. \tag{11}$$

For large $\mu g$, it does not differ appreciably from (8).

Next, the amplified error voltage $e_2$ (see Fig. 3) was measured under the same conditions as above for a fixed input voltage $e_1$. The normalized quantity $e_2/e_1$ is plotted in Fig. 5 vs $R_L$. The small-signal equivalent circuit yields a value for this ratio of

$$e_2/e_1 = [\mu(2g_1 - g_2) + g_1(2 + r_p/R_k)]/[\mu g_2 + 2(1+\mu) + r_p/R_k]. \tag{12}$$

This quantity is slightly greater than unity even for $R_k$ infinite. The solid line of Fig. 5 was calculated from (12), replacing $R_k$ by $R_L$ and using the same values for the tube parameters as those used for Fig. 4. Again, agreement between theory and experiment is exceptionally good. It is of interest to note that at very large loads $e_2/e_1$ may be much greater than unity; its maximum value will be approximately $g$ if this value can be achieved without overdriving the tube $V_4$.

Finally, it should be pointed out that the data of Figs. 4 and 5 were measured with values of $e_1$ of the order of 0.1 volt or less. The equivalent circuit and the resulting formulas only hold as long as operation is in a linear region. When negative peaks are to be produced across a load sufficiently large that the peak current required exceeds the quiescent current in $R_k$, the tubes $V_1$ and $V_2$ will be cut off and negative peak limiting will occur. Only by employing voltages sufficiently small that such limiting did not occur could an accurate undistorted value of $e_k$ be obtained when very low load resistances were used. This negative peak limiting is the reason why a single PACFD or a pair in push-pull cannot be conveniently used to drive a load like a loudspeaker directly even though the small-signal impedances may be matched.

### IV. Comparison of Grid-Driver Circuits

The PACFD is ideally suited for a grid driver. Because it uses two cathode-followers essentially in parallel ($V_1$ and $V_2$), it can supply twice the peak positive grid current of a single unit. In addition, as the current increases, the $g_m$ and $\mu$ of both tubes increase and the $r_p$'s fall. For example, at 50 ma per tube-half, the $\mu$ and $g_m$ of a 5687 are approximately 19 and 12,000 $\mu$mhos, re-

[3] S. Seely, "Electron-Tube Circuits," McGraw-Hill Book Co., Inc., New York, pp. 120–121; 1950.

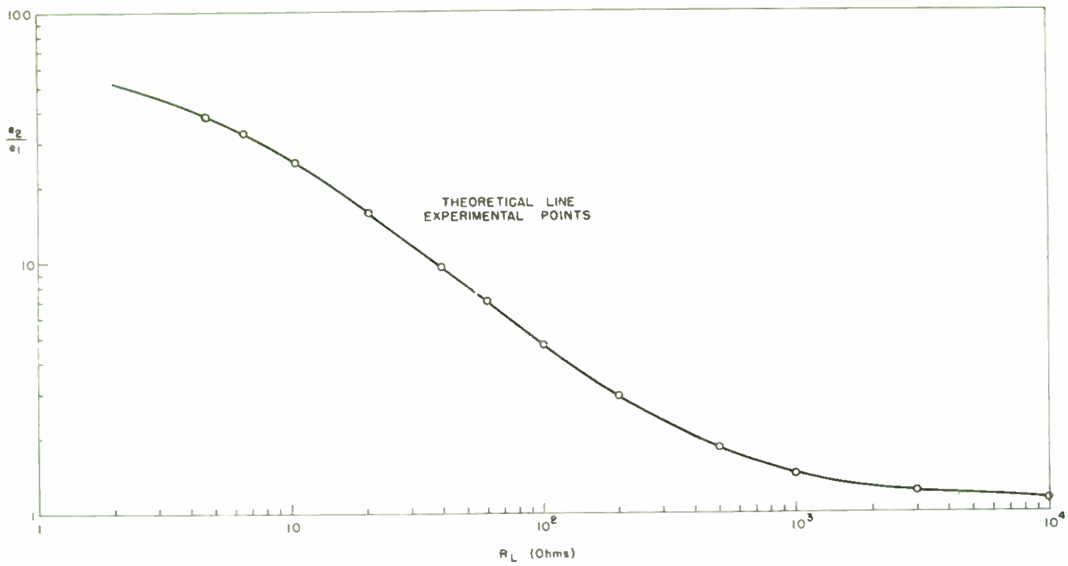



Fig. 5—Dependence on load resistance of the normalized error voltage of the PACFD.

spectively. Using $g=70$, (8) or (11) predict an internal impedance of the PACFD of only about 1.15 ohms instead of the value of 5.6 ohms found for small signals with the circuit of Fig. 3.

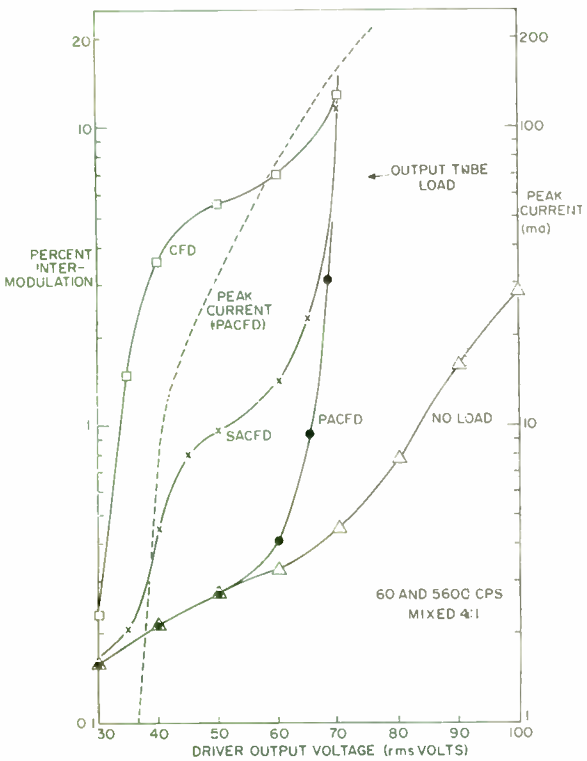



Fig. 6—Comparison between the intermodulation distortion of three drivers when direct-coupled to an output tube grid.

In Fig. 6, we give a comparison between the distortion generated by a CFD, a SACFD, and a PACFD when direct coupled to a power tube grid load. The lowest line, marked "no load," shows the distortion in the unloaded PACFD output. This distortion arises almost entirely from the preceding amplifier stage. The output tube was an 807, triode connected, with 400 volts on the plate. It had an unbypassed 25-ohm cathode resistor and formed half of a push-pull output circuit with output transformer and resistive load. The other half of the push-pull output circuit was, in each case, driven by a driver identical to that measured.[4] The intermodulation distortion was measured at the output of the driver and employed 60 and 5,600 cps signals, mixed 4:1. The dc bias of the output tube was adjusted to $-42.5$ volts so that the grid was driven positive when the rms driver voltage exceeded 30 volts. It is this positive grid region which is presented in Fig. 6.

The dotted line is the approximate peak grid current supplied by the driver. When the rms driver voltage is 70 volts, the grid is driven positive by 56.5 volts peak, and we see that it draws a peak current of about 200 ma. It is obvious from Fig. 6 that the SACFD is a considerable improvement on the CFD, and the PACFD an improvement on the SACFD over most of the range considered. For applied voltages greater than 60 to 65 volts rms, the grid of the output tube loses control of the output current on positive peaks; the point at which control is lost defines the diode line of the output tube. It is seen from the figure that the distortion of all the drivers increases rapidly for larger voltages. Oscillographic observations showed, however, that the PACFD was capable of driving the grid of the output tube considerably beyond the point where the output voltage of the output tube began to show peak clipping arising from diode-line limiting. Even in this region, however, appreciable distortion of the grid signal could not be observed on the CRO.

[4] The push-pull driver circuit used in these measurements incorporated a special feedback loop which reduced even-order harmonic distortion greatly at the driver outputs. Therefore, the intermodulation distortion results obtained at one of the push-pull driver outputs may be appreciably smaller, especially for the case of the CFD, than would be attained in practice without such a feedback loop. Nevertheless, the distortion curves still afford a valid comparison between the relative distortion of the three types of drivers.

# A Semiconductor Diode Multivibrator*

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*Summary*—**The operation and design of a novel semiconductor multivibrator circuit is described. Consisting of one double-base diode, one diode, three resistors, and one capacitor, the diode multivibrator affords a circuit economy of two-to-one over the more conventional Eccles-Jordan configuration. The active element is the double-base diode, which is a three-terminal single-junction negative-resistance semiconductor device. Applications of the diode multivibrator to square-wave oscillators and delayed-pulse generators are illustrated. Control of the period of oscillation or variation of the pulse-delay time is accomplished by the variation of a single capacitor. Hence, this new circuit simplifies the design of astable, monostable or bistable multivibrators.**

## Introduction

THE INCREASING use of quantized signals as "information carriers" in modern electronic systems has caused a growing concern over the power requirements, complexity, and expense of pulse-type circuits. Thus the advent of the transistor as a low-power, long-lifetime, and sub-miniature active element was welcomed most enthusiastically by pulse-circuit engineers as a possible solution to the dilemma of increasing system complexity and size.

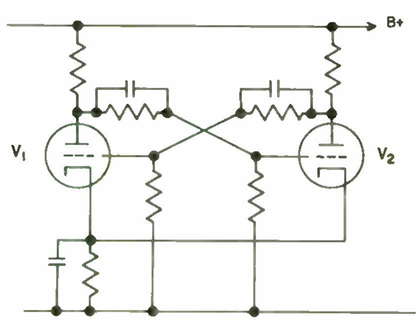


Fig. 1—Vacuum-tube multivibrator.

One of the oldest, and perhaps most fundamental, of the digital-type circuits is the multivibrator.[1] As a two-state circuit, analogous to the mechanical relay, the multivibrator extended the use of relaying to high-speed operations. And since the "on-off" device is virtually the "nerve-cell" of all counting and logic circuits, the multivibrator ultimately became a fundamental building block of modern electronic computers.

The semiconductor multivibrator to be described in this paper may be considered as a diode flip-flop. It has been made possible by the invention of a new semiconductor device—the double-base diode.[2–4] In the diode multivibrator one diode "drives" another in a reciprocal "on-off" relationship. A significant departure from the well-known Eccles-Jordan circuit, which is illustrated in Fig. 1, using vacuum-tubes, and in Fig. 2, using transistors, is that only one active circuit element is required for the regenerative action.

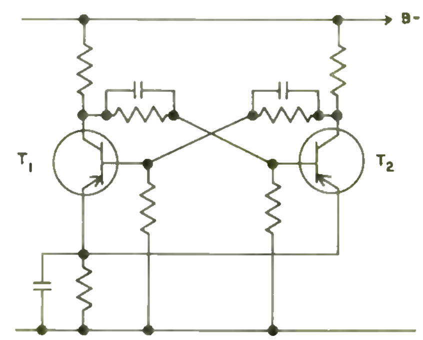


Fig. 2—Junction-transistor multivibrator.

The astable and monostable diode multivibrators will be treated here. The use of the former as a square-wave oscillator and the latter as a delayed-pulse generator will be described. Because of the almost two-to-one reduction in circuit components required by the diode multivibrator, when compared to the Eccles-Jordan transistor configuration, a higher degree of circuit simplicity, miniaturization, and economy may be achieved.

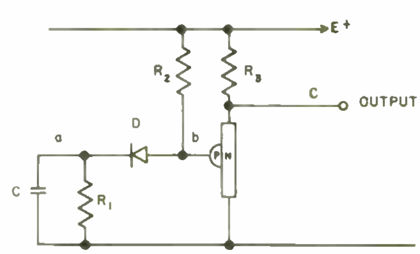


Fig. 3—Diode multivibrator.

## Principles of Operation

The basic circuit configuration of the diode multivibrator is illustrated in Fig. 3. Its relative simplicity is immediately apparent when compared to the transistor

* Original manuscript received by the IRE, January 5, 1955; revised manuscript received, March 3, 1955.

† Electronics Lab., Gen. Elec. Co., Syracuse, N. Y.

[1] H. Abraham and E. Block, Ministere de la guerre Pub. 27; April, 1918.

[2] J. A. Lesk and V. P. Mathis, "The double-base-diode—a new semiconductor device," 1953 *IRE Convention Record*, pp. 2–8.

[3] R. F. Shea, *et al.*, "Principles of Transistor Circuits," John Wiley and Sons, Inc., New York, N. Y., pp. 466–471; 1953.

[4] J. J. Suran, "The double-base diode—a semiconductor thyratron," *Electronics*, pp. 198–202; March, 1955.

multivibrator shown in Fig. 2. Operation of the astable diode multivibrator may be described as follows. Capacitor $C$ is charged from the battery supply through the resistance $R_2$ and the diode, $D$. During the charging cycle of the capacitor, $D$ is conducting but the double-base diode is in the cut-off state. When the potential across the capacitor becomes equal to or greater than the peak-point potential of the double-base diode, the latter becomes unstable and switches into the conducting state. The junction potential ($b$ in the circuit of Fig. 3) is then clamped to a low value—almost to the ground potential, thus causing the diode $D$ to become cut off. When the diode is in its nonconducting state, point $a$ is virtually isolated from point $b$. The capacitor will then discharge through the resistance $R_1$ until the potential at $a$ is approximately equal to the junction potential of the double-base diode. At this instant, the diode becomes conducting again. When the diode reverts to its conduction state, the current through the junction of the double-base diode decreases and the latter is driven into its cut-off state. Capacitor $C$ will then recharge and the cycle will be repetitive.

The waveforms generated by the astable circuit of Fig. 3 are illustrated in Fig. 4. Since capacitor $C$ alternately charges and discharges through a fixed resistance ($R_2$ and $R_1$, respectively), the waveform at $a$ consists of a periodic exponential rise and decay. During the time

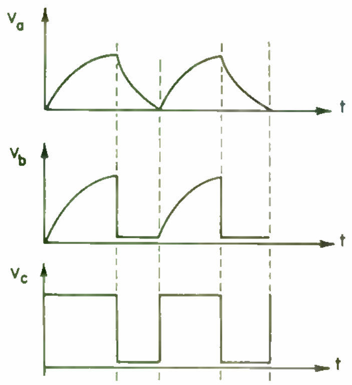


Fig. 4—Diode multivibrator waveforms.

$D$ is conducting, the waveform at $b$ will be almost identical to that at $a$. However, when the double-base diode becomes conducting, the potential at $b$ is clamped to a near-ground value until the capacitor has completed its discharge cycle. When the double-base diode is in its cut-off state, the current through $R_3$ will be comparatively low. However, when the double-base diode switches "on," its bar resistance drops by an order of magnitude and the current through $R_3$ increases. Thus, the current through $R_3$ will be either high or low, depending upon the operating state of the double-base diode. Consequently, the waveform across $R_3$ (at point $C$) will be a square-wave as illustrated in Fig. 4. It is apparent that the frequency and symmetry of this square wave is dependent upon the time constants associated with $R_1$, $R_2$ and $C$.

## Graphical Analysis

A better understanding of the operation of the diode multivibrator may be obtained by considering the equivalent circuits illustrated in Fig. 5. Figure 5(a) illustrates the dc equivalent circuit of the diode multivibrator when $D$ is conducting. The capacitor is omitted from the dc circuit and the diode is assumed to have negligible forward resistance. Writing the loop equations for the circuit of Fig. 5(a), the following set is obtained:

$$E = (R_1 + R_2)I_1 - R_1 I_d \tag{1a}$$

$$O = - R_1 I_1 + R_1 I_d + V_d. \tag{1b}$$

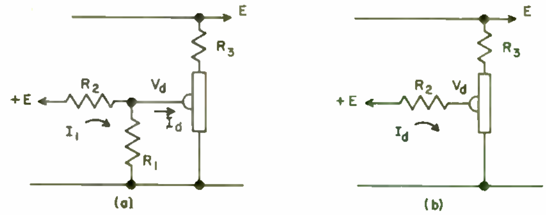


Fig. 5—Dc equivalent circuits; (a) diode conducting, (b) diode nonconducting.

In (1b), $V_d = f(I_d;\ E,\ R_3)$, which represents the input characteristics of the double-base diode for a given battery supply, $E$, and load resistance, $R_3$. Eqs. (1a) and (1b) are easily solved for $V_d$ as a function of $I_d$. Thus

$$V_d = \frac{R_1}{R_1 + R_2} E - \frac{R_1 R_2}{R_1 + R_2} I_d. \tag{2}$$

When the diode $D$ in the circuit of Fig. 3 is nonconducting, the steady-state equivalent circuit of Fig. (5b) may be obtained. It is assumed that the resistance $R_1$ is effectively isolated from the double-base diode by the very high-back resistance of the diode. For the circuit of Fig. (5b),

$$V_d = E - R_2 I_d. \tag{3}$$

The application of (2) and (3) to the operating characteristics of the double-base diode permits a graphical load-line analysis similar to that used in vacuum-tube and transistor circuits design. This is illustrated in Fig. 6 (next page). For the condition that the diode conducts, the steady-state input load line is determined by (2) and is represented by the dashed line. The intersection of load line with ordinate axis is at a point

$$V_d = E \frac{R_1}{R_1 + R_2},$$

and the slope of the load line is the parallel combination of $R_1$ and $R_2$. For the condition that the diode is nonconducting, the load-line characteristic is determined by (3) and is represented by the solid line (slope $= R_2$) in Fig. 6.

In order for the diode multivibrator to be astable, or free running, the following conditions must be satisfied.

1. When the diode is conducting, the input load line cannot intersect the double-base diode characteristic in the cut-off region.

2. When the diode is nonconducting, the input load line must intersect the double-base diode operating characteristic in the transition, or negative-resistance, region. Consequently, the circuit conditions for the free-running multivibrator are as follows:

$$\frac{R_1E}{R_1+R_2} > V_p \tag{4a}$$

$$\frac{E}{R_2} \leq I_v. \tag{4b}$$

In (4a) and 4(b), $V_p$ is the peak-point potential of the double-base diode and $I_v$ is the input current corresponding to its valley point. The operating path of the multivibrator, in relation to the input characteristics of the double-base diode, may be approximately determined from the graphical analysis, as indicated in Fig. 6.

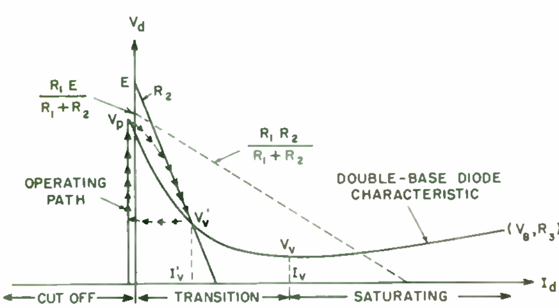


Fig. 6—Operating characteristic of the diode multivibrator.

## Design Analysis

From the steady-state circuit analysis it has been possible to determine the conditions required for sustained oscillation of the multivibrator. The frequency and symmetry of the generated waveforms may be determined from a consideration of the charging and discharging cycles of the capacitor, $C$, in Fig. 3. It may be noted that when $D$ is conducting, the capacitor is charging. This also corresponds to the condition that the double-base diode is in its cut-off state and that the potential at $c$ is comparatively high. Hence, the diode conducts during the positive cycle of the square-wave and is nonconducting during the negative cycle of the square wave. The charging and discharging equivalent circuits, corresponding to the positive and negative portions of the output waveform, are illustrated in Fig. 7.

In terms of the Laplace operator, $s$, the voltage transform for the circuit of Fig. 7(a) is

$$\frac{V_a}{E}(s) = \left[\frac{R_1}{R_1+R_2}\right]\left[\frac{1}{1+\dfrac{R_1R_2Cs}{R_1+R_2}}\right]. \tag{5}$$

Closing the switch, $S$, is assumed to be equivalent to $D$ suddenly becoming conducting. Eq. (5) is solved in the time domain for the following initial conditions:

$$\text{at } t = 0: \qquad V_i = \frac{E}{s}; \qquad V_a = V_v'. \tag{6}$$

Eq. (6) indicates that a step voltage of magnitude $E$ is applied to the circuit of Fig. 7(a) at a time when $C$ is charged to a potential of $V_v'$. Using the standard Laplace transformation techniques (5) and (6) may readily be solved to give the following result in the time domain:

$$V_a(t) = \frac{R_1E}{R_1+R_2}\left[1 - \exp\left(-\frac{(R_1+R_2)t}{R_1R_2C}\right)\right] + V_v' \exp\left(-\frac{(R_1+R_2)t}{R_1R_2C}\right). \tag{7}$$

The time that it takes the capacitor to charge to the peak-point voltage, $V_p$, may be calculated from (7) by substituting $V_a(t) = V_p$ on the left-hand side of the equation. Thus

$$t_p = -\frac{R_1R_2C}{R_1+R_2} \ln\left[\frac{\dfrac{R_1E}{R_1+R_2} - V_p}{\dfrac{R_1E}{R_1+R_2} - V_v'}\right]. \tag{8}$$

where $t_p$ denotes the duration of the positive portion of the square wave.

Similarly, the duration of the negative portion of the square wave can be obtained from the equivalent circuit of Fig. 7(b). For the initial condition that $V_a(0) = V_p$ and calculating the time it takes the capacitor to discharge to a value $V_v'$, it is found that

$$t_n = -R_1C \ln \frac{V_v'}{V_p} \cdot \tag{9}$$

In (9), $t_n$ is the time duration of the negative portion of the output wave.

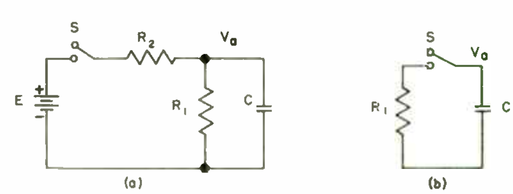


Fig. 7—Low-frequency ac equivalent circuits; (a) positive cycle, (b) negative cycle.

The total period, $t_T$, of the square wave is the sum, $t_p+t_n$. Hence

$$t_T = -R_1C\left\{\frac{R_2}{R_1+R_2}\ln\left[\frac{1-\left(\dfrac{V_p}{E}\right)\left(\dfrac{R_1+R_2}{R_1}\right)}{1-\left(\dfrac{V_v'}{E}\right)\left(\dfrac{R_1+R_2}{R_1}\right)}\right] + \ln\frac{V_v'}{V_p}\right\}. \tag{10}$$

For convenience, the design parameters are defined:

$$P = \frac{V_p}{E}, \qquad U = \frac{V_v'}{E}, \qquad a = \frac{R_1}{R_2}, \qquad X = \frac{t_p}{t_n}. \tag{11}$$

If a symmetrical square wave is desired, $t_p$ must equal $t_n$. From (10) therefore, using constants defined by (11). the following condition for symmetry is established:

$$\frac{1 - P\left(1 + \dfrac{1}{a}\right)}{1 - U\left(1 + \dfrac{1}{a}\right)} = \left[\frac{U}{P}\right]^{1+a}. \tag{12}$$

If the ratio of the positive portion of the square wave to total period is defined as figure of symmetry, $S$, then

$$S = \frac{X}{1 + X} \quad \text{where} \quad X = \frac{t_p}{t_n}. \tag{13}$$

It can then be shown that, for any desired figure of symmetry, (10) must satisfy the condition

$$\frac{1 - P\left(1 + \dfrac{1}{a}\right)}{1 - U\left(1 + \dfrac{1}{a}\right)} = \left[\frac{U}{P}\right]^{(S/1-S)(1+a)}. \tag{14}$$

For a symmetrical waveform, $S=0.5$. Eq. (12), therefore, is a special case of (14).

The output waveform across $R_3$ depends upon the change in the bar resistance of the double-base diode as the latter oscillates between the cut-off and conducting states. Denoting the equivalent bar resistances by $R_{BC}$ and $R_{BS}$, where the subscripts $C$ and $S$ refer to the cut-off and conducting states of the double-base diode, respectively, the equivalent circuit of Fig. 8 may be used

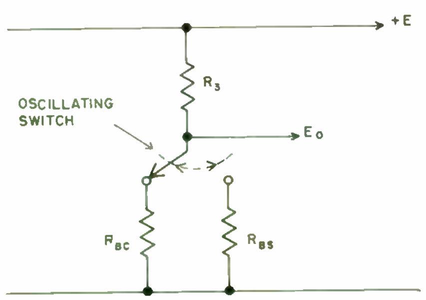


Fig. 8—Low-frequency equivalent-output circuit.

to represent the low-frequency output circuit of the multivibrator. The peak-to-peak amplitude of oscillation across $R_3$ is then given by

$$|E_0| = \left[\frac{1}{1 + \dfrac{R_3}{R_{BC}}} - \frac{1}{1 + \dfrac{R_3}{R_{BS}}}\right][E]. \tag{15}$$

If it is desired to find the value of $R_3$ which maximizes the output voltage amplitude, (15) may be differentiated and solved for a maximum in the usual manner. It is found that the value of $R_3$ required to obtain a maximum $E_0$ is equal to the geometric mean of the two bar resistances, $R_{BC}$ and $R_{BS}$. Thus

$$R_{3m} = \sqrt{R_{BC}R_{BS}}. \tag{16}$$

The maximum output voltage, found by substituting (16) into (15), is

$$|E_0|_{\max} = \left(\frac{1 - \sqrt{R_{BS}/R_{BC}}}{1 + \sqrt{R_{BS}/R_{BC}}}\right)(E). \tag{17}$$

The peak-point voltage of the double-base diode is linearly related to the interbase potential.[4,5] Denoting the constant of proportionality by $H_{12}$, one can write

$$V_p = H_{12}V_B. \tag{18}$$

In (18), the factor $H_{12}$ is always less than unity and is a constant of the device. When the double-base diode is in its cut-off state, the interbase potential is

$$V_B = \frac{R_{BC}}{R_{BC} + R_3}E. \tag{19}$$

Substituting (18) into (19) gives

$$P = \frac{V_P}{E} = \frac{H_{12}}{1 + R_3/R_{BC}}. \tag{20}$$

For maximum output voltage, given by (16), (20) is

$$P_m = \frac{V_P}{E} = \frac{H_{12}}{1 + \sqrt{R_{BS}/R_{BC}}}. \tag{21}$$

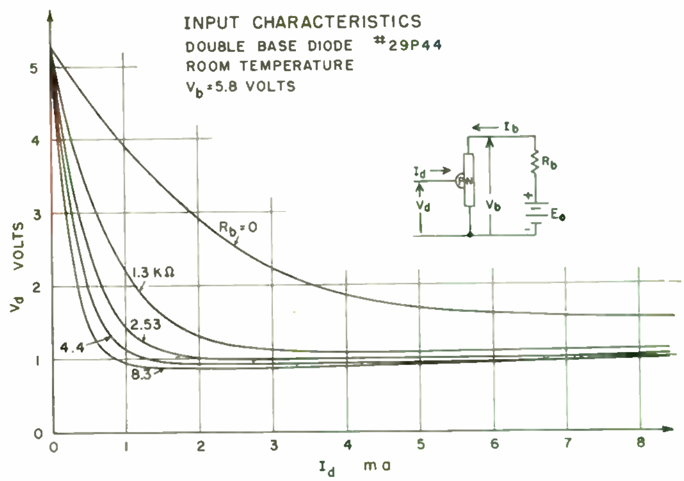


Fig. 9.

Fig. 9 illustrates a typical set of input characteristics for a double-base diode with a load resistance in the base-two lead. The interbase voltage, $V_b$, is constant (5.8 volts) only during the cut-off state. It is important to note that the valley point of the curves in Fig. 9 moves to the left as the load resistance is increased. However, the valley point approaches a limiting value as the load resistance becomes very large.[5] Consequently, $V_v'$ in (6) to (11) depends upon both the input and output load resistances and therefore $U$ in (11) may be treated as an arbitrary circuit design parameter.

[5] J. J. Suran, "Low-frequency circuit theory of the double-base diode," *Trans. IRE*, vol. ED 2; April, 1955.

## Design Procedure

Based upon the foregoing analysis, one can outline an *approximate* design procedure for the free-running diode multivibrator. Procedure given below is based on the maximum voltage criterion as defined by (16) and (17).

Eqs. (4a) and (4b) specify the two necessary conditions required for oscillation. From (4b),

(a) $$\frac{E - V_v}{R_2} < I_v,$$

where $V_v$ and $I_v$ are the voltage and current, respectively, which correspond to the valley point of the double-base diode input characteristic. Since $V_v$ and $I_v$ approach constant limits as $R_3$ is increased, the limiting values for the valley point may be used in condition 1. Thus, $R_2$ can be selected. Using this value of $R_2$, $I_v'$ may be estimated (see Fig. 6). From bar-characteristic curves of the double-base diode, such as illustrated in Fig. 10, $R_{BS}$ can be approximated. $R_{BC}$ may be determined fairly accurately from the characteristic curves of Fig. 10. Thus, using (16),

(b) $$R_3 = \sqrt{R_{BS}R_{BC}}.$$

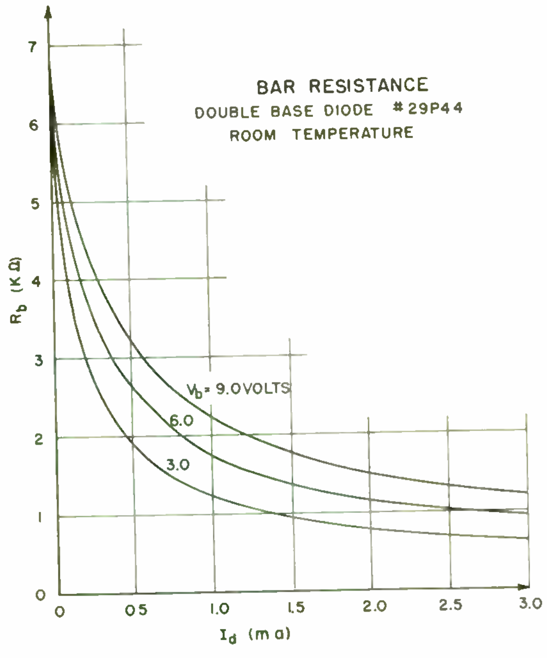


Fig. 10.

The design parameter $P$ can now be calculated from (20) or (21).

(c) $$P = \frac{H_{12}}{1 + R_3/R_{BC}}.$$

In (c), $H_{12}$ is the voltage ratio defined by (18). Knowing $I_v'$, $V_v'$ may be estimated and

(d) $$U = \frac{V_v'}{E}.$$

With $P$ and $U$ determined, the ratio $a = R_1/R_2$ may be found from (14) for any desired figure of symmetry. A family of curves for $S = 0.5$, based upon (12), is illustrated in Fig. 11. The k-values in Fig. 11 correspond to both $U$ and $P$. Hence, the intersections of these curves determine the $a$-values for respective $U$ and $P$ parameters when a symmetrical waveform is desired. $R_1$ may be determined from the design parameter $a$. Thus

(e) $$R_1 = aR_2.$$

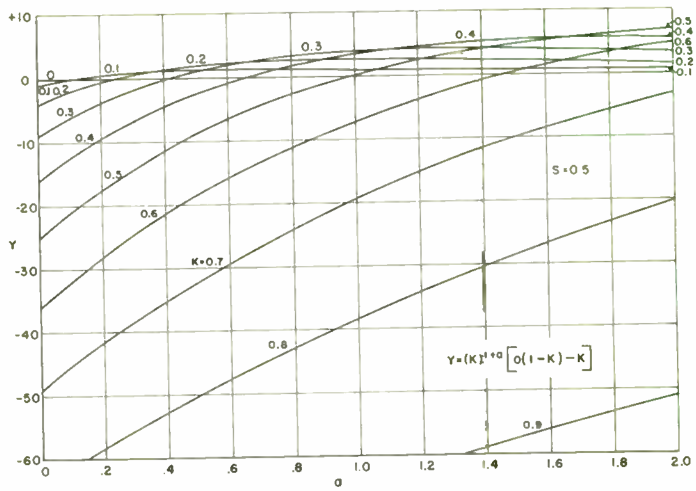


Fig. 11—Symmetry curves for $S = 0.5$.

At this point, $a$ must satisfy the condition [from (4a)]

(f) $$\frac{a}{1 + a} > P.$$

If (f) is not satisfied, (a) must be re-chosen and the design procedure repeated. However, if (f) is satisfied, the design of the multivibrator circuit is complete except for the specification of capacitor $C$. $C$ is calculated on the desired frequency, $f$, from (10). Clearly, for $S = 0.5$,

(g) $$C = \frac{1}{-2R_1 f \ln U}.$$

## Example

Suppose that it is desired to design a 10-kc diode multivibrator, having a symmetrical waveform ($S = 0.5$), and operating from a voltage source of 12 volts. Assume that the double-base diode characteristics are given by Figs. 9 and 10. The first step in the design procedure is to select a value of $R_2$ which will satisfy (4b). From Fig. 9, it is apparent that if the input load line intersects the abscissa axis at $I_{d0} = 1.0$ ma, the astable requirement will almost certainly be satisfied. Hence, from Fig. 6,

$$R_2 = \frac{E}{I_{e0}} = \frac{12}{0.001} = 12,000 \text{ ohms}.$$

The current flowing into the junction of the double-base diode, when the latter is in its conducting state, will be approximately 1 ma. This is readily established by superimposing value of $R_2$ for 12-volt source, on input characteristics of Fig. 9. From Fig. 10, for $I_d = 1$ma, $R_{BS}$

may be estimated. Thus, $R_{BS} \approx 1.8\text{k}$, and $R_{BC} \approx 6.5\text{k}$. Using maximum voltage criterion defined by (16),

$$R_3 = \sqrt{(6.5\text{k})(1.8\text{k})} = 3.5\text{k ohms}.$$

Having determined $R_3$ and $R_{BC}$, interbase potential of double-base diode, during cut-off state, may be calculated.

$$V_{BC} = \left(\frac{6.5\text{k}}{6.5\text{k} + 3.5\text{k}}\right)(12) = 7.8 \text{ volts}.$$

For $V_{BC} = 7.8$ volts, $V_v'$ (see Fig. 6) may be estimated from the input characteristics. Hence $V_v' \approx 2.0$ volts. From Fig. 9,

$$H_{12} = \frac{5.2 \text{ volts}}{5.8 \text{ volts}} = 0.9.$$

The design parameters, $U$ and $P$, defined by (11) and (20), may now be calculated.

$$P = \frac{H_{12}}{1 + R_3/R_{BC}} = \frac{0.9}{1 + \dfrac{3.5\text{k}}{6.5\text{k}}} = 0.58$$

$$U = \frac{V_v'}{E} = \frac{2 \text{ volts}}{12 \text{ volts}} = 0.167.$$

For a symmetry condition of $S = 0.5$, the ratio $a = R_1/R_2$ may be determined from the design curves of Fig. 11. The intersection of the two curves corresponding to $k = 0.58$ and $0.167$, respectively, is the desired value of $a$. from the input characteristics. Hence $N_v' \approx 2.0$ volts. from Fig. 11, $a = 1.55$, and

$$R_1 = aR_2 = (1.55)(12\text{k}) = 18.6\text{k ohms}.$$

$R_1$, $R_2$ and $R_3$ have now been determined to a first approximation. A necessary requirement for oscillation, however, as derived from (4a), is that

$$\frac{a}{1 + a} > P, \quad \text{or} \quad \frac{1.55}{2.55} > 0.58.$$

Since the latter inequality is satisfied, the three resistance values will be adequate to sustain oscillations. Capacitor $C$ may now be determined from the frequency requirement. For the symmetrical condition ($S = 0.5$),

$$C = \frac{1}{-2R_1 f \ln U} = \frac{-10^{-6}}{(2)(18.6)(10) \ln 0.167}.$$
$$= .0015 \ \mu\text{f}.$$

The output amplitude of the waveform is given by (17):

$$|E_0|_{\max} = \frac{1 - \sqrt{\dfrac{1.8\text{k}}{5.6\text{k}}}}{1 + \sqrt{\dfrac{1.8\text{k}}{6.5\text{k}}}}[12] = 3.7 \text{ volts}.$$

It should be noted that (17) does not include the variation of the double-base diode parameters with frequency. Experiment shows that the outlined analysis is accurate to within $\pm 10$ per cent for frequencies up to approximately 20 kc. Waveforms obtained from experimental circuits are shown in Figs. 12 and 13.

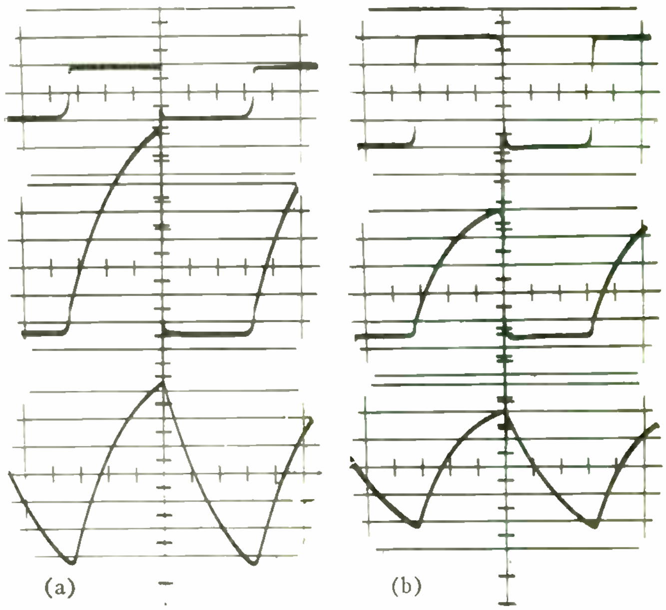

Fig. 12—Experimental waveforms: (a) germanium dbd—1 kc; (b) silicon dbd—10 kc.

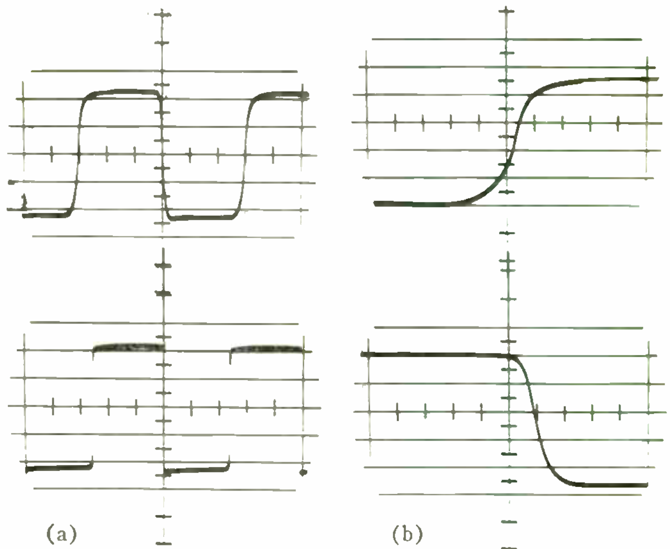

Fig. 13—Silicon double-base diode waveforms: (a) top—55 kc, bottom—7 kc; (b) top—rise time, bottom—fall time, full sweep $= 6\ \mu$sec.

## Pulse-Delay Generator

The diode multivibrator may be made monostable if

$$\frac{R_1 E}{R_1 + R_2} < V_p \tag{22a}$$

and

$$E/R_2 < I_v. \tag{22b}$$

Eq. (22a) fixes the stable operating point of the double-base diode in the cut-off region and (22b) insures that this is the *only* stable operating point. If (22a) and (22b) are satisfied, a positive pulse will trigger the double-base diode from the "off" state to the "on" state. The double-base diode will then remain conductive until the capacitor discharges through $R_1$. When the diode reverses at the end of the capacitor discharge cycle, the double-base diode becomes nonconductive. But since it is stable in cut-off state, the multivibrator circuit remains stable until the next positive trigger pulse is applied. Thus, the regenerated output waveform duration is

$$t_D = -R_1C \ln \frac{V_v'}{V_p} \qquad (23)$$

[see (9)]. Fig. 14 shows the waveform generated by monostable multivibrator. Minimum spacing of the trigger pulses is obviously limited by the circuit's time constants.

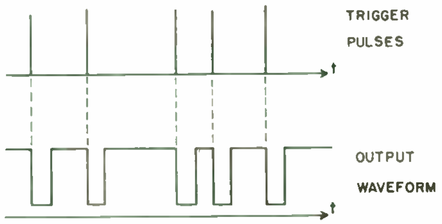


Fig. 14—Monostable multivibrator waveform.

On the other hand, if

$$\frac{R_1E}{R_1+R_2} > V_p \qquad (24a)$$

$$\frac{E}{R_2} > I_v, \qquad (24b)$$

a monostable circuit, having a stable operating point associated with the conductive state of the double-base diode, is obtained. Negative pulses may then be used to trigger the circuit into its regenerative cycle and the output waveform duration will be given [see (8)] by,

$$t_D = -\frac{R_1R_2C}{R_1+R_2} \ln \left[ \frac{1 - \left(\frac{V_p}{E}\right)\left(\frac{R_1+R_2}{R_1}\right)}{1 - \left(\frac{V_v'}{E}\right)\left(\frac{R_1+R_2}{R_1}\right)} \right]. \qquad (25)$$

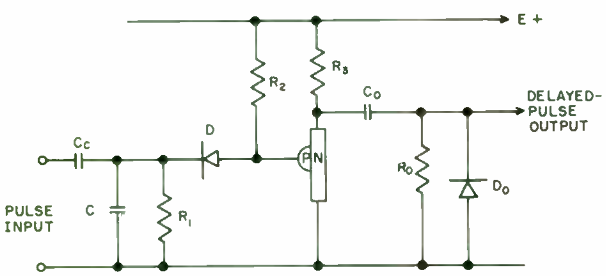


Fig. 15—Delayed-pulse generator.

Use of the monostable multivibrator as a delayed-pulse generator is illustrated in Fig. 15. In this circuit, $R_D$ and $C_D$ are used as a simple differentiating network and diode $D_D$ filters out the pulses of unwanted polarity. The output waveform consists of pulses which are generated by differentiating the trailing edge of the multivibrator output. Diode $D_0$ filters out the pulses which are generated by the leading edge of the multivibrator waveform. Hence, the output of the delayed-pulse generator consists of a train of pulses which have the same polarity and repetition rate of the input pulses but which are delayed in time by an interval, $t_D$, determined by the time constants of the monostable circuit.

Fig. 16 illustrates the relationship between the pulse delay $t_d$ and the magnitude of the multivibrator capacitor $C$, which was obtained for an experimental delayed-pulse generator circuit. In the circuit of Fig. 16, the conditions defined by (22a) and (22b) are required for operation. Time delays from 50 $\mu$sec to 2 msec have been obtained for pulse repetition rates from 0–5 kc. As is apparent from Fig. 16, the time delay is related to the magnitude of the capacitor $C$ in a linear manner. This relationship is convenient in design and facilitates constructing very simple variable-delay pulse generators.

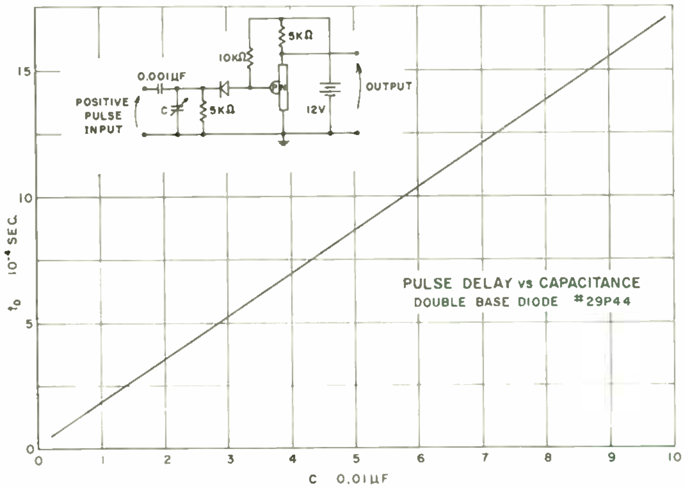


Fig. 16—Experimental delayed-pulse generator characteristics.

## Conclusion

The diode multivibrator consists of three resistors, one capacitor, one diode and one double-base diode. When compared to the corresponding Eccles-Jordan transistor circuit, the diode configuration affords a two-to-one economy in circuit components. Furthermore, since the diode multivibrator consists of one diode and one double-base diode, as compared to the two transistors required for the Eccles-Jordan circuit, an additional cost economy may be achieved.

Such factors as circuit simplicity, easy designability and component and device economy make the diode multivibrator a considerable competitor to the corresponding transistor circuit. The advantages of the diode circuit should be particularly significant in such complex systems as digital computers and counters where component cost and network complexity can be restrictive.

## Acknowledgment

The authors wish to thank Mr. C. M. Hackett for his laboratory assistance.

# The Effect of the Source Distribution on Antenna Patterns*

S. MATT†, ASSOCIATE MEMBER, IRE, AND J. D. KRAUS‡, FELLOW, IRE

*Summary*—**The response pattern of an antenna is modified when the source subtends an appreciable angle. Under these conditions the observed or resultant pattern is a function of both the true antenna pattern and the source distribution. This problem is important in radio astronomy and in radar. The general problem is formulated and solved for the particular case of a radio telescope antenna consisting of a 40 wavelength broadside array. Graphs are presented showing the effect of sources of various widths on the observed patterns. Using such graphs it is possible to deduce the approximate source extent from an observed pattern. The direct analytical solution for the source distribution from the observed pattern is also considered and the limitations of the various methods are discussed.**

## Introduction

THE TRUE RESPONSE pattern of a receiving antenna is obtained when the radiator is a point source situated at a sufficient distance from the antenna. The distance is sufficient if an increase in the distance produces no detectable change in the pattern. Let the true pattern of a receiving antenna be as shown in Fig. 1(a). If the point source is replaced by an extended source at the same distance, the observed pattern is modified as suggested in Fig. 1(b).

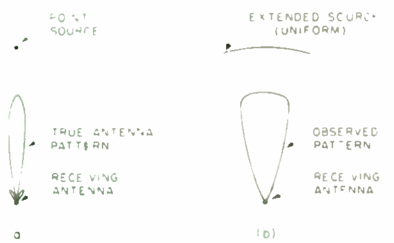


Fig. 1—(a) Antenna pattern for a point source, (b) for an extended source.

In radio astronomy many of the celestial sources are of sufficient angular extent to modify the observed pattern and it is desirable to be able to deduce the source distribution (extent and shape) from the observed pattern. A similar situation exists in radar when the object under observation is of sufficient angular extent.

Referring to Fig. 2, the general problem of the effect of the source distribution on the observed antenna pattern may be stated as follows:

$$G(\phi_0) = \frac{1}{A}\int F(\phi + \phi_0)f(\phi)d\phi; \qquad (1)$$

where

$G(\phi_0)$ = observed or resultant pattern,

$F(\phi+\phi_0)$ = true antenna pattern (as measured with a point source),

$f(\phi)$ = source distribution,

$A = \int f(\phi)d\phi$ = effective angle subtended by source.

All patterns in (1) are proportional to power.

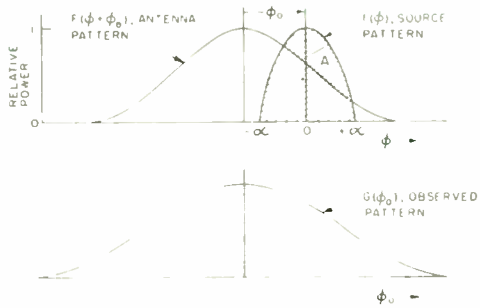


Fig. 2—Antenna pattern, source pattern, and resultant or observed pattern.

In (1) the general problem has been simplified to the one-dimensional case where the patterns are functions only of one co-ordinate. For the purpose of the following discussion this has the advantage that the fundamental problem is retained intact but the analysis is greatly simplified. The simplified situation stated in (1) often occurs in practice as, for example, when the antenna pattern in the direction ($\theta$), normal to $\phi$, is sufficiently broad compared to the source extent in this direction. In (1) the antenna pattern, $F(\phi+\phi_0)$, and observed pattern, $G(\phi_0)$, are usually known while the source distribution, $f(\phi)$, is unknown. To determine the source distribution it is necessary to solve the integral equation (1). This can be done by assuming various source distributions and calculating the corresponding distributions, $G(\phi_0)$. If a calculated $G(\phi_0)$ distribution can be obtained that agrees with the actual observed distribution $G(\phi_0)$, one can conclude that the assumed source distribution $f(\phi)$ used for this case represents true source distribution or its equivalent. Or one can solve the integral equation in a straightforward manner. This is usually the more difficult procedure. The indirect and direct methods of solution will be discussed in that order.

It is assumed that the cosmic signals under consideration are incoherent so that the power received is proportional to the incident power flux from the source integrated over the solid acceptance angle of the antenna.

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Eq. (1) applies not only to antenna problems but to many other situations where instead of an antenna pattern there is a modifying function $F(\phi+\phi_0)$ which disturbs the actual distribution $f(\phi)$ so as to yield the observed function $G(\phi_0)$.

## Cases of Point Source and Large Extended Source

Referring to Fig. 2 the value of the observed distribution $G(\phi_0)$, when the main lobe of the antenna is displaced by an angle $\phi_0$ from the center line of the source, is given by

$$G(\phi_0) = \frac{1}{A}\int_{-\alpha}^{+\alpha} F(\phi + \phi_0)f(\phi)d\phi, \qquad (2)$$

where

$$A = \int_{-\alpha}^{+\alpha} f(\phi)d\phi.$$

The over-all source extent is $2\alpha$. The distributions are all power patterns so that $A$, the area under the source pattern, is in fact the total power flux of the source.

For a *point source* the source pattern in Fig. 2 collapses to a single vertical spike at $\phi=0$ ($\alpha=0$). The observed pattern is then given by (2) with $f(\phi)=0$ except at $\phi=0$. For this case (2) reduces to $G(\phi_0)=F(\phi_0)$ or, in general,

$$G(\phi) = F(\phi). \qquad (3)$$

Thus, for a point source the observed pattern is identical with the true antenna pattern.

At the other extreme let us consider the case of an *extended source* that is much wider than the antenna pattern as suggested in Fig. 3. Let the source be a step

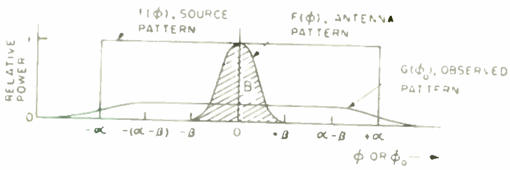


Fig. 3—Case of source pattern that is much wider than antenna beam width.

function equal to unity between $+\alpha$ and $-\alpha$ and zero elsewhere. The resulting observed pattern $G(\phi_0)$ from (2) is as shown in the figure. It is to be noted that in the range of $\phi_0$ between $\alpha-\beta$ and $-(\alpha-\beta)$ the observed distribution is a constant like the source although reduced by a factor $B/A$, where $B$ is the area under the antenna pattern and $A$ is the area under the source pattern ($=2\alpha$).

## O.S.U. Radio Telescope Antenna

The preceding cases are idealized. Turning now to the case of an actual antenna, such as the Ohio State University radio telescope antenna, let us consider the effect of the source distribution for three conditions of antenna operation: (1) All elements in phase (single-lobe pattern), (2) two halves of antenna in phase opposition (split-lobe pattern), and (3) comparison arrangement [resultant pattern equal to the difference of (1) and (2)].

The O.S.U. radio telescope antenna, shown by the photograph in Fig. 4, consists of an array of 96 helical beam antennas mounted on a steel ground plane 160 feet long (east-west) by 22 feet wide.[1] The entire antenna pivots like a meridian transit instrument. At 250 mc the antenna is approximately 40 wavelengths long by 5.6 wavelengths wide. With all helices in phase the beamwidths at 250 mc are approximately 1 degree in right ascension (east-west) by 8 degrees in declination.

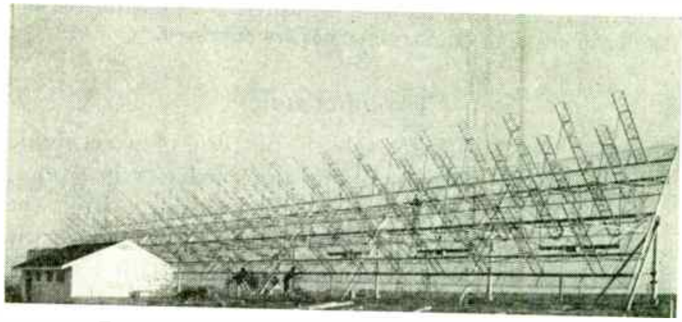

Fig. 4—Photograph of the Ohio State University radio telescope antenna.

Although the antenna can be operated at frequencies between 200 and 300 mc the patterns at only the center frequency of 250 mc are discussed in this article. All helices are right-handed so that the antenna is responsive to the right circularly polarized component of the incident radiation, which is usually of an incoherent nature. In operation the antenna is set at a fixed declination and as the earth rotates a trace or profile is obtained on the recorder as a celestial radio source crosses the meridian. This recorder profile is the observed pattern $G(\phi)$. The helices are arranged in 4 rows with 24 helices in each row. The long (24-helix) rows are east-west and determine the pattern of the antenna in right ascension while the short (4-helix) rows are at right angles and determine the pattern in declination. Since the beamwidth in declination is sufficiently wide (8 degrees) it will be convenient to reduce the problem to the one-dimensional case and consider only the right ascension pattern and the effect of the source shape on it.

The east-west rows of 24 helices are as shown in Fig. 5(a) with a uniform spacing $d$ between helices. The total field pattern of this array is given by the product of the individual helix pattern and the pattern of an array of 24 isotropic point sources with a spacing $d$ (array pattern). However, the array pattern is so much sharper than the helix pattern that for angles near the meridian (broadside to the antenna array) total antenna

[1] J. D. Kraus and D. Van Stoutenburg, "A High Resolution Radio Telescope," (in preparation).

pattern is substantially the same as array pattern. The array has a uniform amplitude distribution.

*Case with All Helices in Phase (Single-Lobe Pattern)*

Considering first the case where all helices are in phase the normalized field pattern is substantially that of a linear array of 24 isotropic point sources of equal amplitude and spacing as given by[2]

$$E_A = \frac{1}{n}\,\frac{\sin\dfrac{n\psi}{2}}{\sin\dfrac{\psi}{2}}\,; \qquad (4)$$

where

$\psi = (2\pi d/\lambda)\sin\phi_H$,
$d$ = spacing between helices (= 1.69 wavelengths at 250 mc),
$n$ = 24 (number of helices in a row in $\phi$ direction),
$\phi_0$ = angle from meridian to center of source,
$\lambda$ = wavelength.

The normalized power pattern of the antenna is equal to $E_A^2$. This is the true antenna pattern $F(\phi)$ as shown in Fig. 5(b). Let the source distribution be a step function equal to unity between $+\alpha/2$ and $-\alpha/2$ and zero elsewhere. For this case (2) then becomes

$$G(\phi_0) = \frac{1}{\alpha}\int_{-\alpha/2}^{+\alpha/2} \frac{\sin^2\left[\dfrac{nd_r}{2}\sin(\phi+\phi_0)\right]}{n^2\sin^2\left[(d_r/2)\sin(\phi+\phi_0)\right]}\,d\phi, \quad (5)$$

where $d_r = 2\pi d/\lambda$ = spacing of helices in radians.

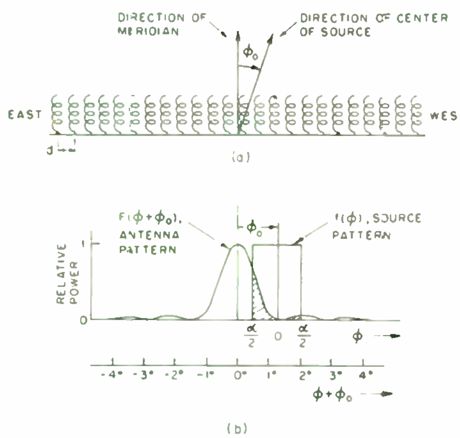


Fig. 5—(a) 24-helix array, (b) antenna pattern with assumed source pattern.

Eq. (5) for the observed distribution $G(\phi_0)$ can be obtained from (5) by graphical integration [by measuring shaded area in Fig. 5(b)], or (5) can be evaluated analytically. Proceeding with the latter solution one can put $\sin(\phi+\phi_0) = \phi+\phi_0$ provided both $\phi_0$ and $\alpha$ are small. Introducing this approximation into (5) and integrating yields

$$G(\phi_0) = \frac{1}{n} + \frac{2}{n^2}\sum_{m=1}^{n-1}\frac{m\sin\left[(n-m)d_r\dfrac{\alpha}{2}\right]}{(n-m)d_r\dfrac{\alpha}{2}}\cdot\cos\left[(n-m)d_r\phi_0\right]. \qquad (6)$$

By evaluating (6) for various values of $\alpha$ the curves of Fig. 6 were obtained. These show the effect of the source extent $\alpha$ on the observed pattern $G(\phi_0)$ of the O.S.U. radio telescope antenna. These curves have been normalized (maximum value set equal to unity). For small source extent ($\alpha$ small) the observed pattern is nearly the same as the true antenna pattern ($\alpha = 0$), while for larger source extent the observed pattern tends to conform more to the source shape. If the source distribution is a step function but $\alpha$ is not known, one can deduce its value from Fig. 6 provided $\alpha > \frac{1}{2}$ degree.

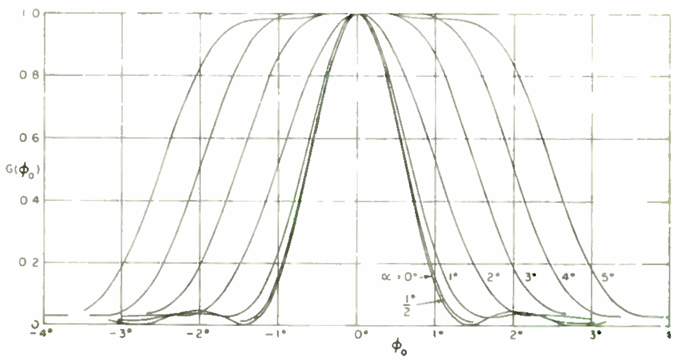


Fig. 6—Single-lobe patterns that would be observed with 40-wavelength broadside array for assumed uniform source distributions of various angular extent ($\alpha$). The sharpest pattern is for the case of a point source ($\alpha = 0$).

If $\alpha < \frac{1}{2}$ degree the observed pattern differs so little from the antenna pattern that it is impractical to deduce its value unless the source is sufficiently strong for the small amount of broadening to be accurately measured. It is also to be noted in Fig. 6 that with increase in $\alpha$ the minor lobe amplitude of the observed distribution tends to decrease. Since the array is a long one the curves of Fig. 6 apply not only to the 24-helix array but also approximately to any uniform rectangular broadside array or aperture 40 wavelengths across.

*Case of Two Halves of Array in Phase Opposition (Split-Lobe Pattern)*

If the helices to the right of the center of the array are reversed in phase with respect to those to the left, the total field pattern is given closely by

$$E_A = \frac{1}{n'}\,\frac{\sin^2\dfrac{n'\psi}{2}}{\sin\dfrac{\psi}{2}} = \frac{1}{n}\,\frac{\left[1-\cos\dfrac{n\psi}{2}\right]}{\sin\dfrac{\psi}{2}}\,; \qquad (7)$$

where

$\psi$ = same as in (4), $n' = 12$, and $n = 24$.

[2] J. D. Kraus, "Antennas," McGraw-Hill Book Co., Inc., New York, N.Y., p. 78; 1950.

Squaring (7) gives the power pattern. Introducing this in (2) and integrating yields, for $n$ even,

$$G(\phi_0) = \frac{1}{n} - \frac{2}{n^2}\sum_{m=1}^{n-1} m \frac{\sin\left[\frac{(n-m)}{2} d_r\alpha\right]}{\frac{n-m}{2} d_r\alpha} \cdot \cos\left[(n-m)d_r\phi_0\right] + \frac{8}{n^2}\sum_{m=1}^{(n-2)/2} m \frac{\sin\left[\frac{(n-2m)}{4} d_r\alpha\right]}{\frac{(n-2m)}{4} d_r\alpha} \cdot \cos\left[\frac{(n-2m)}{2} d_r\phi_0\right]. \quad (8)$$

Evaluating (8) for various values of $\alpha$, curves of Fig. 7 were obtained. These show effect of uniform source distributions of width $\alpha$ on observed pattern. Curves have been normalized. Note that the value of curves in Fig. 7 at minimum occurring at $\phi_0=0$ is an effective indicator of angular extent $\alpha$ of source for $\alpha$ values between about $\frac{1}{2}$ degree and 2 degrees.

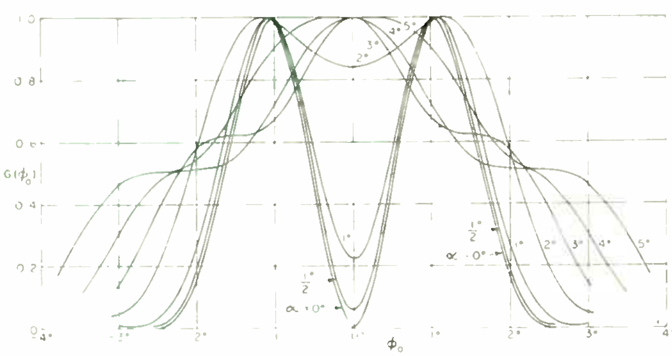


Fig. 7—Split-lobe patterns that would be observed with 40-wavelength broadside array with two halves in phase opposition for assumed uniform source distributions of various angular extent ($\alpha$).

For a source of small extent more large lobes of reduced beamwidth will be obtained as the spacing between the two halves of the antenna is increased. Under these conditions the two halves of the antenna act as the two units of a Michelson interferometer. If the spacing is increased sufficiently the minimum to maximum ratio of the observed lobe amplitude will tend to approach unity. Further increase in spacing will cause fluctuations in the ratio. By using an interferometer with a number of different spacings[3,4] it is theoretically possible to deduce the source distribution with approximately the same accuracy as an array (with single-lobe pattern) extending continuously over a distance equal to the largest spacing between the interferometer units.

### *Case of Comparison Arrangement*

In this case the antenna and receiving system are operated in such a manner[5] that the difference of the single-lobe and split-lobe patterns is obtained.[6] Fig. 8 shows the curves for this case for various values of $\alpha$. Each curve is obtained by applying a scale factor to a curve of Fig. 7 and then subtracting this curve from corresponding curves of Fig. 6. The scale factor adjusts corresponding curves of Figs. 6 and 7 to the same scale.

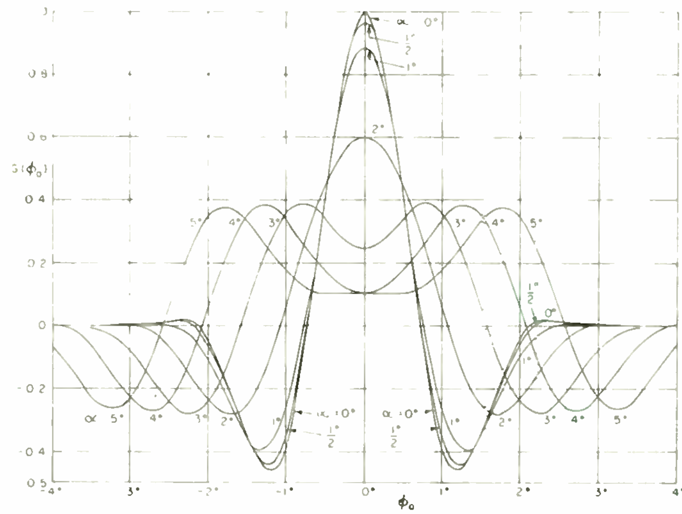


Fig. 8—Patterns that would be observed with 40-wavelength broadside array for the comparison method of operation.

The location of the zeros for the curves of Fig. 8 is independent of receiver gain. This is an advantage. Furthermore the slope is a maximum at the zero points so that these can be used to obtain an accurate time of transit (average of the two zero points), giving precise position data for the celestial source.

## Direct Solution

Finally, in the most general situation where both antenna and source patterns are two-dimensional distributions on a sphere surface, the problem can be stated

$$G(\phi_0, \theta_0) = \frac{1}{\Omega_0}\iint F(\phi, \theta) f(\phi, \phi_0, \theta, \theta_0)\, d\Omega; \quad (9)$$

where

$G(\phi_0, \theta_0)$ = observed or resultant distribution,
$F(\phi, \theta)$ = true antenna pattern,
$f(\phi, \phi_0, \theta, \theta_0)$ = source distribution,
$\Omega_0$ = equivalent solid angle subtended by source.

In (9), the observed and true antenna patterns are usually known, while the source distribution is unknown and is the quantity desired. The source distribution can be obtained indirectly, as done in the special cases above,

[3] L. L. McCready, J. L. Pawsey, and R. Payne-Scott, "Solar radiation at radio frequencies and its relation to sunspots," *Proc. Roy. Soc.*, vol. A 190, pp. 357–375; August, 1947.
[4] B. Y. Mills, "The radio brightness distributions over four discrete sources of cosmic noise," *Aust. Jour. Phys.*, vol. 6, p. 452; 1953.
[5] M. Ryle, "A new radio interferometer and its application to the observation of weak radio stars," *Proc. Roy. Soc.*, vol. A 211, pp. 351–375; March, 1952.
[6] J. D. Kraus and E. Ksiazek, "New techniques in radio astronomy," *Electronics*, vol. 26, pp. 148–152; September, 1953.

by assuming various source distributions and calculating the corresponding observed distributions. Then by comparison with the actual observed distribution, the source distribution or its equivalent can be deduced.

The source distribution can also be obtained by a direct solution of (9) for $f(\phi, \phi_0, \theta, \theta_0)$. This may be done by expanding the patterns into sets of orthogonal functions. For example, in the one-dimensional case the distributions can be expanded into Fourier series.[7] Thus, the source distribution can be written as

$$f(\phi - \phi_0) = a_0 + \sum_{m=1}^{\infty} a_m \cos m(\phi - \phi_0) + d_m \sin m(\phi - \phi_0), \tag{10}$$

the observed distribution as

$$G(\phi_0) = b_0 + \sum_{n=1}^{\infty} b_n \cos n\phi_0 + e_n \sin n\phi_0, \tag{11}$$

and the antenna pattern (symmetrical) as

$$F(\phi) = c_0 + \sum_{p=1}^{\infty} c_p \cos p\phi. \tag{12}$$

Then

$$G(\phi_0) = \int_{-\pi}^{\pi} f(\phi - \phi_0)F(\phi)d\phi = \int_{-\pi}^{\pi} a_0c_0 d\phi + \int_{-\pi}^{\pi} [\sum c_p \cos p\phi \sum (a_m \cos m(\phi - \phi_0) + d_m \sin m(\phi - \phi_0))]d\phi. \tag{13}$$

Integrating,

$$G(\phi_0) = 2\pi a_0 c_0 + \pi \sum_{m=1}^{\infty} c_m(a_m \cos m\phi_0 + d_m \sin m\phi_0). \tag{14}$$

Equating (14) and (11) term by term, the coefficients of the source distribution are found to be

$$a_0 = \frac{b_0}{2\pi c_0}, \qquad a_n = \frac{b_n}{\pi c_n}, \qquad d_n = \frac{e_n}{\pi c_n}. \tag{15}$$

A necessary and sufficient condition for there to be a solution for the source distribution is that $b_n$ and $e_n$ be zero for all $n$ for which $c_n$ is zero and that[8]

$$\sum_0^{\infty} \left|\frac{b_n}{c_n}\right|^2 + \left|\frac{e_n}{c_n}\right|^2 < \infty. \tag{16}$$

The significance of the $b_n$ and $e_n$ coefficients being zero for all $n$ for which $c_n$ is zero is to limit the components of the Fourier series of the actual observed distribution to those components which are contained in the series expansion of the antenna pattern $F(\phi)$. That is, if the series expansion of the source distribution contains terms which do not appear in the series expansion of $F(\phi)$, then the solution is not unique. Actually the antenna will not respond to variations of the source distribution whose period is less than that required for $\frac{1}{2}$ the beamwidth between first nulls (approximately equal to the half-power beamwidth).

As an example of this method it will be applied to find the source distribution where the antenna pattern $F(\phi)$ and the observed distribution $G(\phi)$ are given. Let the antenna pattern be that of the O.S.U. antenna as shown by the $\alpha = 0$ curve of Fig. 6, and let $G(\phi)$ be the $\alpha = 2$ degree pattern in the same figure. Proceeding by the above analytical (Fourier) method the source distribution shown by the solid curve in Fig. 9 was obtained. The assumed distribution is given by a step function (dashed line).

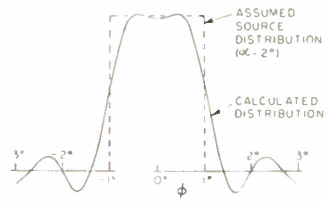


Fig. 9—Assumed source distribution compared with calculated distribution.

The Fourier series for $F(\phi)$ and $G(\phi)$ in the above calculation were obtained graphically by the 24-ordinate method over a range of $\phi$ between $+3$ degrees and $-3$ degrees. The calculated distribution plotted in Fig. 9 contains terms to the 4th harmonic of the Fourier series. The higher harmonic terms introduced by the rectangular distribution cannot be determined because of the lack of response of the antenna to these terms. Thus, the antenna tends to smooth out the source variations.[9,10] However, amplitude and equivalent rectangular extent, as measured at half-power points, are indicated properly by calculated distribution.

It should be noted that for sources of small angular extent where the observed and antenna patterns differ almost imperceptibly it becomes impractical to deduce the source distribution with any certainty (except to state its maximum possible extent).

## Acknowledgment

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[7] S. Matt, "Some Characteristics of Waves of Extraterrestrial Origin," Ph.D. dissertation, Dept. of Electrical Engrg., Ohio State University, Columbus, Ohio; 1953.

[8] H. S. Carslaw, "Fourier's Series and Integrals," 3rd ed., Dover Publications Inc., New York, p. 285; 1930.

[9] R. N. Bracewell and J. A. Roberts, "Aerial smoothing in radio astronomy," *Aust. Jour. Phys.*, vol. 7, pp. 615–640; December, 1954.

[10] P. Elias, "Optics and communication theory," *Jour. Opt. Soc. Amer.*, vol. 42, p. 869; August, 1952.

# Nonsaturating Pulse Circuits Using Two Junction Transistors*

J. G. LINVILL†, ASSOCIATE, IRE

*Summary*—Junction transistors have been supposed to be too slow for many pulse applications. However, if they are used in a way in which their collector voltage is never permitted to become zero, saturation does not occur and the switching times achieved may be as low as several times the reciprocal of $\omega_0$, the radian cut-off frequency of $\alpha$. This time will be less than a microsecond for junction triodes presently available.

Saturation is prevented through use of breakdown diodes which terminate switching transients by their breakdown. They may also serve other functions in the circuit. A binary counter described has stable points dependent upon the breakdown diodes and passive components, these points being virtually independent of the transistor or its temperature-sensitive $I_{c0}$.

A class of two-transistor pulse circuits is described including a binary counter which with one kind of junction triode operates at 1.25 mc. Monostable and astable circuits of the same general nature are shown. The pulse requirements for switching of the binary counter are analyzed in some detail. It is shown that the switching charge is the significant quantity and that a crude estimate of the charge required is $1/\omega_0$ times the difference in conduction currents of the two transistors in the stable state.

## Introduction

JUNCTION transistors have many attributes which are of importance in pulse applications. Their greater designability as devices over point-contact units is important in pulse work as it is elsewhere. The low values of saturation current obtainable in junction units is of particular interest since it is the significant factor determining smallness of power level which can be employed. The principal question usually raised about the application of junction units in switching is the question of their supposed low speed. Analysis indicates and experiment verifies that pulse circuits using junction transistors currently available can switch in fractions of a microsecond provided that one prevents the collector voltages from going to zero at the terminus of each switching operation. It has been shown[1] that this condition of saturation floods the base region with minority carriers. The recovery from saturation can be time consuming and can enormously slow operating speeds. R. L. Wallace suggested that one could avoid saturation with two breakdown diodes placed in the circuit in such a way that the switching transient is terminated by their breakdown rather than transistor saturation.[2] This technique has been employed in a binary counter which operates on pulses occurring at 1.25 mc with experimental *n-p-n* alloy-junction[3] transistors. A similar circuit using grown-junction[4] triode transistors operated at a 600 kc rate. In this circuit both transistors are continuously in their active region; they never reach either zero emitter current or zero collector voltage. Thus the role played by the transistors is essentially that of a linear amplifier, the necessary nonlinear functions being performed by breakdown diodes.

Through the use of two additional breakdown diodes one can fix the stable points of a bistable circuit virtually independent of transistors or of the temperature sensitive $I_{c0}$ (saturation current) of the transistors. Moreover, similar techniques to those applicable to bistable circuits can be applied to monostable and astable circuits with the result that similar transition times are obtained and the pulse height can be accurately set.

## Two-Transistor Pulse Circuits

The elements shown in Fig. 1 are all found in the vast majority of two-transistor pulse circuits. Additional elements may be found in some embodiments but those shown are essential.

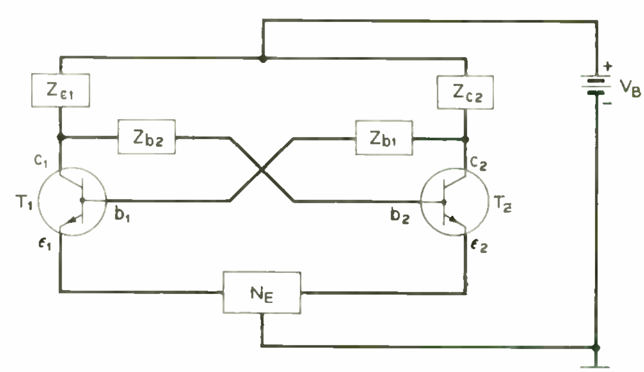


Fig. 1—Usual form of two-transistor pulse circuit.

All pulse circuits have one characteristic in common. It is that in certain conditions they are unstable. This simply means that for these conditions transient currents and voltages exhibit a growing or self-perpetuating behavior. In pulse circuits the attribute of instability is associated with the self-completion of a switching operation once the circuit is triggered appropriately. The process of triggering to initiate the switching operation successfully involves bringing the circuit into its unstable region and leaving initial conditions at the end of the switching pulse such that the circuit completes whatever is left of the switching transient "on its own."

* Original manuscript received by the IRE, January 17, 1955; revised manuscript received, March 2, 1955.

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

[1] J. L. Moll, "Large-signal transient behavior of junction transistors, Proc. IRE, vol. 42, pp. 1773–1784; December, 1954.

[2] A. W. Carlson and others have earlier used fast-recovery diodes to prevent transistors from saturating.

[3] Alpha cut-off frequency is about 10 mc.

[4] Alpha cut-off frequency is 2–5 mc.

With reference to Fig. 1, the instability mentioned above is physically apparent. The circuit, like all trigger circuits of the Eccles-Jordan variety, is essentially a two-stage amplifier with positive feedback.

A second characteristic common to all pulse circuits is that growing transients are always terminated by a change in the characteristics of some component. In the circuits to be described here this component is a breakdown diode (not shown in Fig. 1). The volt-ampere characteristic of an idealized breakdown diode is shown in Fig. 2. In region *b* the diode approximates an open circuit, but as the voltage is brought to the breakdown point $V_{bd}$, the diode's incremental resistance approximates a short circuit. In the circuits described in the following it is the diode voltage attaining $V_{bd}$ which terminates the unstable transient. For these circuits when the operating point of the diode is in range "*a*" the circuits are stable, that is, all of their natural modes

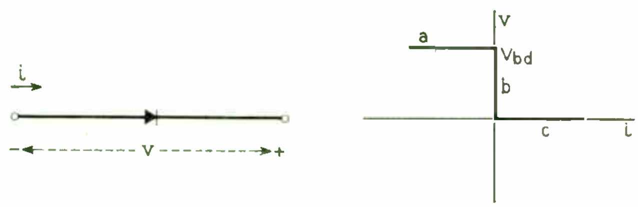


Fig. 2—Volt-ampere characteristic of idealized breakdown diode.

correspond to decaying transients. When the operating point is in range "*b*," at least one of the natural modes corresponds to a growing transient. The remainder of the circuit is essentially linear. In fact two sets of linear analyses, one for the diodes as an open circuit and one for the diodes as short circuits give results in substantial agreement with experiment.

With respect to the pulse circuit shown in Fig. 1, breakdown diodes connected from the collector terminals $c_1$ and $c_2$ to $B+$ or ground can serve to terminate a growing transient. In general the growing transients of circuits of this type tend to increase one collector voltage and decrease the other. Thus a diode connected to ground from a collector prevents the increase of collector voltage beyond its breakdown point. This prevention ordinarily limits the corresponding decrease in collector voltage of the companion transistor. In a similar way a breakdown diode connected from the collector to $B+$ in the proper orientation terminates the fall of collector voltage at $V_B - V_{bd}$. The connection of two breakdown diodes in the arrangement shown in Fig. 3 placed between $c_1$ and $c_2$ in Fig. 1 terminates switching transients when the difference in collector voltages rises to $V_{bd}$ in either direction. (This suggestion was due to R. L. Wallace.) This particular arrangement has the desirable feature that it can also be used for the additional function (described in detail later) of "pulse routing" in a binary counter.

The diagram of Fig. 1 can represent either bistable, monostable or astable circuits. Moreover, it is possible to employ breakdown diodes in all of these types to prevent the transistors from going into the saturated condition. The distinction between these is simply whether or not the circuit maintains the broken-down diode in its low impedance condition once the switching has taken place. Bistable circuits can remain permanently in either of two states, monostable circuits can remain permanently in only one state and astable circuits have no stable condition in which they will remain.

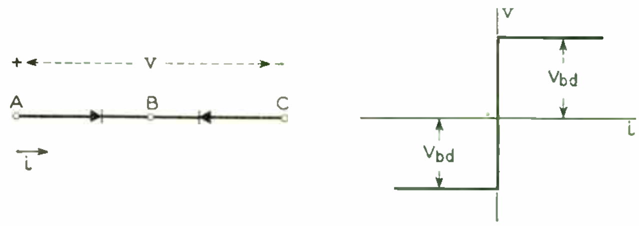


Fig. 3—Connection of two breakdown diodes and volt-ampere characteristic.

In any case switching, or unstable, transients occur which terminate with the breakdown of a diode. Until the diode breaks down, at least one of the natural frequencies of the circuit corresponds to a growing transient. Once the diode has broken down to terminate the switching, all of the natural frequencies correspond to decaying transients. If the circuit is not to remain in this stable condition, the current in the diode decays to zero and at this point a new switching transient ensues.

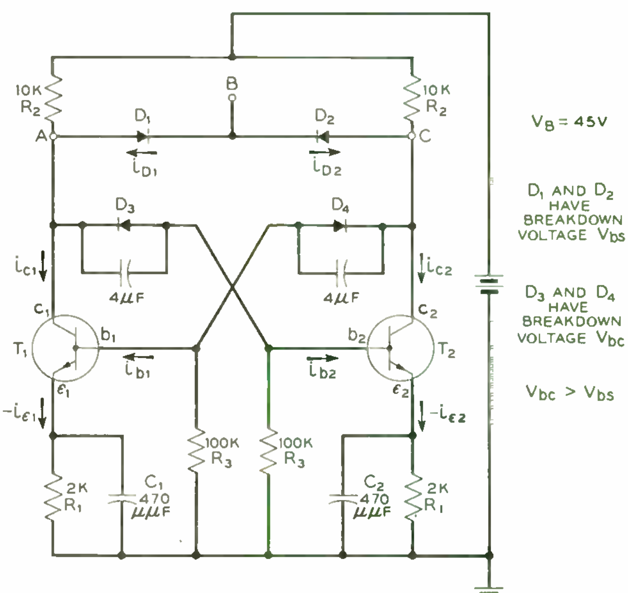


Fig. 4—A nonsaturating binary counter.

## Analysis of Binary Counter

A two-transistor nonsaturating binary counter of the general type described in the foregoing section is shown in Fig. 4. In the two stable states the circuit will maintain different collector voltages by the amount of $V_{bs}$, that is, one of the two diodes $D_1$ and $D_2$ will be broken down, the other conducting in the forward direction. Diodes $D_3$ and $D_4$ are continuously broken down and hence with their bypass condensers maintain constant potential drops between the points $c_1$ and $b_2$ and the points $c_2$ and $b_1$. The collector voltage of the high-conducting transistor will be $V_{bc} - V_{bs}$ ($>0$) and the collector voltage of the low-conducting transistor will be $V_{bc} + V_{bs}$. When a switching pulse of either polarity is applied at point $B$, transistors $T_1$ and $T_2$ exchange the roles of high and low conduction. Thus the circuit acts as a binary counter in that either transistor assumes a given state alternately as pulses are applied. Binary counters ordinarily require auxiliary diode circuits called routing circuits which direct an incoming pulse to the appropriate point to turn on the low-conducting unit and to turn off the high-conducting unit. The diodes $D_1$ and $D_2$ serve this function in addition to their function of preventing saturation of transistors $T_1$ and $T_2$.

The interesting properties in the application of binary counters are the switching time, pulse requirements for switching and the dependence of the counter's behavior on changes in the transistor with temperature or exchanges of transistors. These properties can be evaluated for the circuit of Fig. 4 quite directly. Moreover the same analysis with only minor change applies to monostable and astable circuits as well.

## Analysis of Switching Time of Binary Counters

The speed with which the switching of a binary counter is effected is substantially dependent upon the natural frequencies of the circuit when it is in the unstable state. A switching pulse will have been applied and will have caused the diodes $D_1$ and $D_2$ (Fig. 4) both to become open circuits in a manner considered in detail in the next section. At the point that they become open circuits (arrive at range $b$, Fig. 2) the circuit will have certain initial conditions which don't change instantaneously. Hence the evaluation of the switching time of the circuit of Fig. 4 is really a computation of its transient response employing the initial values of the pertinent variables. This analysis can be particularly simple if one makes approximations in the representation of the components. Subsequent to the analysis in the simple form the accuracy of the approximations can be assessed. An approximant to the circuit of Fig. 4 for transients is shown in Fig. 5. In Fig. 5, the transistor is approximated as having negligible base and emitter resistance and collector conductance. The differential equation applied to the collector current [see Fig. 5(b)] is a first approximation to the diffusion equation for the transistor. The alpha cut-off frequency is $\omega_0$ radians per second. From the equation in Fig. 5 one sees that collector currents cannot change instantaneously with finite emitter current. Hence for the circuit starting a transient, the pertinent initial conditions are the collector currents, $i_{c1}$ and $i_{c2}$ and the voltages on the condensers connected to the emitters.

The circuits shown in both Fig. 4 and Fig. 5 are symmetrical. In such circuits one can effect an economy in analysis through considering symmetrical and anti-symmetrical components of the variables. Thus we define:

$$\begin{aligned} v_{cs} &= \frac{v_{c1} + v_{c2}}{2} & v_{ca} &= \frac{v_{c1} - v_{c2}}{2} \\ i_{cs} &= \frac{i_{c1} + i_{c2}}{2} & i_{ca} &= \frac{i_{c1} - i_{c2}}{2} \\ i_{es} &= \frac{i_{e1} + i_{e2}}{2} & i_{ea} &= \frac{i_{e1} - i_{e2}}{2} \end{aligned} \tag{1}$$

Moreover, all initial conditions and excitations are split into symmetrical and anti-symmetrical components in the same way. The economy arises since symmetrical excitations, initial conditions and variables are independent of the anti-symmetrical ones and the equilibrium equations can be written separately for each, each involving only half the total number of variables. In addition, in considering switching in pulse circuits one is principally interested in the anti-symmetrical components as these are the only ones involved when one side of the circuit changes with respect to the other.

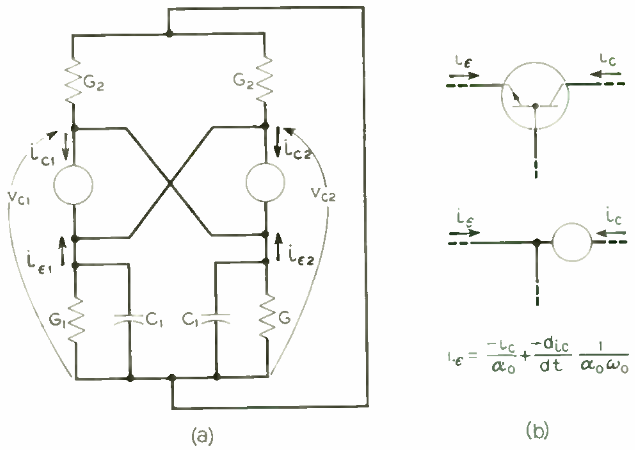


Fig. 5—Approximant of circuit of Fig. 4.

Expressing the fact that current leaving each emitter flows to the adjacent $G$-$C$ circuit of Fig. 5. one has,

$$v_{e1}G_1 + C_1\frac{dv_{e1}}{dt} = -\frac{i_{c1}}{\alpha_0} - \frac{di_{c1}}{dt}\frac{1}{\alpha_0\omega_0}, \tag{2a}$$

and

$$v_{e2}G_1 + C_1\frac{dv_{e2}}{dt} = -\frac{i_{c2}}{\alpha_0} - \frac{di_{c2}}{dt}\frac{1}{\alpha_0\omega_0}. \tag{2b}$$

Subtracting, using (1),

$$v_{ea}G_1 + C_1\frac{dv_{ea}}{dt} = -\frac{i_{ca}}{\alpha_0} - \frac{di_{ca}}{dt}\frac{1}{\alpha_0\omega_0}. \tag{2c}$$

Expressing the fact that the sum of currents leaving the cross-coupling wires must add to zero, one has,

$$v_{c1}(G_1 + G_2) + C_1\frac{dv_{e1}}{dt} - i_{c2} + i_{e1} = 0, \tag{3a}$$

and

$$v_{c2}(G_1 + G_2) + C_1 \frac{dv_{c2}}{dt} - i_{c1} + i_{c2} = 0. \tag{3b}$$

Subtracting, using (1),

$$v_{ca}(G_1 + G_2) + C_1 \frac{dv_{ca}}{dt} + 2i_{ca} = 0. \tag{3c}$$

From (2c) and (3c) one obtains the characteristic equation

$$p^2 + p\left[(1 - 2\alpha_0)\omega_0 + \frac{G_1 + G_2}{C_1}\right] + \frac{\omega_0 G_1(1 - 2\alpha_0)}{C_1} + \frac{\omega_0 G_2}{C_1} = 0. \tag{4}$$

If $G_2$ were zero, the roots of (4) become $(2\alpha_0 - 1)\omega_0$ and $-(G_1/C_1)$, the first being associated with the growing transient, the latter being simply associated with the discharge of the rc circuit at the emitter. However, if $G_2$ is present but much smaller than $G_1$, a normal situation, one finds the roots to be approximately:

$$\left.\begin{aligned} p_1 &\cong (2\alpha_0 - 1)\omega_0 - \frac{\dfrac{2\alpha_0\omega_0 G_2}{C_1}}{\omega_0(2\alpha_0 - 1) + \dfrac{G_1}{C_1}} \\ &\text{(associated with growing component)} \\ p_2 &\cong -\frac{G_1}{C_1} + \frac{\dfrac{G_2}{C_1}\left(\omega_0 - \dfrac{G_1}{C_1}\right)}{(2\alpha_0 - 1)\omega_0 + \dfrac{G_1}{C_1}} \\ &\text{(associated with decaying component).} \end{aligned}\right\} \tag{5}$$

Finally where $(G_1/C_1) \ll \omega_0$ and $\alpha_0 \to 1$, the roots of the characteristic equation become approximately

$$(2\alpha_0 - 1)\omega_0 - 2\frac{G_2}{C_1} \quad \text{and} \quad -\frac{G_1}{C_1} + \frac{G_2}{C_1}.$$

The switching time is primarily dependent upon the exponent of the growing component of the transient, $(2\alpha_0 - 1)\,\omega_0 t$ or about $\omega_0 t$. The time for the growing component of the transient to multiply itself by $\epsilon$ from the value left by the switching pulse is about $1/\omega_0$ second, which is less than 0.1 $\mu$sec for grown-junction triode transistors and smaller by several times in some of the fastest units. The time required for the growing exponential to increase by 10 times is $2.3/\omega_0$ seconds since $\epsilon^{2.3}$ is 10.

At the termination of the switching period, the growing component of the transient is ordinarily the largest component. The time required for it to grow to its size at the termination from 0.1 of that value is a good approximation of the rise time of the output pulse or the switching time. Thus the rise time should be about $2.3/\omega_0$ seconds. Further, the analysis points up the fact that $G_2$ should be as small as possible and that the emitter resistors should be bypassed, conclusions toward which one is led if he considers the circuit as a two-stage amplifier with positive feedback in which loop amplification and bandwidth should be maximized.

The circuit representation shown in Fig. 5 is optimistic in that base and emitter resistances are approximated by short circuits. The influence of these resistors is to slow the transients a bit. The resistors $R_3$ in Fig. 4 are essentially in parallel with the two $R_2$'s and these parallel combinations should be identified with $G_2$ in Fig. 5.

The switching time is dependent not only upon the natural frequencies of the circuit in its unstable state, but also upon the initial conditions left by the driving pulse. A strong pulse naturally causes the circuit to switch somewhat faster than a weak one, simply because the larger initial value of the switching transient requires less time to grow to the size required to establish the opposite stable state. More than this it is essential that the driving pulse leave initial conditions which will ultimately result in the opposite stable state to that which was in effect before the pulse. Here the final state is associated with the growing transient which with time dominates the decaying one, and one must insure that the sign of the growing component is in the proper direction. Referring to Fig. 5, one recognizes from the earlier discussion that $G_2$ should be small in comparison with $G_1$.

If one considers the limiting case when $G_2 = 0$,

$$-i_{e1} = i_{c1} - i_{c2} = \frac{i_{c1}}{\alpha_0} + \frac{di_{c1}}{dt}\frac{1}{\alpha_0\omega_0}, \tag{6a}$$

and

$$-i_{e2} = i_{c2} - i_{c1} = \frac{i_{c2}}{\alpha_0} + \frac{di_{c2}}{dt}\frac{1}{\alpha_0\omega_0}. \tag{6b}$$

Subtracting and using (1),

$$-i_{ea} = 2i_{ca} = \frac{i_{ca}}{\alpha_0} + \frac{di_{ca}}{dt}\frac{1}{\alpha_0\omega_0}. \tag{6c}$$

Since (6) includes only the variable $i_{ca}$ one can solve it separately finding that the transient is of the form

$$i_{ca} = A\epsilon^{(2\alpha_0 - 1)\omega_0 t}. \tag{7}$$

On the basis of (7) one concludes that the asymmetrical component of collector current will, after the beginning of the transient, simply increase; it always keeps the direction of the initial value. The current fed into the rc branch at the emitter terminals is proportional to $i_{ca}$. The voltage appearing there includes a component growing with this current.

The requirements on the switching pulse are now clear. It must simply cause transistor to be switched to high conduction to carry a larger collector current at termination of pulse than the companion transistor.

The excess grows as the transient develops. The approximation of $G_2$ by an open-circuit is somewhat optimistic. Since the current taken by it is small compared to that taken by $G_1$, the simple result obtained above is not seriously incorrect.

## The Mechanism of Switching

The preceding section has identified the role of the switching pulse to be the establishment of higher conduction in the transistor which after the switching transient will be the high conducting one. A complete analysis of the switching mechanism is quite complicated, but a semi-quantitative analysis of simplified circuits can show what the important factors are and attach some numerical values to them.

The circuit of Fig. 4 can be approximated by that shown in Fig. 6. The cross-coupling breakdown diodes are represented by batteries. The anti-saturation diodes are approximated by open circuits when not broken down and by batteries of voltage $V_{bs}$ when broken down. A simplified model of the transistor is employed as shown in Fig. 6.

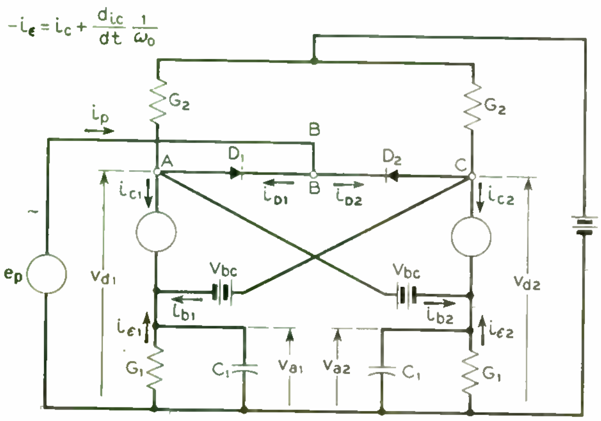


Fig. 6—Approximant of binary counter.

Before the switching pulse is applied at point $B$, one of the breakdown diodes $D_1$ or $D_2$ will be broken down, the other will be carrying a current in its forward direction. It is assumed that the pulse source supplies no current to point $B$ in the quiescent condition. For this simple model one can evaluate the stable point voltages and currents easily. It is convenient to consider symmetrical and asymmetrical components separately. The only asymmetrical voltage source is that represented by the diode which is broken down. For definiteness we take this to be $D_1$. The symmetrical component of collector current is

$$i_{cs} = \frac{V_b - V_{bc}}{\frac{1}{G_1} + \frac{1}{G_2}}. \tag{8}$$

The asymmetrical component of collector current is

$$i_{ca} = \frac{V_{bs}G_1}{2}. \tag{9}$$

The current flowing in $D_1$ is

$$i_{D1} = \frac{V_{bs}}{2}(G_1 - G_2) = -i_{D2}. \tag{10}$$

The discussion of the switching mechanism at the application of a pulse is easier after one considers what happens as the voltage $e_p$ is slowly changed. If $e_p$ is slowly increased, $i_p$ simultaneously increases and this current divides equally between diodes $D_1$ and $D_2$. The admittance seen by the source $e_p$ is $2(G_1+G_2)$. Thus one sees that total current taken by diode $D_1$ increases while that taken by $D_2$ decreases. This situation persists as $e_p$ is continuously raised until diode $D_2$ is carrying no current, whereupon it opens, severing the connection between $B$ and $C$. Further increases in $e_p$ beyond this point result in a *decrease* of $i_p$, since the admittance seen at $B$ with $D_2$ open is $-(G_1^2/G_2)+G_2$. If one increases $e_p$ further, the current $i_p$ goes to zero and diode $D_1$ opens removing the pulse source from the circuit. At that point the two collectors carry equal currents. The circuit is in an unstable condition at this point. In the normal switching case one desires the unstable transient which ensues to cause $D_2$ to break down. However, for this "slow-motion" case at the severing of the switching source $e_p$, both transistors carry the same current and the direction of the subsequent switching transient is uncertain. The presence of appropriate condensers $C_1$ modifies the above description in the case of faster changing switching voltage, making it possible for a switching pulse to leave higher current in the transistor which initially carried the smaller current.

It is simple to trace the sequence of events which occur when with $D_1$ broken down a step of voltage is applied at $B$ of Fig. 6. The condensers $C_1$ begin to charge. $D_2$ opens as the current impulse brings its total current to zero. Thus $i_{b2}$ exhibits an impulse as does $-i_{e2}$ charging the right condenser. The impulse of emitter current causes $i_{c2}$ to jump to a larger value. Since $G_2$ is a small conductance, the principal part of $i_{c2}$ is drawn from the base of the left transistor. The result is that $i_{e1}$ exhibits a positive increment and $i_{c1}$ develops a negative increment which decreases the current in the diode $D_1$ ultimately bringing it to zero. At this point the circuit is free of the switching source and executes its own natural transient as described earlier.

Under the assumption that $G_2$ approaches zero, one finds the transients sketched in Fig. 7 for a range of different circuit parameters and switching pulses. The transients shown there are evaluated from the beginning of the switching pulse for a circuit representation which does not change during the pulse. In the actual circuit the switching transient is always terminated by the breakdown of one or both of the diodes, $D_1$ and $D_2$, whereupon the circuit is no longer unstable and a new set of decaying transients occurs. The description at the bottom of the columns in Fig. 7 indicates when the switching transient terminates. If the variables are considered in the order presented, the relationships given in

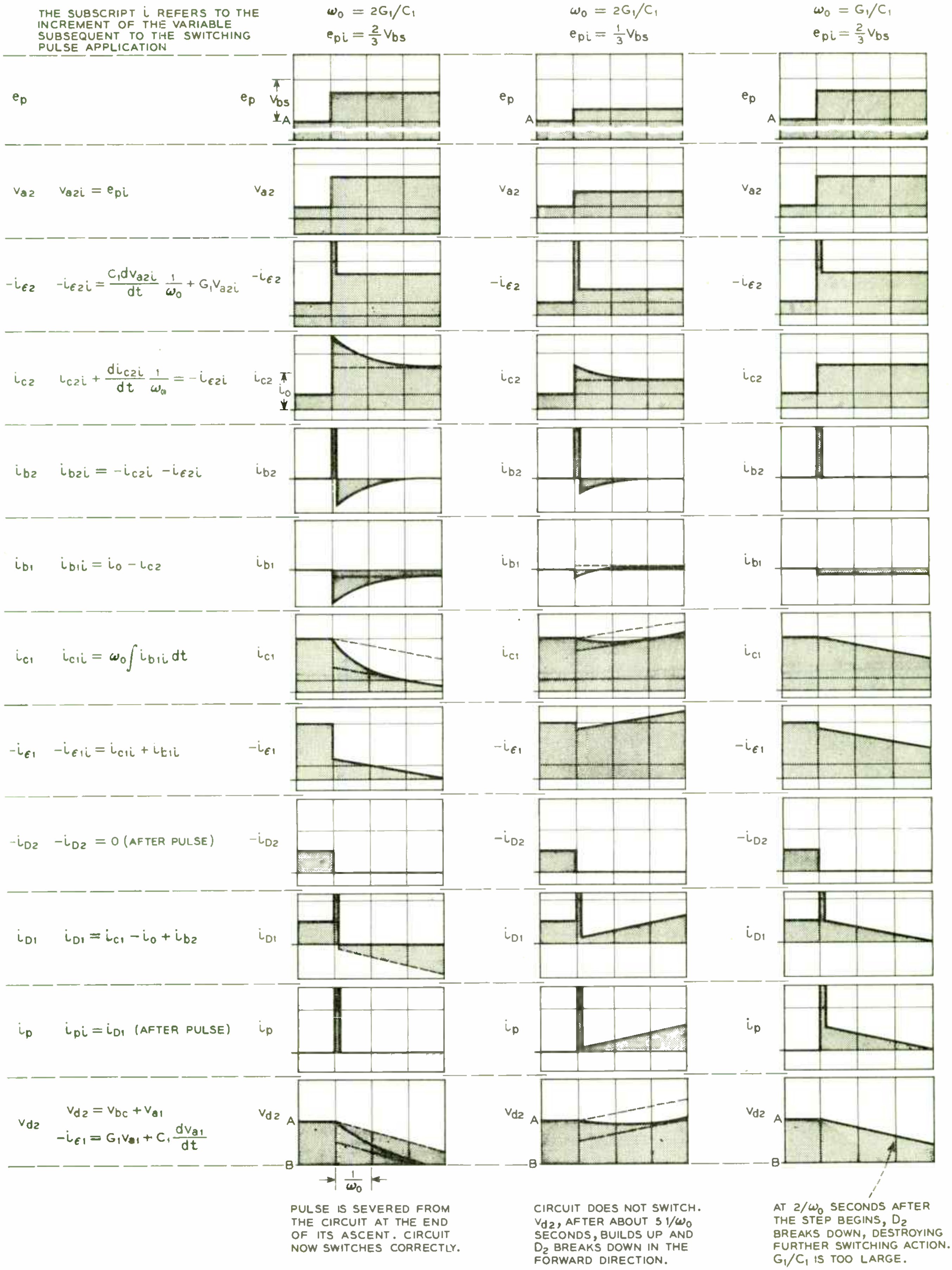


Fig. 7—Switching pulse shapes.

Fig. 7 are easily verified. From a study of Fig. 7 and consideration of other similar plots one can draw several interesting conclusions.

1. The switching pulse must supply a sufficient impulse to the low-conducting transistor's base to increase its collector current to a value which exceeds that in the companion transistor. A unit impulse of current, a coulomb of charge, applied to a transistor base causes a step in collector current of $\omega_0$ amperes. An increase in the size of the emitter condenser $C_1$ makes it possible to supply this impulse with lower values of switching voltage and less energy. However, increasing $C_1$ while one maintains $G_1$ fixed increases the time constant of the emitter branch and results in slower decay of some transient components. Transistors with high values of alpha-cut-off frequency $\omega_0$ are inherently easier to switch than poorer units. A simple but very approximate relationship for the minimum charge required for switching is

$$q = \frac{V_{bs}G_1}{\omega_0}. \tag{11}$$

From (9) one recognizes that this is the amount of charge which if passed through the low-conducting transistor's base will bring its collector current to the value possessed by the high-conducting transistor.

2. The source of the switching pulse may be severed from the counter either by its own high impedance after its impulse has been delivered or it may be severed by the opening of the second anti-saturation diode. The self-severing is enhanced by values of $C_1/G_1$ which are several times larger than $1/\omega_0$ or by values of switching pulse greater than $V_{bs}/2$ which gives a linear decrease in $i_p$ with time.
3. Switching failures can result from two kinds of improper pulsing. Too large pulses, greater than $V_{bs}$, for example, may cause both $D_1$ and $D_2$ to break down and after the pulse is over the counter becomes uncertain as to which way to switch. A similar kind of difficulty is experienced with smaller pulses if during the pulse duration $v_{d2}$ (see Fig. 7) decreases far enough to induce breakdown of $D_2$. The second type of failure results from an unsufficient pulse which after initially opening $D_2$ (Fig. 7) fails to get the second diode open and permits $D_2$ to begin conducting again in the forward direction.

## Effects of Temperature Changes on Circuit of Fig. 4

An interesting and useful property of the binary counter shown in Fig. 4 is that the stable voltages with respect to ground of points $c_1$, $\epsilon_1$, $c_2$ and $\epsilon_2$ are essentially independent of the transistor parameters and saturation current $I_{c0}$, being dependent upon the resistors and breakdown diodes and the applied voltage. This would be exactly so if there were no voltage drops between base and emitter of a conducting transistor and if the breakdown diodes have zero incremental impedances, but it is nearly so in any practical case since the emitter-base impedance of a transistor and the conduction impedance of a breakdown diode are small fractions of the other impedances in the circuit. The reason for the independence is easily explained. An examination of Fig. 6 in which the emitter-base voltages are neglected and the voltages of the diodes $D_1$ and $D_2$ are taken as $V_{bs}$ and zero reveals that the voltages across the resistors are determined without any further characterization of the transistors. The transistors simply determine how the current divides between them and the cross-coupling diodes without influencing conditions in the resistors at all.

## Experimental Results

A number of different two-transistor pulse circuits have been built and operated. Oscillograms of interesting voltages of a number of them are presented in the remaining figures. In particular the circuits described include a binary counter operated by a blocking oscillator, a two-stage binary counter, the second stage being driven by the first and a monostable or astable two-transistor pulse circuit which is also nonsaturating.

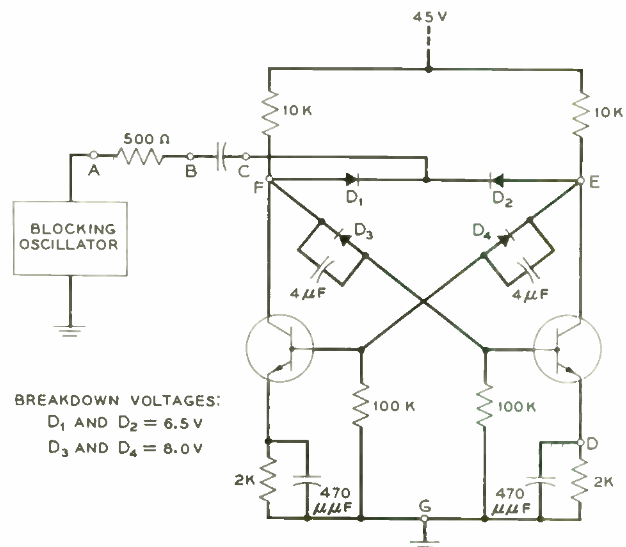


Fig. 8—Binary counter driven by blocking oscillator.

Fig. 8 gives the circuit arrangement of a binary counter driven by a blocking oscillator which operated at a maximum of 600 kilocycles using grown-junction triodes. Fig. 9 shows a number of oscillograms of the pertinent voltages when the counter is driven by a 200-kilocycle wave from the blocking oscillator. Figs. 9(a), 9(b) and 9(d) are taken with the same attenuator setting on the oscilloscope and with the sweep synchronized with the blocking oscillator is about 2 volts with the peak current from the blocking oscillator being about 3 ma as seen from Fig. 9(b). The switching charge

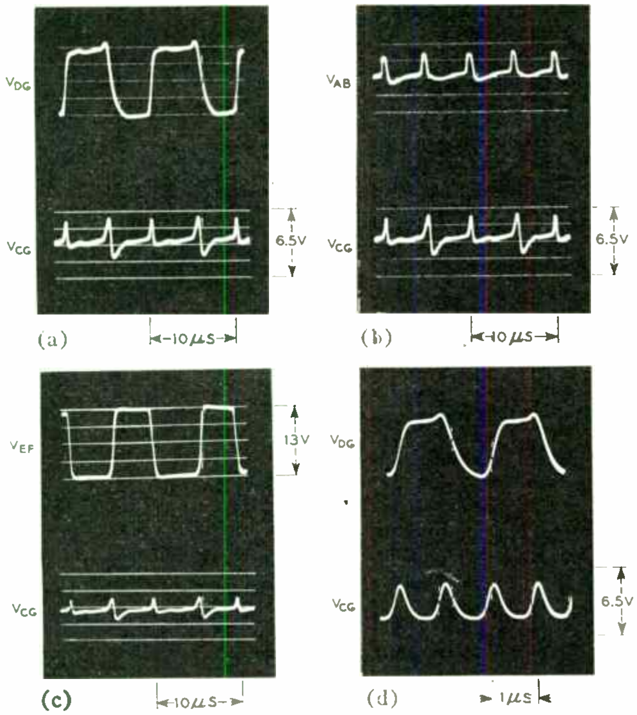

Fig. 9—Oscillograms for binary counter of Fig. 8.

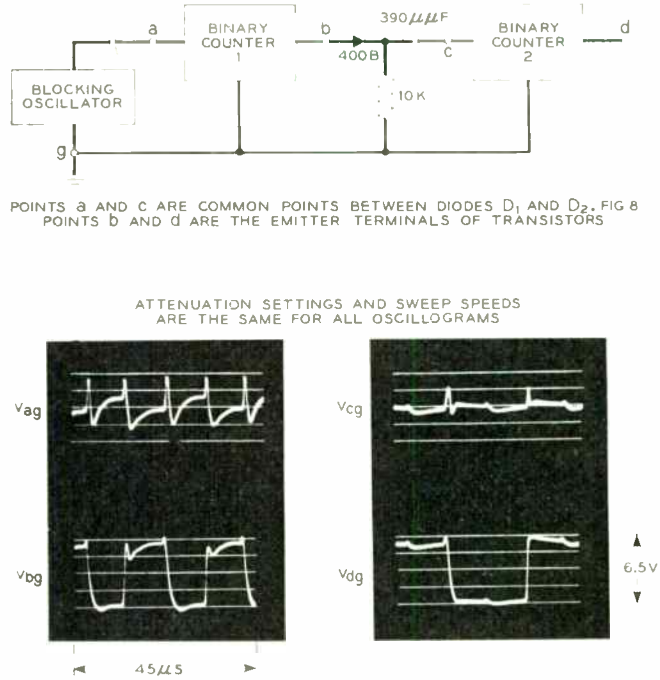

Fig. 10—Circuit arrangement and oscillograms for two-stage binary counter.

is very crudely estimated to be $1.5\times10^{-9}$ coulombs which compares with $3.2\times10^{-10}$ coulombs obtained from the formula of (11) if one uses the value of $10^7$ as the radian cut-off frequency of the grown-junction transistors. Fig. 9(d) applies to a binary counter in which grown-junction transistors are replaced by *n-p-n* alloy-junction transistors and the blocking oscillator frequency is increased to 1 megacycle. Fig. 9(c) has a 2 to 1 change in oscilloscope attenuator. It is necessary to observe also that the 470 μμf condensers shown in Fig. 8 were replaced by 150 μμf condensers. With grown-junction triodes the counter will not operate with 150 μμf condensers and with the alloy-junction transistors the counter cannot be driven at 1 mc with 470 μμf condensers. Naturally it operates with less driving pulse amplitude at lower frequencies with the larger condensers than is possible with the 150 μμf condensers.

Fig. 10 shows the circuit arrangement and oscillograms for a two-stage binary counter of the type shown in Fig. 8. The counters are coupled by a diode network which prevents operation of the second binary counter on down swings of the voltage at *B*.

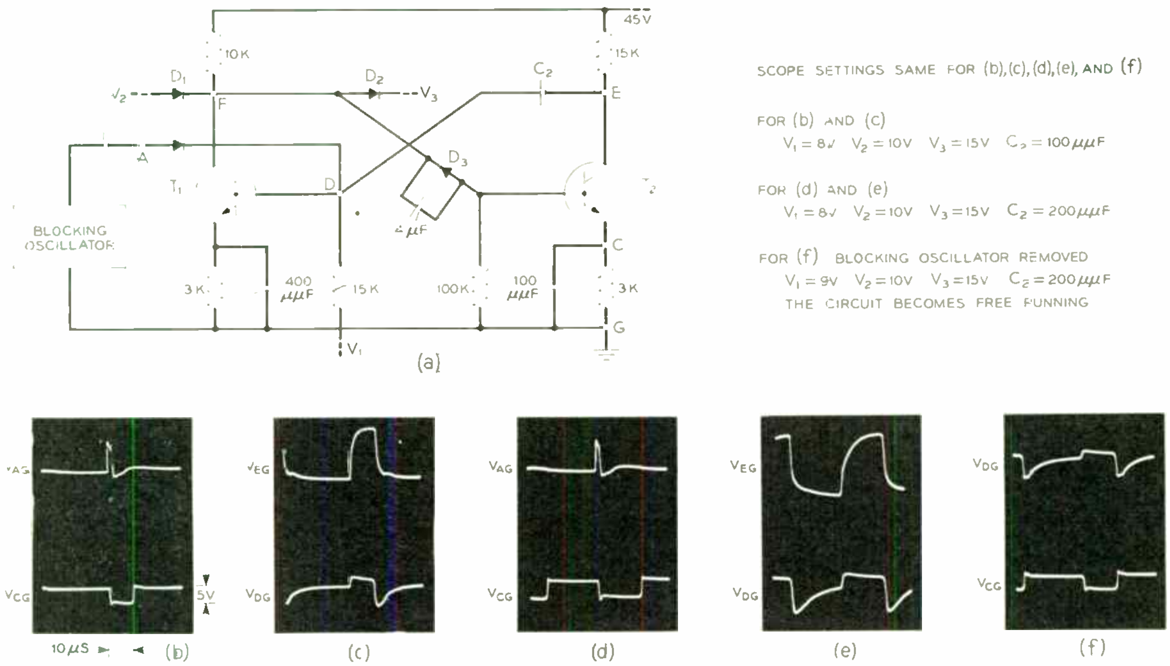

Fig. 11—Circuit diagram and oscillograms for nonsaturating astable or monostable circuits.

Most of the foregoing material has been applied directly to binary counters. The same general ideas, nonsaturation, switching, speed, etc., apply nearly unaltered in the case of monostable and astable circuits of similar configuration. Fig. 11 (preceding page) shows a circuit which can be made monostable or astable by proper selection of the voltage $V_1$. This circuit, which is not symmetrical, differs from the binary counter in two principal respects. First, one of the cross couplings is capacitive, and this accounts for the maximum of one stable state. Second, saturation is prevented by two diodes, both point diodes in the example illustrated, connected to one of the collector terminals. One, $D_1$, prevents the collector voltage on the adjacent transistor from dropping to zero. The second, $D_2$, prevents a sufficient rise of the collector voltage of the adjacent transistor to drive the collector voltage of the companion transistor off. Both of these diodes could have been breakdown diodes of the appropriate breakdown voltage. Both would have been reversed in orientation, $D_1$ being connected to $B^+$, $D_2$ being connected to ground. The third diode, $D_3$, is used simply as a coupling element, which, in combination with $D_1$ and $D_2$, assures that the output voltage wave at $c$ has tops and bottoms independent of the transistors. Since at the time of switching, the equivalent representation of this circuit is essentially the same as that of the binary counter, one expects similar switching times and experimental results verify this.

When $V_1$ is set to zero or to a value close to zero, diode $D_2$ will be broken down and the circuit is in a stable state with

$$V_{CG} = V_3 - V_{D3} \text{ and}$$
$$V_{D3} = 8.5 \text{ v for diodes } D_3 \text{ used.} \tag{13}$$

A positive pulse at $A$ opens $D_2$ and the circuit switches $T_1$ on more heavily, switching being terminated by breakdown of $D_1$. The condenser $C_2$ cannot maintain the voltage of point $D$ indefinitely and when the voltage at $D$ falls to a value which permits $D_1$ to open, the circuit switches to its original stable state.

When the voltage $V_1$ is raised sufficiently that $D_2$ is opened the circuit becomes astable with an output wave of amplitude $V_3 - V_2$ and with pulse length being principally determined by $C_2$ and the connected resistors.

# A Two-Emitter Transistor with a High Adjustable Alpha*

R. F. RUTZ†

*Summary*—**The current amplification, $\alpha$, of a point contact transistor can be increased to values in excess of 20 by the addition of a third point contact which is biased so as to act as an emitter. The amount of increase in $\alpha$ can be adjusted by varying the second emitter current. A qualitative explanation of the $\alpha$ enhancement is discussed which involves an internal positive feedback action that varies the hole transport factor, $\beta$, associated with the second emitter as the first emitter current is changed. The effect of varying the second emitter-to-collector spacing is discussed and experimental results are given.**

THE CURRENT amplification, $\alpha$, of a point contact transistor can be materially increased and controlled by adding a third whisker placed far from the collector and biased so as to act as an emitter. This paper will present a discussion of the mechanism whereby this $\alpha$ enhancement is achieved and describe the characteristics of some experimental two-emitter transistors. Occasionally, in the literature, mention has been made of other types of $\alpha$ variations brought about by the use of additional electrodes or whiskers.[1]

The operation of a conventional point contact transistor has been discussed by Shockley.[2] This transistor consists of a small block of $N$-type germanium and three electrodes attached to it. These electrodes are an ohmic connection known as the base and two-point contacts, placed a few thousandths of an inch apart, known as the emitter and collector. In normal operation, the emitter is biased positively and injects holes into the germanium, and the collector is biased negatively and collects these holes. The collection mechanism is such

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[1] H. J. Reich, P. M. Schultheiss, J. G. Skalnik, T. Flynn, and J. E. Gibson, "Effect of auxiliary current on transistor operation," *Jour. Appl. Phys.*, vol. 22, p. 682; May, 1951.
J. G. Skalnik, H. J. Reich, J. E. Gibson, and T. Flynn, "Auxiliary current alters transistors characteristics," *Electronics*, vol. 24, p. 142; September, 1951.
T. R. Scott, "Crystal triodes," *Proc. IEE*, part III, vol. 98, pp. 169–177; May, 1951.

[2] W. Shockley, "Electrons and Holes in Semiconductors," D. Van Nostrand Company, Inc., New York, N. Y., pp. 101 ff.; 1950.

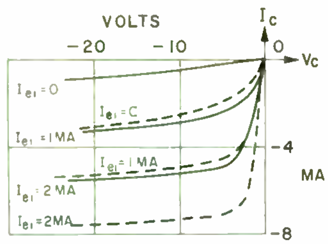


Fig. 1—Collector $V$-$I$ characteristics for a typical transistor with a second emitter spaced close to the collector.

that holes arriving in the neighborhood of the collector may allow additional electrons to flow from the collector to the base. Thus, a given change in the emitter current may cause a greater change in collector current. The current amplification factor $\alpha$ is defined by:

$$\alpha = -\left(\frac{\partial I_c}{\partial I_e}\right)_{V_c=\text{const.}}$$

where the subscripts $e$ and $c$ refer to the emitter and collector respectively and $V_c$ is the collector to base voltage. Since only the holes injected by the emitter and arriving at the collector are effective in changing the collector current, it is convenient to express $\alpha$ as the product of three factors:

$$\alpha = \alpha^*\beta\gamma$$

where

$$\alpha^* = \frac{\partial I_c}{\partial I_{cp}}, \qquad \beta = -\frac{\partial I_{cp}}{\partial I_{ep}}, \qquad \gamma = \frac{\partial I_{ep}}{\partial I_e},$$

and the subscript $p$ means that part of the current carried by holes. Here $\gamma$ is the hole injection efficiency of the emitter, $\beta$ is the transport efficiency, and $\alpha^*$ is the intrinsic $\alpha$ of the collector. The emitter of a point contact transistor is normally placed so close to the collector that $\beta$ is virtually unity. The $\gamma$ will not in general exceed unity and may be less. The intrinsic $\alpha$ may be considerably greater than unity.

With this background, we are in a position to consider the effect on the $\alpha$ of a transistor of adding a second emitter. It has been found that the effect depends upon the spacing of the second emitter contact from the collector contact. A closely-spaced second emitter also will have a $\beta$ factor of nearly unity and there will be little direct interaction between the emitters. This has been discussed by Haegele.[3] The collector hole current will be the sum of the two-emitter hole currents. The average $\alpha$ of the transistor will not be much affected by current in the second emitter. The $\alpha$ referred to here and subsequently in this paper is defined as:

$$\alpha = -\left(\frac{\partial I_c}{\partial I_{e1}}\right)_{V_c,\ I_{e2}=\text{const.}}$$

[3] R. W. Haegele, "A crystal tetrode mixer," *Sylvania Tech.*, vol. 2, pp. 2–4; October, 1949.

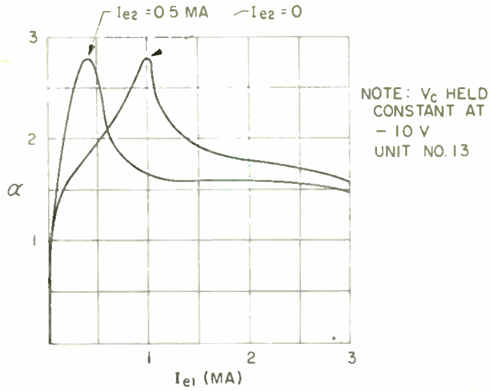


Fig. 2—$\alpha$ vs $I_{e1}$ for a transistor with a second emitter spaced close to the collector.

where the subscripts $e1$ and $e2$ refer to the first and second or added emitter respectively.

Fig. 1 shows collector $V$-$I$ characteristics of a typical transistor with a second emitter close to the collector. The solid lines are for the case where $I_{e2}=0$ and correspond to the conventional two-whisker transistor characteristics. The dashed lines are for the case where $I_{e2}=1.0$ ma. It can be seen that outside of the region of small collector voltage, a displacement of the lines of constant $I_{e1}$ takes place, but there is no significant change in separation of adjacent lines. Hence, the average $\alpha$ of the original transistor has not been altered appreciably. It has been observed in cases where the curve of $\alpha$ vs $I_{e1}$ has a peak, that this peak may be shifted and somewhat modified when a constant current is applied through the second emitter. Fig. 2 shows an example of this. Here $\alpha$ is shown as a function of $I_{e1}$ for two different values of $I_{e2}$, namely, $I_{e2}=0$ and $I_{e2}=0.5$ ma. It will be noticed that the peak in the $I_{e2}=0$ curve is displaced to the left by approximately 0.5 ma in the $I_{e2}=0.5$ ma curve. This may be explained on the basis that the peak in $\alpha$ is due to a peak in the intrinsic $\alpha$ of the collector which occurs at some definite value of collector hole current. When some of the hole current is supplied by the second emitter the peak occurs at a lower value of first emitter current. If the second emitter hole current is increased beyond the value at which the intrinsic $\alpha$ has its maximum, then the peak will no longer appear in $\alpha$.

If a second emitter is placed far from the collector so that in the absence of first emitter current its transport efficiency is small, a large enhancement of the $\alpha$ of the transistor is possible without greatly changing the other transistor characteristics. Fig. 3 shows a typical example of this effect in the form of a set of collector $V$-$I$ characteristics for an experimental transistor of this type for the two cases of $I_{e2}=0$ (conventional two-whisker transistor) and $I_{e2}=8$ ma. It is evident that the main effect, outside of the region of small collector voltage, is an increased separation of the lines of constant $I_{e1}$, which means that the average $\alpha$ of the original transistor has been noticeably increased. The effect is due to an increase in the transport efficiency of the second emitter as a result of an increase in current through the first emitter.

The following qualitative explanation of how this may be accomplished has been suggested by R. W. Landauer.[4] Consider Fig. 4 which is a schematic representation of the hole and electron flow in the transistor with a second emitter placed a relatively large distance from the collector. The holes injected by the second emitter

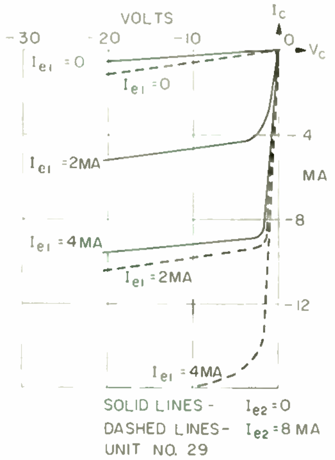


Fig. 3—Collector $V$-$I$ characteristics for a transistor with a second emitter spaced far from the collector.

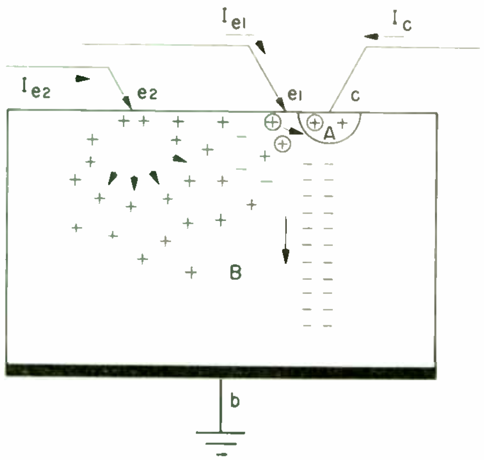


Fig. 4—Hole injection by second emitter spaced far from the collector.

are indicated symbolically by + and those injected by the first emitter indicated by ⊕. Electrons are indicated by −. The region near the collector where the amplifying mechanism is concentrated is indicated by the region $A$ and the rest of the germanium block is labelled region $B$. Let a constant current $I_{e2}$ be flowing into the second emitter and let the collector voltage be held at some constant negative value. Now consider what happens when the first emitter current is turned on. If the intrinsic $\alpha$ of the collector is sufficiently high, then the increase in collector current due to holes from the first emitter arriving in region $A$ must be accompanied by an increased electric field in region $B$. Holes from the fringe of the reservoir of holes created by $I_{e2}$ will be drawn to the collector by this increased field. The collecting of these holes in region $A$ liberates even more electrons to flow into region $B$. This further increases the electric field there so that still more holes are drawn from the reservoir. In this way, we have a positive feedback mechanism which, for a given $I_{e1}$, will make the current gain greater than would have been present without the reservoir of additional holes. The depletion of the reservoir by the feedback action is equivalent to an increase in the $\beta$ factor of the second emitter.

It is observed that the back resistance of the transistor is lowered by a current flowing into the second emitter. This is because a small number of holes will arrive at the collector from the second emitter even when there is no current through the first emitter, and the number arriving will depend on the collector voltage. This means that $\beta$ factor associated with the second emitter depends slightly upon the collector voltage. A typical example of the change in back resistance due to second emitter current is shown by the change in slope of the $I_{e1}=0$ line in Fig. 3.

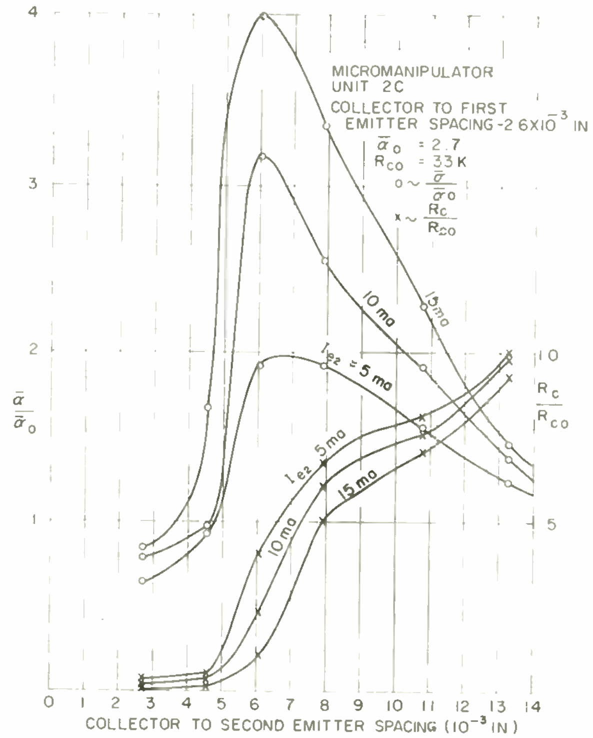


Fig. 5—Average $\alpha$ and back resistance as a function of second emitter spacing for different values of second emitter current.

The curves in Fig. 5 show how the increase in average $\alpha$ varies with the second emitter to collector spacing in a typical transistor for three different values of second emitter current. Also shown in Fig. 5 is the decrease in collector back resistance for the same transistor. The average alpha, $\bar{\alpha}$, is the alpha associated with the collector and first emitter averaged over the interval

[4] Private communication.

of first emitter current from 0 to 2 ma, with the collector voltage held constant at $-10$v and the second emitter current held at some constant value. The back resistance, $R_c$, is the ratio of the collector voltage to collector current when $V_c=-20$v and $I_{e1}=0$ and the second emitter current is held constant. For the special cases where $I_{e2}=0$ the average $\alpha$ and back resistance are designated as $\bar{\alpha}_0$ and $R_{c0}$ respectively.

It is apparent that for second emitter to collector spacings of less than four-thousandths of an inch, $\bar{\alpha}$ actually decreases. This is because the second emitter gives sufficient collector hole current so that $\alpha^*$ has passed its peak. Except for this close spacing, $\bar{\alpha}$ increases as second emitter current increases. At the same time back resistance decreases. In general, a compromise must be made between these two effects. As might be expected, for a given second emitter current, the $\alpha$ enhancement falls off for large spacings.

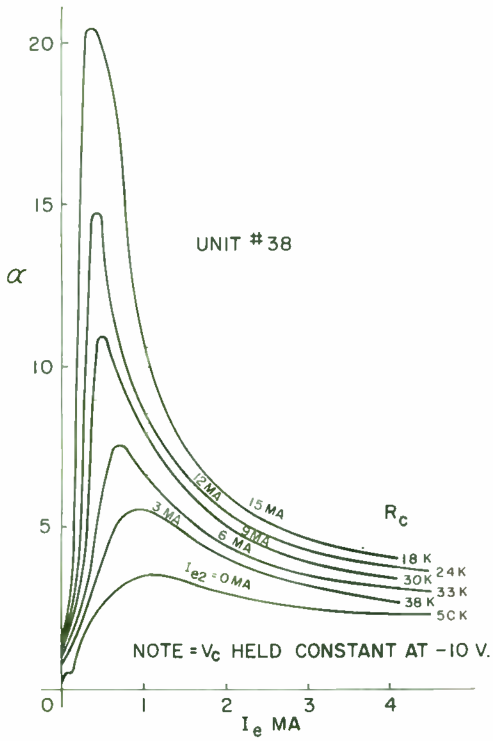


Fig. 6—$\alpha$ vs $I_e$ for various values of second emitter current for a transistor with a second emitter spaced far from the collector.

Fig. 6 shows a detailed picture of the relationship between $\alpha$ and $I_{e1}$ for different values of $I_{e2}$ for a single transistor. The back resistance, $R_c$, associated with each value of the $I_{e2}$ is also shown. In this particular transistor, as has been found to be the case generally, the $\alpha$ enhancement due to $I_{e2}$ is greatest in the region of low first emitter current and falls off for high values of $I_{e1}$. This clearly must happen since the $\beta$ of the second emitter cannot increase above unity.

To show what can be realized in practice, the characteristics of four encapsulated transistors with widely spaced second emitters are given in Table I. In these transistors back resistance decreases approximately fifty per cent and average $\alpha$ about doubles when the second emitter current is raised from zero to 5ma.

TABLE I

Average $\alpha$ and Collector Back Resistance for Four Transistors with Widely-Spaced Second Emitters

| Unit No. | Back Resistance (ohms) | | Average $\alpha$ | |
|---|---|---|---|---|
| | $R_{c0}(I_{e2}=0)$ | $R_c(I_{e2}=5$ ma$)$ | $\alpha_{avg}(I_{e2}=0)$ | $\alpha_{avg}(I_{e2}=5$ ma$)$ |
| A | 20,000 | 10,000 | 2.5 | 5.0 |
| B | 21,000 | 10,000 | 1.8 | 3.7 |
| C | 29,000 | 14,000 | 3.0 | 5.9 |
| D | 50,000 | 35,000 | 2.6 | 4.6 |

In general, it has been found that the greatest $\alpha$ enhancement occurs in transistors which have an initially high value of $\alpha$. This result might be expected since a large $\alpha$ implies a large intrinsic $\alpha$ and hence large sweeping fields. The grestest $\alpha$ enhancement is obtained when the second emitter is so placed that the first emitter lies on a line between it and the collector. It has also been found that best results are obtained with short lifetime germanium. This means that the reservoir of holes created by the second emitter can be located relatively close to the collector and hence easily be affected by the sweeping fields. Finally, it has been found desirable to make the base connection farther from the second emitter than is the collector. A nearer base would presumably tend to drain holes from the reservoir.

In circuit applications, the high $\alpha$ two-emitter transistor has the advantage of permitting higher current gain than is normally obtainable from conventional point contact transistors. Also, it is useful for applications requiring the control or modulation of the gain of an amplifier. Measurements on one transistor in a circuit with a collector load resistor of a few hundred ohms indicate that the rise time of a rectangular current pulse amplified by the transistor is substantially the same with moderate second emitter currents as it is with zero second emitter current (i.e., as it is for the original transistor comprised of the first emitter and collector alone). The rise time of an output pulse at the collector when a rectangular positive current pulse switches the second emitter current from zero to 5 ma, for a steady first emitter current of a few milliamperes, has been found to be less than a half-microsecond in an experiment on one transistor. Thus, the modulating properties of the second emitter are apparently not limited to excessively low frequencies.

## Acknowledgment

The author wishes to thank the members of the IBM Research Laboratory and especially L. P. Hunter and J. A. Swanson for their many clarifying discussions in connection with the development of the high-$\alpha$ two-emitter transistor. Also he wishes to acknowledge the assistance of G. A. Beutel and P. Fiore who fabricated the transistors used in these experiments.

# Internal Feedback and Neutralization of Transistor Amplifiers*

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*Summary*—**Transistors are nonunilateral amplifying devices. The most important effects of internal feedback are reflected immittances and potential instability of amplifiers in certain frequency ranges. These phenomena are undesirable in many applications.**

**Considering the various matrix representations of a two-terminal pair one can show that by connecting appropriate networks in a suitable manner to the active nonunilateral element, the internal feedback of the latter can be removed (neutralized).**

**Several neutralized transistor arrangements have been investigated experimentally. The circuits are based on an analysis of the properties of transistor feedback parameters at "higher" frequencies. Neutralization throughout relatively wide bands of frequencies can be achieved using simple feedback arrangements. The maximum available power gain of transistor amplifiers is only moderately affected by neutralization.**

## Introduction

A TWO-TERMINAL pair network is unilateral if an excitation applied to one of its terminal pairs produces a response at the second terminal pair, whereas an excitation applied to the second terminal pair does not result in a response at the first terminal pair, or vice versa. Networks with bidirectional transmission between terminal pairs are nonunilateral; they may be bilateral (if they obey the theorem of reciprocity) or nonbilateral.[1]

It is well known that vacuum tubes operated at low frequencies can be considered as unilateral devices. Transistors, however, are nonunilateral: the three transistor configurations (common base, emitter and collector) exhibit bidirectional transmission between input and output terminals pairs. Although, in general, the "backward" transmission is considerably lower than the "forward" transmission, its effects on the circuit properties of the transistor cannot be neglected. The major effects of the backward transmission (or "internal feedback") on circuit behavior are:

1. The input and output driving point immittances are functions of the load and source immittances respectively.
2. Internal feedback may lead to instability of amplifiers and circuits may oscillate even in the absence of an external feedback loop.

These phenomena are undesirable in numerous transistor applications and often represent major difficulties to the circuit designer and user. In other cases, these properties may be of no consequence or are even desirable.

The problem of internal feedback has been studied by Mason,[2] who has shown that unilateralization can be achieved by lossless reciprocal coupling. The approach of this paper is somewhat different. The discussed methods of neutralization involve both resistive and reactive elements. This may result in a sacrifice of power gain, but often "simplifies" the neutralizing network and makes it easier to design an amplifier neutralized throughout a relatively wide band of frequencies. The discussion is mainly concerned with transistor amplifiers; the principles can, however, be applied to other active devices.

## Reflected Immittances

The behavior of any transistor configuration, considered as an active linear two-terminal pair, can be described by one out of six possible sets of two linear equations. Using the "series-parallel" representation for the generalized transistor amplifier of Fig. 1 one can write:

$$E_1 = h_{11}I_1 + h_{12}E_2 = E_G - Z_G I_1 \qquad (1)$$
$$I_2 = h_{21}I_1 + h_{22}E_2 = - E_2 Y_L.$$

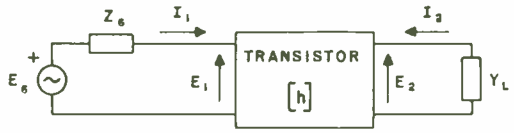


Fig. 1—Schematic representation of transistor amplifier.

The source impedance is $Z_G = R_G + jX_G$ and the load admittance is $Y_L = G_L + jB_L$. The parameters $h_{ij}$ are functions of the dc operating point, the signal frequency and are, of course, different for different transistor configurations. The existence of internal feedback manifests itself by $h_{12} \neq 0$. The input impedance $Z_i$ is a function of the load admittance:

$$Z_i = h_{11} - \frac{h_{12}h_{21}}{h_{22} + Y_L}. \qquad (2)$$

In a similar manner, the output admittance $Y_0$ is a function of the source impedance:

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[1] These definitions may be disagreed with, and the literature is not unanimous on the subject. However, these definitions seem to be compatible with E. A. Guillemin, "Introductory Circuit Theory," John Wiley & Sons, Inc., New York, N. Y., p. 148, 1953, and other textbooks.

[2] S. J. Mason, "Power Gain in Feedback Amplifiers," Res. Lab. Elec., M.I.T. Tech. Rep. No. 257; August 25, 1953

$$Y_0 = h_{22} - \frac{h_{12}h_{21}}{h_{11} + Z_G}. \tag{3}$$

Due to "reflected immittances" a transistor amplifier stage cannot be designed as an isolated unit. To obtain proper performance, the designer must take into account the effect of adjacent and often even that of more remote amplifier stages. The internal feedback of transistors makes it difficult to use them in various applications, e.g., in certain types of laboratory equipment.

The problem of reflected immittances is particularly serious in the case of high frequency amplifier stages, where transistor parameters and external circuit elements are complex quantities. Fig. 2 shows the variation of input and output impedances of a common emitter amplifier with tuned input and output as functions of frequency. The complicated nature of the reflected immittances results in a distortion of the bandpass characteristic of tuned amplifiers (nonsymmetrical bandpass).

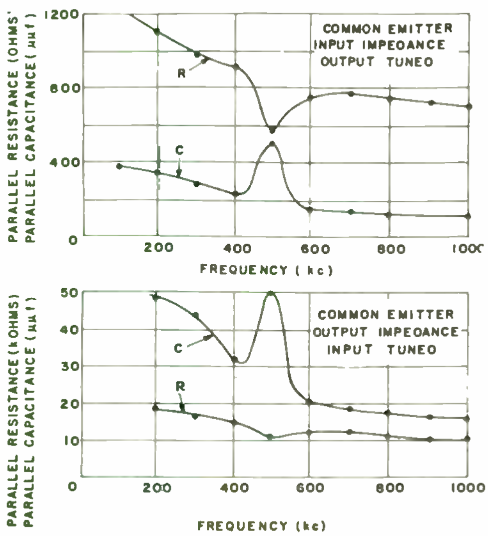


Fig. 2—Input and output impedances of the common emitter stage as functions of frequency with load and source tuned to 500 kc respectively.

Furthermore, it is not only difficult to design and compute the performance of a multistage tuned amplifier for desired bandshape and gain, but even the alignment of such an amplifier may be a laborious task. The design and alignment of multistage staggered tuned amplifiers present particularly difficult problems.

It can, therefore, be stated that reflected immittances are undesirable in many circuit applications and their elimination could solve numerous design problems

## Stability Considerations

The transducer gain of the transistor amplifier (defined as the power delivered to the load divided by the available power of the source) as calculated from (1), is:

$$G = \frac{4|h_{21}|^2 R_G G_L}{|(h_{11} + Z_G)(h_{22} + Y_L) - h_{12}h_{21}|^2}. \tag{4}$$

$h_{12}$ being different from zero, internal feedback may lead to instability even without an external feedback loop. A well-known example of such behavior in vacuum tube circuitry is the tuned plate-tuned grid oscillator, in which the grid-to-plate capacitance of the vacuum tube furnishes the internal feedback that is required for oscillation.

Analogous phenomena exist in transistor circuits. At higher frequencies the transistor parameters $h_{ij}$ are complex quantities:

$$h_{11} = h_{11}^{(R)} + jh_{11}^{(I)}$$
$$h_{22} = h_{22}^{(R)} + jh_{22}^{(I)}$$
$$h_{12}h_{21} = H = H_R + jH_I.$$

It can be shown[3] that, restricting $R_G$ and $G_L$ to positive values, instability may occur, provided that:

$$H_I^2 \geqq 4h_{11}^{(R)}h_{22}^{(R)}(h_{11}^{(R)}h_{22}^{(R)} - H_R). \tag{5}$$

In inequality (5) does not hold, the transistor configuration is unconditionally stable, whatever load and source immittances are connected to it. Condition (5) can also be written in the form

$$|H| + H_R \geqq 2h_{11}^{(R)}h_{22}^{(R)} \tag{5a}$$

An analysis of the three transistor configurations in the light of conditions (5) or (5a) shows that they exhibit potential instability throughout wide frequency ranges. This potential instability can be eliminated by neutralization of the internal feedback.

## Fundamental Aspects of Internal Feedback

A two-terminal pair network is characterized by two voltages, $E_1$ and $E_2$, and two currents, $I_1$ and $I_2$. Consequently, depending on which of these four quantities are considered as independent and dependent variables, the behavior of the network can be described by any one of six possible sets of two-linear equations.[4] Two of these sets (those separating input variables from output variables) are useful if one considers cascaded networks, whereas the four other sets are important from the point of view of internal feedback. Using the matrix notation, these are:

$$\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} = \begin{bmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{bmatrix} \times \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \tag{6}$$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \times \begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \tag{7}$$

$$\begin{bmatrix} E_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \times \begin{bmatrix} I_1 \\ E_2 \end{bmatrix} \tag{8}$$

$$\begin{bmatrix} I_1 \\ E_2 \end{bmatrix} = \begin{bmatrix} g_{11} & g_{12} \\ g_{21} & g_{22} \end{bmatrix} \times \begin{bmatrix} E_1 \\ I_2 \end{bmatrix}. \tag{9}$$

[3] A. P. Stern, "A stability criterion for active networks and its application to transistors," (to be published).

[4] E. A. Guillemin, "Communication Networks," John Wiley & Sons, Inc., New York, N. Y., vol. 2, p. 132 ff.; 1935.

If the matrix elements corresponding to one of the representations (6) to (9) are known, those pertaining to the other sets can be computed. Eqs. (6) to (9) can be written in a symbolic form:

$$\begin{bmatrix} D_1 \\ D_2 \end{bmatrix} = \begin{bmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{bmatrix} \times \begin{bmatrix} J_1 \\ J_2 \end{bmatrix}, \tag{10}$$

where $D_i$ and $J_i$ symbolize dependent and independent variables pertaining to terminal pair $i$ respectively. The matrix elements of (10) have the following significance:

$k_{11}$ is an input immitance,

$k_{12}$ is a backward transfer ratio or immittance indicative of internal feedback,

$k_{21}$ is a forward transfer ratio or immittance,

$k_{22}$ is an output immittance.

The network is unilateral if the feedback parameter $k_{12}=0$. Using a matrix conversion table one sees that if $k_{12}=0$ in one of the representations (6) to (9), it is zero in the others. For example, if $h_{12}=0$, than $z_{12}=y_{12}=g_{12}=0$. This, of course, is not surprising, because $h_{12}=0$ implies that an output excitation results in zero input response.

Although the feedback parameters in the representations (6) to (9) vanish simultaneously, it is useful to note that they *symbolize different aspects* of the internal feedback mechanism.

The transfer impedance $z_{12}$ represents series feedback proportional to the output current;

The transfer admittance $y_{12}$ represents parallel feedback proportional to the output voltage;

The voltage transfer ratio $h_{12}$ represents series feedback proportional to the output voltage; and

The current transfer ratio $g_{12}$ represents parallel feedback proportional to the output current.

## Theory of Neutralization

Let the transistor $T$ in a given configuration be represented by

$$\begin{bmatrix} D_1 \\ D_2 \end{bmatrix} = \begin{bmatrix} k_{11} & k_{12} \\ k_{21} & k_{22} \end{bmatrix} \times \begin{bmatrix} J_1 \\ J_2 \end{bmatrix}. \tag{10}$$

One can then consider a network $N$ ($N$ is not necessarily passive) described by:

$$\begin{bmatrix} D_1' \\ D_2' \end{bmatrix} = \begin{bmatrix} k_{11}' & k_{12}' \\ k_{21}' & k_{22}' \end{bmatrix} \times \begin{bmatrix} J_1' \\ J_2' \end{bmatrix}. \tag{11}$$

If $T$ and $N$ are connected to each other in a manner that forces $J_1'$ and $J_2'$ to be equal to $J_1$ and $J_2$ respectively, the composite network $C$ is defined by:

$$\begin{bmatrix} D_1'' \\ D_2'' \end{bmatrix} = \begin{bmatrix} k_{11}+k_{11}' & k_{12}+k_{12}' \\ k_{21}+k_{21}' & k_{22}+k_{22}' \end{bmatrix} \times \begin{bmatrix} J_1 \\ J_2 \end{bmatrix}. \tag{12}$$

If the feedback parameters of $T$ and $N$ satisfy the condition

$$k_{12}' = -k_{12} \tag{13}$$

the composite network $C$ is unilateral: one can say that the internal feedback of $T$ is neutralized.

These considerations can be applied to all representations (6) to (9) and result in the four basic neutralizing arrangements shown schematically in Fig. 3. The method of interconnection of $T$ and $N$ in these arrangements is consistent with the significance of the feedback parameters outlined at the end of the previous section.

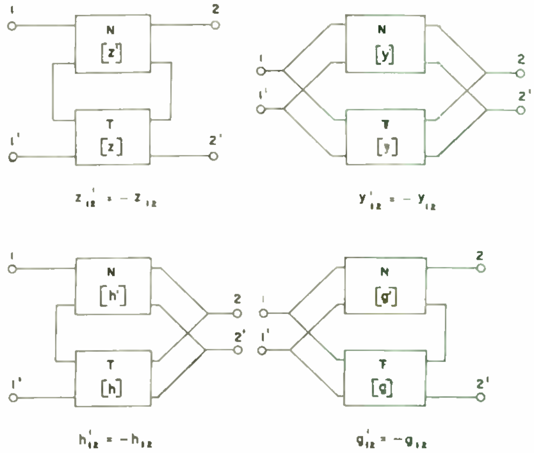


Fig. 3—Basic neutralizing arrangements.

If a practical neutralizing circuit is designed on the basis of the schematic arrangements of Fig. 3, caution must be exercised to interconnect $T$ and $N$ in a "permissible" manner[5] insuring the validity of (12). Problems of interconnection can be solved, as is well-known, using isolating transformers. Such a solution, however, is often undesirable because of circuit performance or cost. Consequently, the most practical circuits will be those which circumvent the difficulties of interconnection.

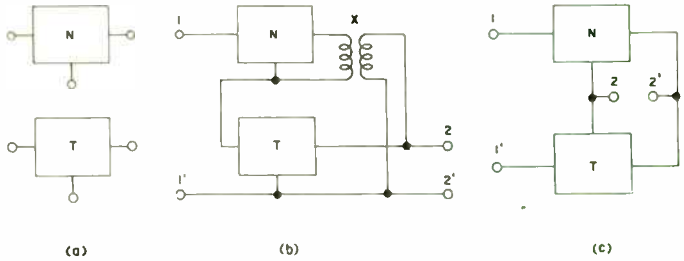


Fig. 4—$h$-type neutralization with and without transformer.

For example, one may consider the case of "$h$-type," ("series-parallel") neutralization. One transistor terminal is common to input and output, the transistor being in reality a three-terminal device. Neutralizing network $N$ may also be of the three-terminal type [Fig. 4(a)].

The two networks can be connected in series-parallel using an isolating transformer $X$ as shown in Fig. 4(b). If the transformer is close to "ideal" with unity transformation ratio and negligible phase shift, the composite network is unilateral for $h_{12}=-h_{12}'$; if the transformer provides 180-degree phase shift, neutralization

[5] Guillemin, *ibid.*

occurs for $h_{12}=h_{12}'$. However, the transformer is not necessary for interconnecting the networks. $T$ and $N$ can be directly connected in series-parallel, without transformer, in eight different ways. Only one of these connections is "permissible" and is shown in Fig. 4(c). For this arrangement neutralization occurs provided $h_{12}'=h_{12}$. This "bridge" circuit, is, of course, more desirable than its equivalent using a transformer.

Similar considerations apply to the "parallel-series" type interconnection of $T$ and $N$, for "$g$-type" neutralization. The use of a transformer can be avoided, as shown schematically in Fig. 5.

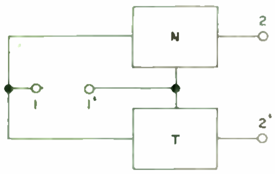


Fig. 5—$g$-type neutralization without transformer.

The composite networks of Fig. 4(c) and 5 are of the "four-terminal" type; no terminal is shared by input and output. Consequently, if such amplifying stages are cascaded, interstage transformers are usually required for reasons of common ground and power supply. If, on the other hand, a transformer is used *within* the neutralizing network, the composite network is of the three-terminal type and stages can be cascaded without an interstage transformer, if interstage impedance matching is considered unnecessary. It is, however, often preferable to use an interstage transformer rather than a transformer inside the neutralizing loop. The reason for this is that the network $N$ required for neutralizing a transistor throughout a reasonable frequency range is usually simple, provided no transformer is used (or for the case of an "ideal transformer"). With an actual transformer in the neutralizing loop the design of $N$ must be modified to account for the deviation of the transformer from an "ideal" one and this may complicate the design of $N$ considerably. This problem does not arise if the transformer is external to the neutralizing loop and is used solely for interstage coupling.

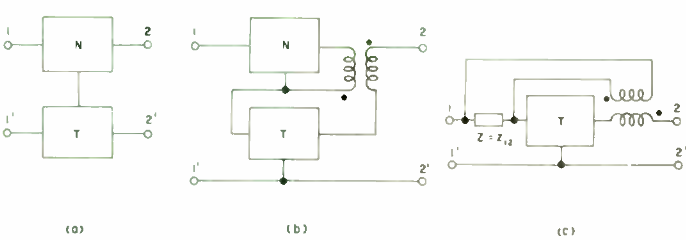


Fig. 6—$z$-type neutralization.

The problem of interconnecting $T$ and $N$ has somewhat different aspects in the case of $z$-type ("series") or $y$-type ("parallel") neutralization. The circuit of Fig. 6(a) is a "permissible" and transformerless series connection of $T$ and $N$ for $z$-type neutralization. The composite network is neutralized provided $z_{12}'=-z_{12}$. For different transistor configurations the real part of $z_{12}$ is usually positive and consequently $z_{12}'$ must have a negative real part of prescribed magnitude to match $z_{12}$. This can be realized easily for one given frequency, but it is generally difficult to obtain the desired $z_{12}'$ throughout a band of frequencies without using relatively complicated neutralizing networks consisting of four or more elements. The number of elements, however, should be kept minimum to avoid difficulties of alignment, and consequently, the transformerless circuit can be neutralized easily only in the neighborhood of one frequency. This may be satisfactory in certain applications, but if neutralization throughout a wider band is desired the circuit of Fig. 6(b) [or its practical version of Fig. 6(c)], using a transformer, would be preferable. The transformerless circuit may, however, be adequate for certain narrowband tuned amplifiers.

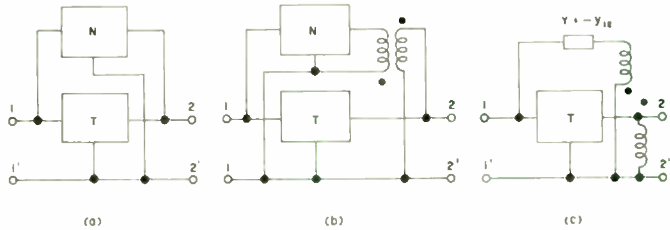


Fig. 7—$y$-type neutralization.

Similar considerations apply to $y$-type neutralization. In Fig. 7(a) no phase-inverting transformer is used and consequently neutralization will occur for $y_{12}'=-y_{12}$. In general, for transistors, $y_{12}$ has a negative real part [due to (7) and the sign of the currents and voltages in Fig. 1] and consequently $y_{12}'$ is required to have a prescribed magnitude with a positive real part. Without a transformer this can be realized only for one frequency using relatively simple networks. In most tuned amplifiers an interstage transformer is used anyway for interstage matching and the interstage transformer can simultaneously be used to provide neutralization in conjunction with a feedback network, as shown in Fig. 7(b). A practical realization is shown in Fig. 7(c).

The discussion can be summarized by stating that while no transformer is necessary for $h$- and $g$-type neutralization (except for interstage coupling purposes), in the case of $z$- or $y$-type neutralization a transformer within feedback network may be desirable. For wide band neutralization, $h$ and $g$-methods are often superior.

## Discussion of the Feedback Parameters

### *Equivalent Circuits and Approximations*

The low frequency values of the transistor feedback parameters $h_{12}$, $g_{12}$, $y_{12}$ and $z_{12}$ can be easily determined using any of the well-known low frequency equivalent circuits of the transistor. This section is concerned mainly with the feedback parameters at "higher" frequencies.

The $T$-type equivalent circuit of the common emitter configuration [Fig. 8(a)] is derived from a simplified version of the well-known equivalent of the common base amplifier [Fig. 8(b)]. Using these representations the following approximations and assumptions are made, in order to simplify the expressions obtained:

1. The common base transistor is composed of a simplified "ideal transistor," derived from the diffusion equation, collector barrier capacitance "base spreading impedance" $Z_b'$ [Fig. 8(b)]. The "ideal transistor" is considered as having no internal feedback, feedback in the actual transistor being due to $Z_b'$. At higher frequencies, this is a justifiable simplification.

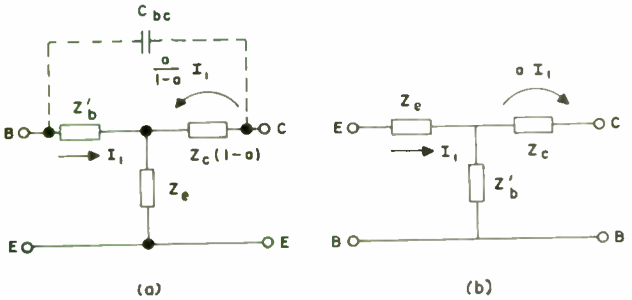


Fig. 8—Equivalents of common emitter and common base transistor circuits.

2. The frequency dependence of $a$ (short-circuit current amplification of the common base amplifier) is represented as

$$a = \frac{a_0}{1 + j\omega/\omega_a}, \tag{14}$$

where $\omega_a$ is the "$\alpha$-cut-off frequency" and $a_0$ the low frequency value of $a$.

3. The diffusion impedance of the emitter junction $Z_e$ is a function of frequency

$$Z_e = \frac{r_e}{1 + j\omega/\omega_a}, \tag{15}$$

where $r_e$ is the low frequency value of $Z_e$. This means that $Z_e$ is considered as the parallel connection of $r_e$ and of a capacitance

$$C_e = \frac{1}{r_e\omega_a}. \tag{16}$$

4. The collector impedance $Z_c$ is considered as the parallel connection of the collector resistance $r_c$ and the collector capacitance $C_c$($C_c$ being the sum of collector diffusion and barrier capacitances):

$$Z_c = \frac{r_c}{1 + j\omega C_c r_c} = \frac{r_c}{1 + j\omega/\omega_c}. \tag{17}$$

At frequencies considerably higher than $\omega_c/2\pi$:

$$Z_c \cong \frac{1}{j\omega C_c}. \tag{18}$$

5. The base spreading "resistance" $r_b'$ is considered complex ($Z_b'$) for grown junction $n$-$p$-$n$ transistors. This is due to the distributed nature of transistor parameters and $r_b'$, as shown by Pritchard and Coffey.[6]

The simplifying assumptions 1 to 4 are justified in the range of intermediate and higher frequencies and do represent the behavior of the transistor adequately up to a considerable fraction of $\omega_a$.

### *Common Emitter Configuration*

To explain the frequency dependence of the feedback parameters in common emitter configuration, it is necessary to consider the effect of a capacitance $C_{bc}$ connected between base and collector. The effect of this capacitance is particularly important in the case of grown junction $n$-$p$-$n$ transistors, as Pritchard has shown[7] (base overlap capacitance). In the case of other transistor types $C_{bc}$ is smaller, but, of course, still exists, due to transistor lead, socket and wiring capacitances.

Using the equivalent circuit of Fig. 8(a), the $h$-parameters of the common emitter amplifier are found to be approximately:

$$h_{11} \cong Z_b' + \frac{Z_e}{1 - a} \tag{19}$$

$$h_{12} \cong \frac{Z_e}{(1 - a)Z_c} + j\omega C_{bc}\left[Z_b' + \frac{Z_e}{1 - a}\right] = h_{12}' + h_{12}'' \tag{20}$$

$$h_{21} \cong \frac{a}{1 - a} \tag{21}$$

$$h_{22} \cong \frac{1}{(1 - a)Z_c} + \frac{j\omega C_{bc}}{1 - a}. \tag{22}$$

It must be remembered that, even if $\omega \gg \omega_c$, in view of the phase shift of $(1-a)$, $h_{22}$ is not purely capacitive (it is, in fact, almost purely resistive throughout a wide range of frequencies).

The feedback parameter $h_{12}$ consists of two components $h_{12}'$ and $h_{12}''$. According to (14) and (17)

$$h_{12}' = \frac{Z_e}{(1 - a)Z_c} = \frac{r_e/(1 - a_0)}{r_c} \frac{1 + j\omega/\omega_c}{1 + j\omega/(1 - a_0)\omega_a}. \tag{23}$$

$h_{12}'$ has the constant value $r_e/(1-a_0)r_c$ at low frequencies, increases between $\omega_c$ and $\omega_a$ $(1-a_0)$ and at higher frequencies is equal to $r_e\omega_a C_c$. The phase shift of $h_{12}'$ vanishes at low and high frequencies and has a peak between $\omega_c$ and $\omega_a$ $(1-a_0)$. The schematic frequency response of $h_{12}'$ is shown in Fig. 9 (opposite), whereas $h_{12}''$ increases with frequency and has a positive phase shift.

[6] R. L. Pritchard and W. N. Coffey, "Small-signal parameters of grown-junction transistors at high frequencies," 1954 IRE Convention Record, Part 3, "Electron Devices and Component Parts," vol. 2, pp. 89–98.

[7] R. L. Pritchard, "Effect of base-contact overlap and parasitic capacities on small-signal parameters of junction transistors," Proc. IRE, vol. 43, pp. 38–40; January, 1955.

The measured frequency dependence of $h_{12}$ of a grown-junction *n-p-n* transistor is shown in Fig. 10. At low frequencies $h_{12}'$ prevails. At higher frequencies $h_{12}''$ takes over and $h_{12}$ increases with frequency. The phase response corresponds to the amplitude response. The behavior of $h_{12}$ proves that the effect of $C_{bc}$ is considerable.

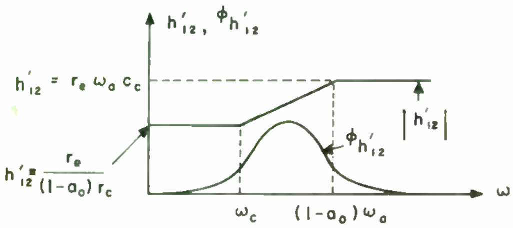


Fig. 9—Schematic frequency response of $h_{12}'$ (common emitter).

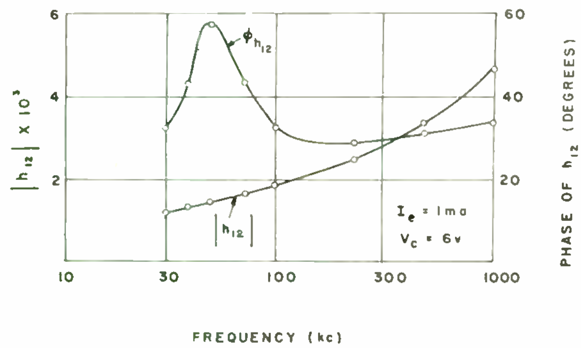


Fig. 10—Measured frequency response of $h_{12}$ for *n-p-n* grown junction transistor (common emitter).

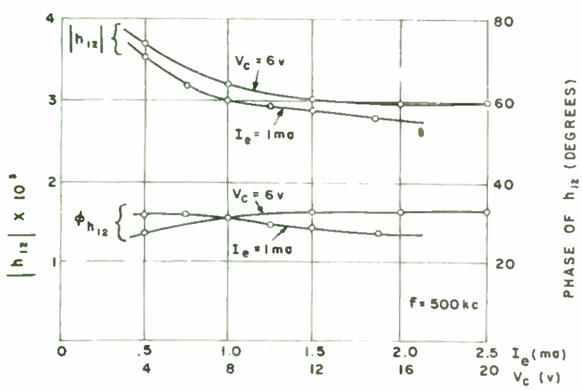


Fig. 11—$h_{12}$ as function of $I_e$ and $V_c$ (common emitter).

$h_{12}$ is function of the dc operating point. The dependence of $h_{12}$ on emitter current $I_e$ and collector voltage $V_c$ is shown in Fig. 11.

The other three feedback parameters can be calculated from (19) to (22). *Current feedback parameter is:*

$$(-g_{12}) = \frac{h_{12}}{\Delta^h}$$

$$\cong \frac{Z_e + j\omega C_{bc} Z_c (1-a)\left[Z_b' + \dfrac{Z_e}{1-a}\right]}{Z_b' + Z_e + j\omega C_{bc} Z_c (1-a)\left[Z_b' + \dfrac{Z_e}{1-a}\right]}. \quad (24)$$

If $Z_b'$ is purely resistive and $C_{bc}$ is very small, the phase shift of $(-g_{12})$ is negative. This, however, is not in agreement with measurements made on all rate grown *npn* and some *pnp* transistors. In the case of grown junction *npn* transistors the discrepancy is due mainly to the previously mentioned capacitive component of $Z_b'$ discussed by Pritchard and Coffey. They also measured the frequency response of $Z_b'$ and found that the magnitude and phase of $Z_b'$ change rather slowly with frequency.[8] Fig. 12 shows the measured variation of $g_{12}$ with frequency for a rate grown *n-p-n* transistor. Considering that in (24) usually $Z_e \ll Z_b'$, the response of $g_{12}$ seems to corroborate the measurements of Pritchard and Coffey. $g_{12}$ is function of $I_e$ and $V_c$ (Fig. 13).

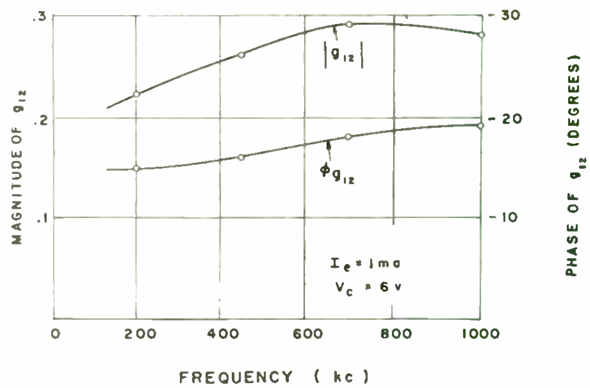


Fig. 12—Measured frequency response of $g_{12}$ for an *n-p-n* grown junction transistor (common emitter).

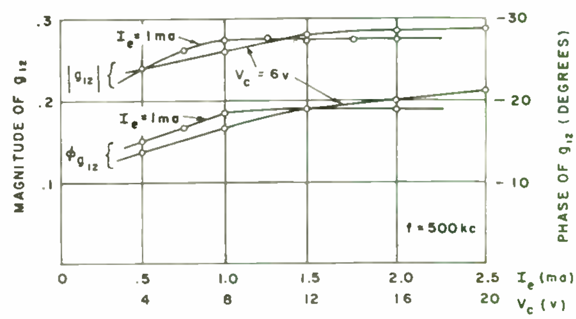


Fig. 13—$g_{12}$ as function of $I_e$ and $V_c$ (common emitter).

The *feedback impedance parameter* is:

$$z_{12} = \frac{h_{12}}{h_{22}} \cong Z_e + j\omega C_{bc} Z_c (1-a)\left[Z_b' + \frac{Z_e}{1-a}\right]$$
$$= z_{12}' + z_{12}''. \quad (25)$$

Here, as in the case of $h_{12}$, the effect of $C_{bc}$ is very important. If $C_{bc}$ could be neglected $z_{12}$ would be equal to $Z_e$ and would have a negative phase shift ($Z_e$ being capacitive). The portion $z_{12}''$ contributed by $C_{bc}$ is inductive and thus $z_{12}$ has a positive phase shift.

*Feedback admittance parameter* can be computed as:

$$(-y_{12}) = \frac{h_{12}}{h_{11}} \cong \frac{1}{Z_c} \frac{Z_e}{Z_b'(1-a) + Z_e} + j\omega C_{bc}. \quad (26)$$

[8] Pritchard and Coffey, *loc. cit.* (Fig. 8).

If $Z_b'$ were zero, $(-y_{12})$ would be purely capacitive at frequencies exceeding $\omega_c$. In actual transistors, however, the feedback admittance has a considerable conductive component. The phase of $y_{12}$ depends, of course, also on the phase of $Z_b'$.

The expressions derived for the feedback parameters can be simplified considerably and their significance visualized if one considers frequencies exceeding $(1-a_0)\omega_a/2\pi$, but smaller than $\omega_a/2\pi$. In this frequency range one may use the approximations:

$$a \cong a_0 \quad (27a)$$

$$Z_c \cong r_e \quad (27b)$$

$$(1-a) = (1-a_0)\frac{1 + j\omega/\omega_a(1-a_0)}{1 + j\omega/\omega_a} \cong j\frac{\omega}{\omega_a} \quad (27c)$$

$$1/Z_c \cong j\omega C_c \quad (27d)$$

$$Z_b' > \frac{Z_e}{1-a}. \quad (27e)$$

With these simplifications the feedback parameters can be written as:

$$h_{12} \cong r_e\omega_a C_c + j\omega C_{bc} Z_b' \quad (28)$$

$$(-g_{12}) \cong \frac{r_e + j\dfrac{\omega}{\omega_a}\dfrac{C_{bc}}{C_c}Z_b'}{Z_b' + r_c + j\dfrac{\omega}{\omega_a}\dfrac{C_{bc}}{C_c}Z_b'} \quad (29)$$

$$z_{12} \cong r_e + j\frac{\omega}{\omega_a}\frac{C_{bc}}{C_c}Z_b' \quad (30)$$

$$(-y_{12}) \cong \frac{r_e\omega_a C_c}{Z_b'} + j\omega C_{bc}'. \quad (31)$$

These approximations are very coarse [especially (27e) applies only to certain transistors having relatively large $Z_b'$ and being biased in a restricted range of dc operating points] and do not permit accurate design, but give an impression of the order of magnitude of the neutralizing network components and the relative phases involved. The expressions can be easily correlated with the neutralized amplifier circuits discussed in the next section. If $Z_b'$ is resistive, the neutralizing network required can be determined with particular ease.

### *Common Base Configuration*

The effect of parasitic capacitances can be neglected in most cases when analyzing the common base configuration. The approximate expressions of the $h$-parameters are:

$$h_{11} \cong Z_e + Z_b'(1-a) \quad (32)$$

$$h_{12} \cong j\omega C_c Z_b' \quad (33)$$

$$h_{21} \cong a \quad (34)$$

$$h_{22} \cong \frac{1}{Z_c} \cong j\omega C_c. \quad (35)$$

If $Z_b'$ is purely resistive $(=r_b')$

$$h_{12} \cong j\omega C_c r_b' \quad (36)$$

and has a positive phase shift of approximately 90 degrees.

The other three feedback parameters can be calculated from the $h$ parameters. The *current feedback parameter* is:

$$(-g_{12}) = \frac{h_{12}}{\Delta^h} = \frac{Z_b'}{Z_e + Z_b'}. \quad (37)$$

If $Z_b' = r_b'$ the phase shift of $(-g_{12})$ is positive; if $Z_b'$ is capacitive, phase shift of $(-g_{12})$ may become negative.

The *feedback impedance parameter* is:

$$z_{12} = \frac{h_{12}}{h_{22}} = Z_b'. \quad (38)$$

The *feedback admittance parameter* is:

$$(-y_{12}) = \frac{h_{12}}{h_{11}} = \frac{j\omega C_c Z_b'}{Z_e + Z_b'(1-a)}. \quad (39)$$

At frequencies higher than $(1-a_0)\omega_a/2\pi$ but lower than $\omega_a/2\pi$, one can write in first approximation, using (27e):

$$(-y_{12}) = \omega_a C_c. \quad (40)$$

According to (40), in this frequency range, the feedback admittance is almost purely conductive, in reality, however, a reactive component is to be expected.

The above expressions can be easily correlated with the neutralizing circuits shown in the following section.

## Neutralized Amplifier Circuits

During the experimental phase of the investigation mainly General Electric grown junction $n$-$p$-$n$ transistors (type ZJ6) were used. In many experiments, however, General Electric fused junction $p$-$n$-$p$ transistors (types 2N43, 2N44, 2N45) were employed.

Due to the great variety of slightly different feasible neutralizing arrangements, a report describing all experiments completed would be impractical and repetitions in substance. Consequently, only a limited number of useful circuits will be discussed in this section. The experimentation was done to determine the practical value of the equations discussed in the preceding section and to answer the following questions:

1. Which types of neutralized circuits are most independent of dc collector voltage ($V_c$) and emitter current ($I_e$) variations? The transistor small-signal parameters being functions of the dc operating point, dependence of the neutralized condition on $V_c$ and $I_e$ can, of course, be expected.
2. Which types of neutralized circuits are most appropriate for "wide-band neutralization" of transistors? The feedback parameters $h_{12}$, $g_{12}$, $z_{12}$, $y_{12}$ are not equally complicated functions of frequency. Consequently, some of them can be represented by simpler networks than the others.

3. How does neutralization affect the maximum available power gain of transistor amplifiers?

*Audio Frequency Circuits*

In audio frequency applications the problem of stability does not arise. Neutralization may still be desired in certain applications to make input and output immittances independent of the terminal immittances.

Using the familier $T$-type equivalent circuit of the common base transistor [Fig. 14(a)], the "$h$-neutralized" common base amplifier circuit can be constructed by inspection [Fig. 14(b)]. In principle, the value of $k$ is arbitrary, but $k$ does, of course, influence the immittance levels and the maximum power gain of the stage. If $k$ is of the order of 0.1 reasonable unilateral performance can be expected: low input impedance, relatively high output impedance and, in most applications, practically unchanged power gain.

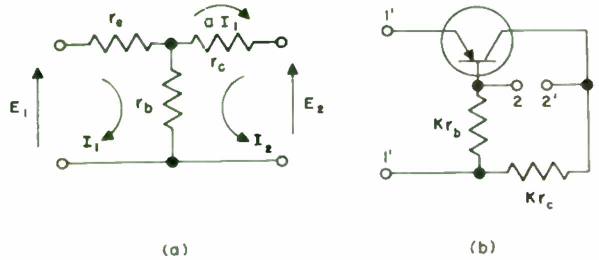


Fig. 14—(a) Low frequency equivalent of the common base amplifier, and (b) neutralized circuit.

Using the $T$-type equivalent circuit of the common collector stage [Fig. 15(a)] a "$g$-neutralized" circuit can be derived [Fig. 15(b)]. In this circuit

$$R_1 = kr_c(1 - a)$$
$$R_2 = k(r_b + ar_c).$$

The value of $k$ influences the immittance levels and the power gain of the neutralized amplifier, but is otherwise arbitrary. At very low frequencies this circuit exhibits very high input impedance and low output impedance. Including appropriate capacitances across $R_1$ and $R_2$ the frequency range of neutralization can be extended.

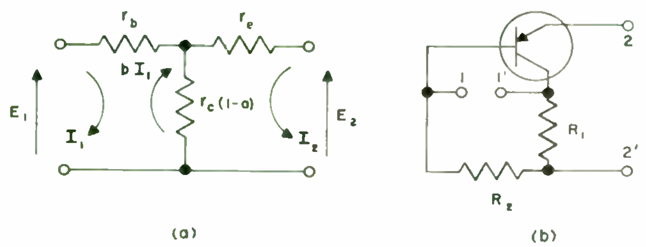


Fig. 15—(a) Low frequency equivalent of the common collector amplifier and (b) neutralized circuit.

Using a $p$-$n$-$p$ transistor with $R_2 = 440$ K ohms and $R_1 = 24$ K ohms an input impedance of 250 K ohms and output impedance of 30 ohms was measured at 500 cps, independently of source and load impedances. Maintaining the ratio $R_1/R_2$ constant and varying $R_1$, the maximum available power gain of the amplifier varied as shown in Fig. 16. This circuit is useful as a high input impedance amplifier.

Other neutralized audio frequency circuits can be designed in a similar manner and also have attractive performance characteristics: their disadvantage lies in the necessary interstage transformer the cost of which may be prohibitive in many applications.

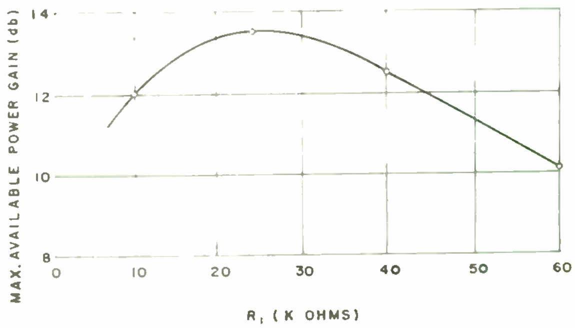


Fig. 16—Gain of a neutralized common collector stage vs $R_1$.

*High Frequency Circuits*

A considerable number of neutralized amplifiers were built and tested in the frequency range between 100 kc and 3 mc. A few typical circuits are shown in Figs. 17 to 21. The investigation was mainly concerned with the common emitter amplifier, this configuration being the most important one from a practical point of view. The common emitter configuration exhibited a strong tendency toward becoming unstable with commonly used values of terminating impedances.

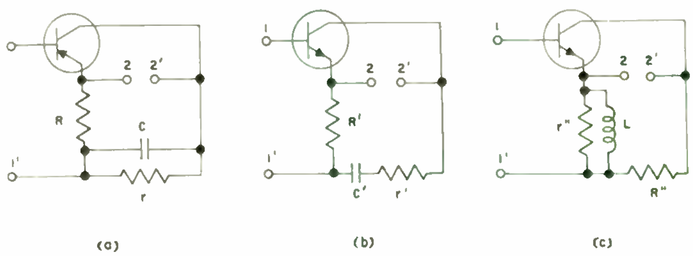


Fig. 17—$h$-neutralized common emitter circuits.

Fig. 17(a) represents an $h$-neutralized common emitter amplifier. The structure of the neutralizing network corresponds to (28). With grown junction $n$-$p$-$n$ transistors the frequency performance of the circuit shown in Fig. 17(b) was somewhat superior to that of Fig. 17(a): the amplifier could be neutralized throughout a wider frequency range, using a given neutralizing network. The reason for this behavior is connected with the complex nature of $Z_b'$. The circuits of Fig. 17(b) and 17(c) are equivalent. Using an inductive neutralizing network one can, by proper choice of the circuit elements, make the input impedance of the composite amplifier purely resistive at one frequency and the reactive component is reduced through a wide range of frequencies. Neutralization in these circuits varies moderately with $I_e$ and $V_c$.

Fig. 18 shows $g$-neutralized common emitter amplifier circuits. Figs. 18(a) and 18(b) are representative of most neutralized $n$-$p$-$n$ and $p$-$n$-$p$ stages, whereas Fig. 18(c) shows a $p$-$n$-$p$ stage with very small $C_{bc}$. The sign of the imaginary component of $g_{12}$ in 18(c) is different from that in 18(a) and 18(b). This phenomenon can be explained by considering the effect of the complex base spreading impedance of grown junction transistors and that of $C_{bc}$. The circuit of Fig. 18(b) was considerably superior to that of Fig. 18(a) from the point of view of dc operating point and frequency variations.

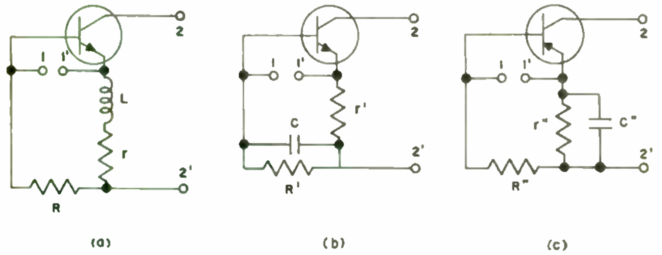

Fig. 18—$g$-neutralized common emitter circuits.

A $y$-neutralized common emitter stage is shown in Fig. 19. Due to the existence of $Z_b'$, $(-y_{12})$ is *not* purely capacitive, as indicated by (31). Neutralization in this circuit was fairly independent of variations of the dc operating point. Fig. 20 represents a $z$-neutralized common emitter amplifier. The structure of the circuit corresponds to (30).

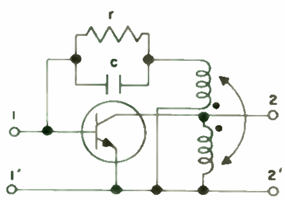

Fig. 19—$y$-neutralized common emitter circuit.

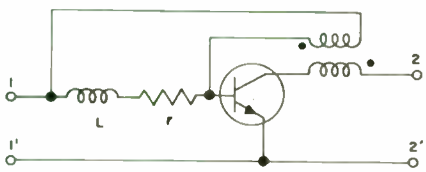

Fig. 20—$z$-neutralized common emitter circuit.

Neutralization of the common base stage can also be achieved in many ways. The $h$-neutralized common base circuit of Fig. 21(a) has been described by Angell and Keiper.[9] They have shown that the neutralizing network follows directly from the high frequency $T$-type equivalent circuit of the common base transistor. At lower frequencies perfect neutralization can be obtained by adding a resistance $r$ in parallel to the capacitor $C$ ($r$ can be omitted if $\omega \gg \omega_c$). The neutralization in this circuit is practically independent of the frequency, but varies with $I_e$ and $V_c$. A $z$-neutralized common base circuit is shown in Fig. 21(b).

If viewed closely, in all circuits described [with the exception of Fig. 21(a)] the elements of the neutralizing network required for perfect neutralization varied with the operating frequency and consequently perfect neutralization was possible only at one given frequency. However, with most circuits approximate neutralization was achieved over a relatively wide band. For example, using the circuits of Figs. 17(c) and 18(b) and adjusting the circuit elements to give "perfect" neutralization at 500 kc, no noticeable dependence of the input and output immittances on the terminal immittances was experienced at 300 and 800 kc. Consequently from a practical viewpoint wide band neutralization can easily be achieved.

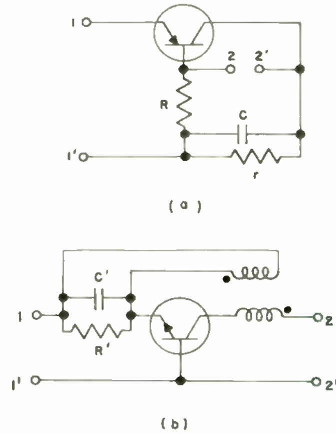

Fig. 21—Neutralized common base circuits.

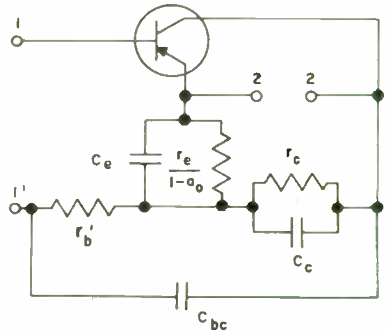

Fig. 22—Circuit with "perfect" neutralization throughout very wide frequency range.

"Perfect" neutralization throughout a wide range of frequencies could be achieved with the aid of complicated feedback networks. For example, in view of the $\pi$-type equivalent circuit of the common emitter stage,[10] the circuit of Fig. 22 could be adjusted for "perfect" neutralization throughout a very wide frequency band. However, the alignment of such a complicated circuit

[9] J. B. Angell and F. P. Keiper, "Circuit applications of surface barrier transistors," Proc. IRE, vol. 41, pp. 1709–1712; December, 1953.

[10] L. J. Giacoletto, "The study and design of alloyed-junction transistors," 1954 IRE Convention Record, Part 3, "Electron Devices and Component Parts," vol. 2, pp. 99–103.

would be exceedingly difficult. Consequently the simple circuits discussed above are definitely preferable, as they contain a maximum of three elements in the neutralizing network.

### *Neutralization and Gain*

The unneutralized transistor amplifier may become unstable provided that condition (5) is fulfilled. Consequently, the maximum unneutralized gain may become infinite and one must be cautious when referring to the maximum available power gain of an unneutralized transistor amplifier. Pritchard and Coffey have shown[11] that introducing certain constraints (e.g. matching at the output and tuning out the short circuit input impedance $h_{11}$) a constrained "maximum power gain" can be established for reference:

$$G_{unn} = \frac{|h_{21}|^2}{h_{11}^{(R)} h_{22}^{(R)} \left[1 + \sqrt{1 - H_R/h_{11}^{(R)} h_{22}^{(R)}}\right]^2}. \qquad (41)$$

The maximum available power gain of the neutralized amplifier is

$$G_n = \frac{|h_{21}''|^2}{4 h_{11}''^{(R)} h_{22}''^{(R)}}. \qquad (42)$$

Assuming now that the neutralizing network does not modify $h_{21}$, $h_{11}^{(R)}$ and $h_{22}^{(R)}$ considerably, one can see that if $H_R > 0$, neutralization will cause a decrease in gain, whereas if $H_R < 0$ an increase in gain *may* result from neutralization.

In practical circuits, however, the neutralizing network modifies $|h_{21}|^2/h_{11}^{(R)} h_{22}^{(R)}$ considerably, and

$$\frac{|h_{21}''|^2}{h_{11}''^{(R)} h_{22}''^{(R)}} < \frac{|h_{21}|^2}{h_{11}^{(R)} h_{22}^{(R)}}.$$

Consequently, even if $H_R < 0$, neutralization usually results in moderate, if any, additional gain. Gain decreases of few db, due to neutralization, have been measured using the common emitter configuration. Small decrease or increase in gain has been measured in the case of the common base circuit.

[11] R. L. Pritchard and W. N. Coffey, private communication.

## Measurement of Feedback Parameters

Neutralization, i.e., bridge methods, can be used advantageously to measure certain feedback parameters of transistors. A simplified version of a measuring arrangement used throughout this study for measuring $h_{12}$ is shown schematically in Fig. 23.

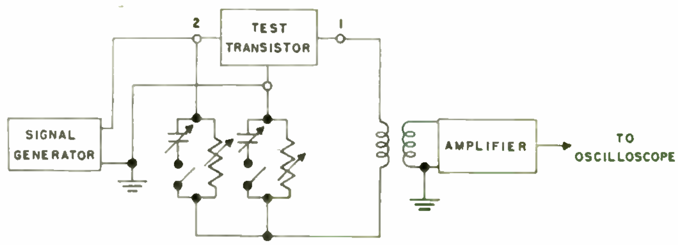


Fig. 23—Arrangement for measuring $h_{12}$.

The signal generator feeds the output 2 of the transistor and the signal appearing at the input 1 is amplified and displayed on an oscilloscope. If the neutralizing network is properly adjusted, the backward transmission becomes zero and $h_{12}$ can be computed from the values of the neutralizing network elements. Similar methods can be used for the measurement of $g_{12}$.

## Conclusion

Internal feedback of transistor amplifiers can be cancelled using different methods of neutralization. Neutralized transistor circuits are stable with input and output immittances independent of terminal immittances.

The elements of the circuits required for neutralization depend on the dc operating point and the frequency. Neutralization throughout relatively wide bands of frequencies can be achieved using simple feedback arrangements. The gain of the amplifier may be increased or decreased as a result of neutralization, depending on the nature of the internal feedback of the unneutralized amplifier.

## Acknowledgment

The authors are indebted to R. L. Pritchard for his valuable suggestions and his support of this work with unpublished material. The interest and suggestions of K. Fong and S. K. Ghandhi are also appreciated.

# Backward-Wave Oscillator Efficiency*

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*Summary*—**The theoretical and experimental results of a study of the factors which determine the efficiency of backward-wave oscillators are described. The dependence of power output upon space charge, circuit loss, beam thickness, velocity spread, and circuit mismatches is determined by a combination of theoretical and empirical means. In addition, the effect of circuit mismatches upon the starting current and frequency is discussed. The study shows that efficiency can be increased by increasing either the gain parameter $C$ or the space-charge parameter $\omega_q/\omega$. Circuit loss, beam thickness, and velocity spread are found to decrease the efficiency. The use of the results reported here permits the designer to control the factors affecting the power output and predict the efficiency of a tube with reasonable accuracy before it is built.**

## Introduction

The DISCOVERY of backward-wave oscillations[1] in traveling-wave devices has led to a new type of microwave tube—the backward-wave oscillator, or "carcinotron" as it is called by French workers.[2] Backward-wave oscillations occur when an electron beam interacts with a periodic structure which is in general shorter than that used for a traveling-wave amplifier. This fact explains the occasional occurrence of backward-wave oscillations in conventional traveling-wave amplifiers. When the electron velocity is synchronized with the phase velocity of a backward wave of a slow-wave structure, the device behaves as a backward-wave amplifier with internal positive feedback and will amplify for values of current below a critical value. Above the critical value of beam current the device will oscillate. Since backward waves are dispersive with a phase velocity which is a function of frequency, the frequency of oscillation may be changed continuously by changing the electron velocity. Thus the voltage of the electron beam may be used to tune the frequency of oscillation of a backward-wave oscillator. This brief description shows that backward-wave oscillators are both new and interesting microwave devices. It shows too that such oscillators have characteristics which are not common to other tubes.

The purpose of this paper is to add to the existing knowledge an understanding of some of the factors which determine the level of oscillation and hence the power output or efficiency to be expected. Other workers have described their experimental and theoretical results.[3–12] In general, the definitions and notation introduced by Pierce[3] will be used and it should be noted that all of the analysis contained in the paper concerns the extension of linear theory[6] to the region of oscillation. No attempt has been made to carry out a nonlinear theory[13] of backward-wave oscillators because of the complexity of such a study. The objective has been to obtain theoretical and experimental information that could be applied easily to predict the efficiency of oscillators with reasonable accuracy. The factors which are shown to influence the efficiency may be listed as follows: space charge, circuit loss, beam thickness, and velocity spread. A study of the effect of imperfect matches on the starting conditions and on the efficiency is also included. The results contained in this paper make it possible to design backward-wave oscillators with a fairly clear idea of the factors that influence the efficiency so that the amount of rf output power to be obtained may be controlled in the design. We will assume that the efficiency of a backward-wave oscillator can be written in the following form

$$\eta = \eta_0(QC)F_1(L_{db})F_2\left(\frac{\beta l}{\beta b}\right)F_3(S)F_4(R), \tag{1}$$

where each of the factors on the right is defined and discussed in the following sections. The form of (1) may not be valid if the correction factors $F_1$—$F_4$ differ greatly from unity.

The symbols used above and throughout this paper are as follows:

$b$ = velocity parameter defined by Pierce,
$C$ = gain parameter defined by Pierce,
$E(z)$ = rf circuit electric field at the plane $z$,

* Original manuscript received by the IRE, March 21, 1955.
† Stanford University, Stanford, Calif.
[1] R. Kompfner, "Backward-wave oscillator," *Bell Lab. Rec.*, vol. 31, pp. 281–285; August, 1953.
[2] P. Guenard, O. Doehler, B. Epsztein, and R. Warnecke, "New uhf oscillator valves with wide electronic tuning band," *Compt. Rend. Acad. Sci. (Paris)*, vol. 235, pp. 236–238; July, 1952.
[3] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., Inc., New York, N. Y.; 1950.
[4] L. R. Walker, "Starting currents in the backward-wave oscillator," *Jour. Appl. Phys.*, vol. 24, pp. 854–859; July, 1953.
[5] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. IRE, vol. 41, pp. 1602–1611; November, 1953.
[6] H. Heffner, "Analysis of the backward-wave traveling-wave tube," Proc. IRE, vol. 42, pp. 930–937; June, 1954.
[7] D. A. Watkins and E. A. Ash, "The helix as a backward-wave structure," *Jour. Appl. Phys.*, vol. 25, pp. 782–790; June, 1954.
[8] P. K. Tien, "Bifilar helix for backward-wave oscillators," Proc. IRE, vol. 42, pp. 1137–1143; July, 1954.
[9] M. Mueller, "Traveling-wave amplifiers and backward-wave oscillators," Proc. IRE, vol. 42, pp. 1651–1658; November, 1954.
[10] J. W. Sullivan, "A wide-band voltage-tunable oscillator," Proc. IRE, vol. 42, pp. 1658–1665; November, 1954.
[11] A. Karp, "Traveling-wave tube experiments at millimeter wavelengths with a new, easily built, space harmonic circuit," Proc. IRE, vol. 43, pp. 41–46; January, 1955.
[12] W. V. Christensen and D. A. Watkins, "Helix millimeter-wave tube," Proc. IRE, vol. 43, pp. 93–96; January, 1955.
[13] A. Nordsieck, "Theory of the large-signal behavior of traveling-wave amplifiers," Proc. IRE, vol. 41, pp. 630–637; May, 1953.

$F_1(L_{db})$ = efficiency reduction factor due to the circuit loss,
$F_2\left(\frac{\beta t}{\beta b}\right)$ = efficiency reduction factor due to beam thickness,
$F_3(S)$ = efficiency reduction factor due to velocity spread of the beam,
$F_4(R)$ = efficiency reduction factor due to reflections,
$G$ = efficiency parameter defined by (4),
$I_0$ = average convection current of the beam,
$i(z)$ = rf convection current of the beam,
$L_{db}$ = total uniform circuit loss in decibels,
$L$ = length of circuit,
$P(z)$ = power on the circuit at the plane $z$,
$P_{av}$ = power output obtained from the average value of $C$,
$QC$ = space-charge parameter defined by Pierce,
$R$ = reflection coefficient defined by (33),
$S$ = electron beam velocity distribution parameter,
$V_0$ = dc beam voltage,
$V(z)$ = rf circuit voltage at the plane $z$,
$y = 2\pi CN$ = length parameter defined by Pierce,
$z$ = distance parameter of one dimensional model,
$\alpha$ = attenuation constant of the circuit,
$\beta$ = cold circuit propagation constant,
$\beta_e$ = propagation constant of the electron beam,
$\xi$ = incremental propagation constant,
$\Delta$ = incremental propagation constant defined by (9)
$\rho$ = reflection coefficients,
$\omega$ = frequency of oscillation in radians per second,
$\omega_q$ = reduced plasma frequency in radians per second,
$\eta$ = over-all efficiency,
$\eta_0(QC)$ = basic efficiency as a function of space charge.

## General Method of Solution

The calculation of efficiency of a one-dimensional, lossless backward-wave oscillator is based on the assumption that the oscillation level of the tube is limited by the saturation of the beam current. The degree of saturation is assumed by specifying the relationship between the magnitude of the rf portion of the convection current and the dc portion of the convection current at the collector end of the interacting structure. At least two logical values may be assumed. One corresponds to the assumption that the beam has no harmonics and has the appearance of a sinusoid plus a constant value where the components have magnitudes at the collector end of the structure given by

$$i = I_0,$$

which is a choice which will be used to a large extent in the following. Another value of special interest which will be used to a smaller extent is based on the maximum theoretical value which the fundamental rf component can have with reference to the dc component. This maximum value occurs when the waveform consists of a series of equally spaced delta functions and has equal-amplitude harmonics related to the steady component by

$$i = 2I_0.$$

In general the value $i = I_0$ will be used for reasons which will become evident later.

The definition of $C$ will be used to determine the value of the rf power on the circuit of the backward-wave oscillator. The efficiency may be written thus:

$$\eta_0 = \frac{P(0)}{I_0 V_0} = \frac{|E(0)|^2}{V_0^2} \frac{1}{8\beta^2 C^3}. \tag{2}$$

The efficiency is determined from (2) by relating $E(0)$ and $i(L)$ which is easily accomplished by relating the respective waves of the convection current and the circuit electric field from the linear theory. Thus, the convection current may be written in the following manner:

$$i(L) = -j \frac{I_0 G}{2V_0 C^2 \beta_e} E(0), \tag{3}$$

where $G$ has the definition given below

$$G = \left[\frac{e^{iy\xi_1}}{(\xi_1 - \xi_2)(\xi_1 - \xi_3)} + \frac{e^{iy\xi_2}}{(\xi_2 - \xi_1)(\xi_2 - \xi_3)} + \frac{e^{iy\xi_3}}{(\xi_3 - \xi_1)(\xi_3 - \xi_2)}\right] e^{-i\beta_e L}. \tag{4}$$

Then letting the current at the collector be equal to the dc current $I_0$ and solving for $E(0)$ from (3) we obtain

$$E(0) = j \frac{2V_0 C^2 \beta_e}{G}. \tag{5}$$

Eq. (5) may be substituted into the efficiency relation given by (2) with the following result:

$$\eta_0 = \frac{C}{2|G|^2}. \tag{6}$$

Eq. (6) represents the general expression for efficiency. The efficiency may be found by substituting into (6) the value of $G$ as defined by (4) and as determined from the start oscillation condition. We should note that the efficiency relation will hold for various conditions providing that the correct values of the roots for any given degree of space charge are substituted into (4). As an example for the space-charge case the $\xi$'s are the three roots of the equation

$$\xi^3 + b\xi^2 - 4QC\xi - 4QCb + 1 = 0. \tag{7}$$

Thus it should be observed that the $\xi$'s used here differ from Pierce's $\delta$'s by the quantity $j$. The general equations (4) and (6) will be used throughout this paper.

### Efficiency for Large Space Charge

In order to solve (6) for large space charge and small $C$ it is necessary to determine the behavior of the roots and to evaluate both $2\pi CN$ and $|G|^2$. One important equation for large space charge which will be demonstrated in this section for start oscillation is given as

$$b \cong \sqrt{4QC}. \tag{8}$$

The roots of (7) for this value of $b$ all have approximately the same magnitude for large space charge so that a simplification results from the substitution

$$\Delta = b + \xi, \tag{9}$$

which will separate the two equivalent roots. Substituting (9) into (7) yields

$$\Delta^3 - 2b\Delta^2 + (b^2 - 4QC)\Delta + 1 = 0. \tag{10}$$

If the real root of (10) is called $\Delta_1$, then the other two roots may be written approximately as follows:

$$\Delta_2 \cong -\Delta_1$$
$$\Delta_3 \cong 2b. \tag{11}$$

Substitution of these approximate roots into the equation for start oscillation leads to the conclusion that

$$y\Delta_1 \cong \frac{\pi}{2} \tag{12}$$

which result depends on the assumption that $\Delta_1$ is much smaller than $b$. This assumption will be justified later. However, substituting $\Delta_1$ into (10) and applying the same assumption along with (8) yields the result

$$1 = 2b\Delta_1^2. \tag{13}$$

Combining (8), (12), and (13) leads to the conclusion

$$(2\pi CN)^2 = y^2 = \frac{\pi^2}{2}\sqrt{4QC}. \tag{14}$$

The validity of (8) may now be verified by constructing (10) from the roots of (11) and noting that

$$b^2 - 4QC = \Delta_1^2. \tag{15}$$

Eqs. (12), (13), and (14) are consistent only where $\Delta_1$ decreases with an increase of $b$. Thus the basic assumption given by (8) is consistent with the oscillation condition. By means of (8), (11), and (14) the evaluation of $|G|$ for large space charge may be carried out and the corresponding efficiency becomes

$$\eta_0 = \sqrt{4QC^3} = \frac{\omega_q}{\omega}, \tag{16}$$

where $\omega_q$ is the reduced plasma frequency and $\omega$ is the frequency of oscillation. We should realize that (16) is based on the conditions which exist at start oscillation where the start oscillation conditions are satisfied. At this point $b$ has the value given by (8). It should be pointed out that although the same result is obtained with a similar saturation assumption for a traveling-wave tube,[14] a backward-wave oscillator with large space charge is found experimentally to adhere much closer to this theoretical value.

It is interesting to note that the waveforms which exist on a backward-wave oscillator for large space charge may be written in analytic form. The current waveform is given by

$$i(z) = i(L) \sin \frac{\pi z}{2L} e^{-j\beta z}, \tag{17}$$

whereas the circuit field waveform is given by

$$E(z) = E(0) \cos \frac{\pi z}{2L} e^{-j\beta z}. \tag{18}$$

Examination of these functions shows that the beam is bunched to the maximum extent when the field is zero and conversely the convection current is zero when the field is maximum. This situation is entirely different from that which exists in a traveling-wave tube where the convection current and the circuit field both increase exponentially at the same rate. For this reason the efficiency of a backward-wave oscillator may be expected to be less than a comparable traveling-wave tube operating at the point of maximum gain.

### Theoretical Efficiency for Small Space Charge

From (4), (6), (7), and the oscillation condition, it is possible to solve for the efficiency for any value of space charge. For example, at zero space charge where

$$b = 1.52$$
$$y = 2\pi CN = 1.97,$$

we obtain the efficiency

$$\eta_0 = 0.21C. \tag{19}$$

Fig. 1 was obtained by plotting this point and others where the assumption of equivalence between the rf convection current and the dc beam current is made. It is consistent with Pierce[14] to suppose that at zero space charge the efficiency will more nearly be given by multiplying the result given by (19) by four to obtain

$$\eta_0 = 0.84C, \tag{20}$$

which is equivalent to the assumption that $i = 2I_0$. The ripples shown in Fig. 1 occur because of ripples which exist in the magnitude of the convection current as functions of distance and space charge. The process of setting the rf convection current equal to the dc beam current causes ripples to exist in the efficiency curve. For $\omega_q/\omega C > 1$ the ripples are seen to vary about the value given by (16). It is important to note that the presence of the ripples is a result of the linear theory and could hardly be expected to occur in actual tubes.

[14] Pierce, *op. cit.*, Chap. XII.

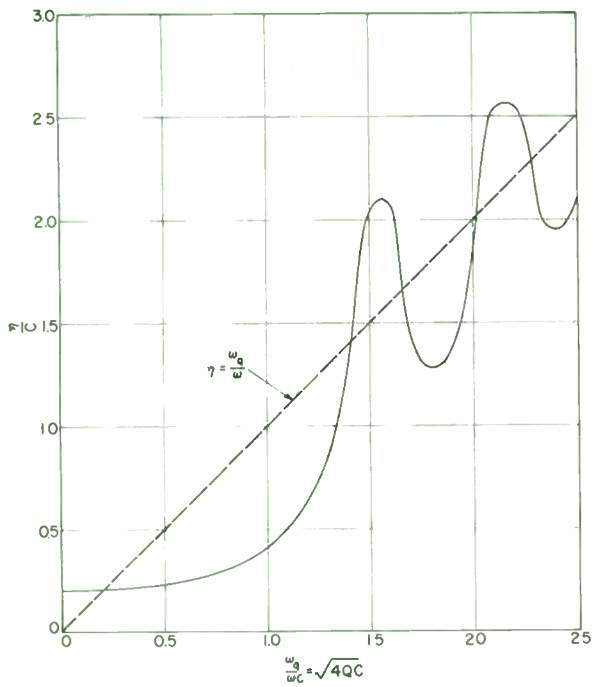


Fig. 1—Theoretical curve of $\eta/C$ vs $\omega_q/\omega C$ obtained from linear theory with the assumption that the electron beam at the collector saturates at a value of $i = I_0$.

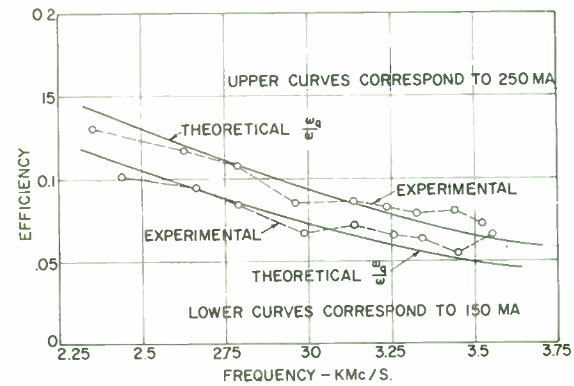


Fig. 2—Comparison of theoretical and measured efficiency of Putz and Luebke's tube operating at 150 and 250 ma.

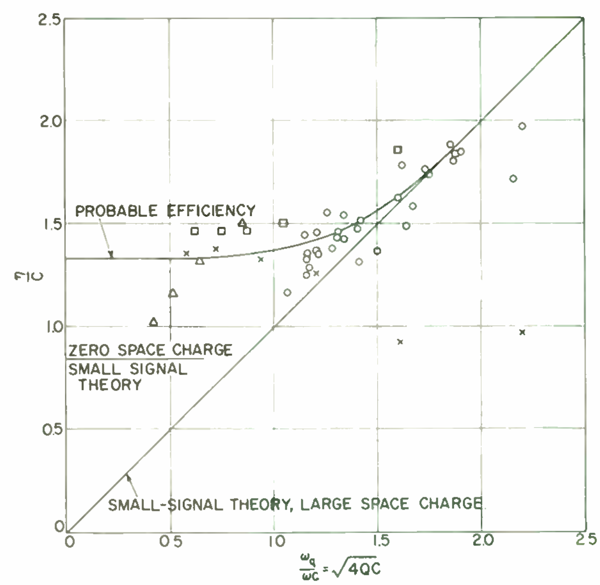


Fig. 3—Comparison of the experimental results obtained for the efficiency of a large number of backward-wave oscillators and theoretical results presented in this paper. A probable efficiency curve is drawn for the best agreement between the theory and experiment and is called the "basic" efficiency curve.

## Probable Efficiency and Experimental Results

Summarizing the results of the previous sections we find that for zero space charge

$$\eta_0 \simeq C$$

and for the large space charge

$$\eta_0 \simeq \frac{\omega_q}{\omega},$$

In the region between zero space charge and large space charge we should expect the efficiency to depend on both $C$ and $\omega_q/\omega$.

Experimental verification of the large space charge relation was obtained with a tube designed by J. L. Putz and W. R. Luebke of this laboratory. Fig. 2 shows plots of efficiency vs frequency for beam currents of 250 and 150 ma. The theoretical values of $\omega_q/\omega$ are also shown in Fig. 2. The close agreement which was achieved between the theoretical curves and the experimental curves was gratifying.

Points which were obtained from a number of tubes at various frequencies are shown in Fig. 3. The cluster of points leads to the combined theoretical and empirical curve shown also in Fig. 3. This curve will be subsequently called the "basic" efficiency curve and the procedure used to predict efficiency will be based upon this curve and upon the correction factors subsequently described. Experimental results indicate that the "basic" efficiency curve may be used when the beam current is more than two times the starting current.

## Efficiency Reduction Due to Circuit Loss

On a lossy structure it is reasonable to suppose that the output of a backward-wave oscillator is reduced because the energy is not transferred to the circuit at the output terminal but must travel through at least a portion of the lossy structure. The reduction in efficiency of an oscillator with uniformly distributed circuit loss is determined in this section by reducing each increment of power according to the amount of loss which it encounters. The magnitude of the field on a structure with large space charge is given by (18) which is not greatly different from the waveform for small space charge. From (18) power on the structure can be expressed as

$$P(z) = P_{out}(0) \cos^2 \frac{\pi z}{2L}. \tag{21}$$

The rate of change of the power with $z$ is found by

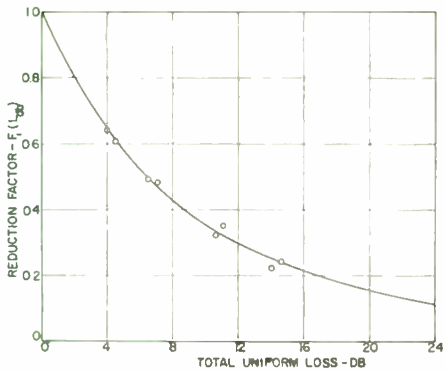


Fig. 4—Theoretical efficiency reduction factor for the case of uniform circuit loss. The experimental points were furnished by W. A. Harman.

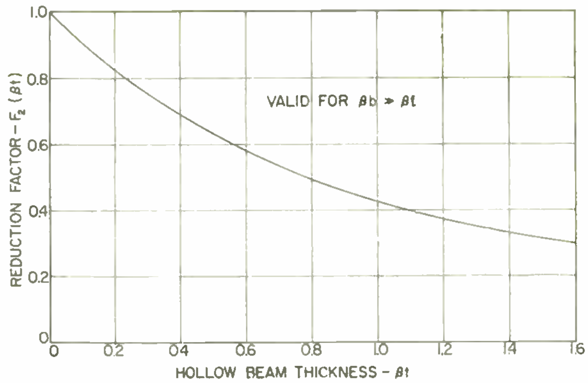


Fig. 5—Theoretical efficiency reduction factor for a thin hollow beam where fields vary as $I_1(\beta r)$ or $I_0(\beta r)$ and $\beta b \gg \beta t$.

differentiating (21) with respect to $z$:

$$\frac{dP(z)}{dz} = \frac{\pi}{2L} P_{\text{out}}(0) \sin \frac{\pi z}{L} \cdot \tag{22}$$

The ratio of transfer of power from the beam to the line is seen to be a sinusoidal function with a zero occurring at both ends of the structure and most of the transfer taking place at the midpoint. This condition may actually be suspected from the nature of the field and current waveforms. A rough approximation may be obtained by considering that all the power orginates in the center of the backward-wave oscillator and traverses one half the length of the structure in reaching the output terminal. This approximation yields the result

$$F_1 = \exp(-0.115 L_{db}). \tag{23}$$

We should expect (23) to approximate the true result. To analyze the problem more exactly we can attenuate all elements of power according to their origin on the line. This procedure yields the result

$$F_1 = \exp(-0.115 L_{db}) \frac{\cosh(0.115 L_{db})}{1 + \left(\dfrac{0.230 L_{db}}{\pi}\right)^2} \tag{24}$$

which is the exact solution. It is apparent that the approximate result appears with a modifying correction to take account of the fact that all the elements of power do not originate at the midpoint of the structure. The resulting function of $F_1$ versus loss $L_{db}$ has been plotted in Fig. 4. Also shown are experimental points provided by W. A. Harman of this laboratory. The experimental points were positioned to obtain the best fit to the theoretical curve inasmuch as the loss could not be completely removed from the oscillator so that the no-loss point could not be determined. The good agreement between the measured points and the theoretical curve is indicative of the fact that the effect of distributed loss on efficiency can be predicted by (24) or Fig. 4.

### Efficiency Reduction for a Thick Beam

In a backward-wave oscillator with a thick beam such that the rf electric field varies across the beam cross section we might expect the magnitude of the saturation convection current of an element of the beam to be proportional to the electric field which acts upon this beam element. It is consistent with the assumption of current saturation to assume that the saturation of the element of the beam which lies in the region of highest impedance will control the degree of saturation of the rest of the beam. These two assumptions have been used in the analysis which follows. The application of these assumptions leads to the relationship given below in which each element $n$ of the beam is assumed to have an efficiency $\eta_n$ which may be written as follows

$$\eta_n = \eta_0(QC) \frac{K_n}{K_{\max}}, \tag{25}$$

where the uncorrected efficiency $\eta_0(QC)$, as determined from the space charge parameter $\omega_q/\omega$ and the average value of $C$, is the same for all elements $n$ of the beam.

It is convenient for a thick beam to use Fig. 3 to determine a value for the efficiency based on the average value of $C$ and then correct this value by an amount depending on the beam thickness. This procedure makes it possible to use the same value of impedance to compute both the starting conditions and the uncorrected efficiency. The correction factor $F_2$ is defined in the following manner:

$$F_2 = \frac{P}{P_{av}} = \frac{\sum_n \eta_n I_n}{\eta_0(QC) I_T} \cdot \tag{26}$$

The combination of (25) and (26) leads to the general conclusion that

$$F_2 = \frac{K_{av}}{K_{\max}} \cdot \tag{27}$$

Eq. (27) may be solved for both a hollow beam and a solid beam. The efficiency correction factor for a hollow beam with thickness $t$ and $\beta b \gg \beta t$ may be written as

$$F_2(\beta t) = \frac{1 - e^{-2\beta t}}{2\beta t} \cdot \tag{28}$$

This expression has been plotted in Fig. 5. It is equally valid for a field variation of $I_0(\beta r)$ or $I_1(\beta r)$ as long as

$\beta b \gg 1$. A similar procedure for a solid beam with radius $b$ and with the electric field varying as $I_0(\beta r)$ yields the relation

$$F_2(\beta b) = 1 - \frac{I_1^2(\beta b)}{I_0^2(\beta b)} \qquad (29)$$

which has been plotted in Fig. 6. These reduction factors can be used to compute the efficiency when the beam is thick and where the $C$ has been computed in the usual small-signal manner. Fig. 5 for a thick hollow beam has been successfully used to compute the output power of a hollow-beam backward-wave oscillator built by L. A. Roberts of this laboratory. A comparison of the theoretical and measured power output for this oscillator is shown in Fig. 7. The agreement between the curves is well within the known accuracy of the tube parameters. It should, however, be noted that the theoretical curve of Fig. 7 includes the correction for the uniformly distributed circuit loss as given by Fig. 4.

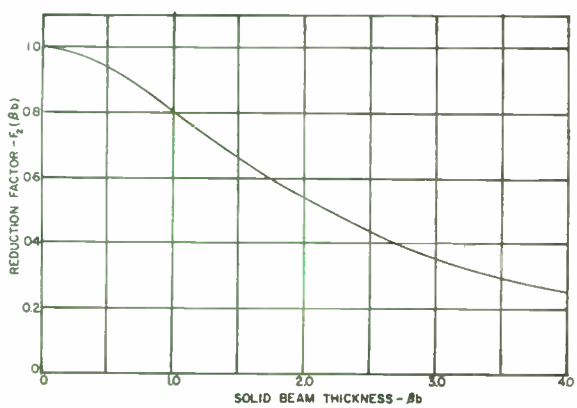


Fig. 6—Theoretical efficiency reduction factor for solid beam where the field varies as $I_0(\beta r)$.

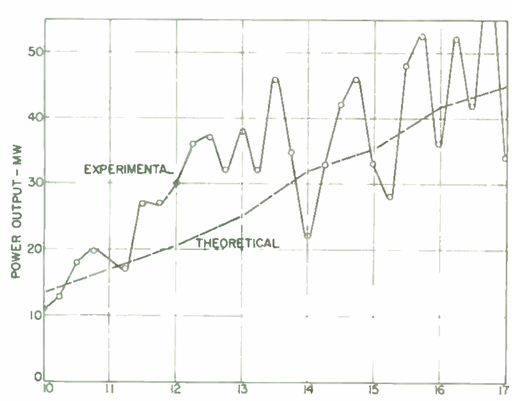


Fig. 7—Comparison of measured and theoretical power output where correction was made for the thickness of a hollow beam. The experimental data were furnished by L. A. Roberts.

### Effect of the Velocity Spread of the Beam

In an electron beam focused by means of an axial magnetic field uniform through the cathode, the dc velocity of the electrons varies over the beam cross section. This variation is caused by the potential depression produced by the charge of the electrons. Thus some elements of the beam travel faster than other elements and the synchronous beam voltage is not clearly defined. Under these conditions Watkins and Rynn[15] have shown that velocity spread in traveling-wave devices has an effect similar to space charge in the region of operation where linear theory applies. It was suspected that linear theory could not be extended in this case to the region of nonlinearity. No reasonable theory has been developed to predict the effect of velocity spread of the electrons of the beam on the efficiency of backward-wave oscillators. This section concerns an experimental approach to the problem. An actual backward-wave oscillator was modified to make possible a measurement of the effect of velocity spread on efficiency.

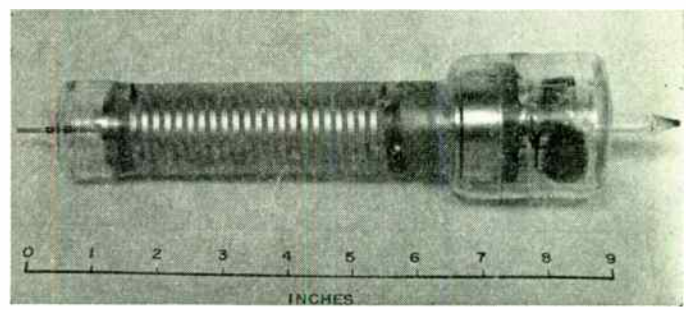


Fig. 8—Photograph of the 500–1,000 mc backward-wave oscillator used to measure the effect of velocity spread on the efficiency. The oscillator used a hollow electron beam placed close to the helix.

This measurement was performed on the backward-wave oscillator shown in Fig. 8. The beam was 0.005 inch thick and was spaced about 0.005 inch from the helix which was about 1.3 inches in diameter. This normal arrangement was modified by inserting a 0.5 inch diameter cylinder down the axis of the tube, as suggested by P. D. Lacy[16] of the Hewlett-Packard Co. The cylinder could be operated at any desired voltage with respect to the helix in order to introduce artificially a velocity spread to the electrons of the beam. In order to obtain meaningful data all of the parameters of the tube were held constant except the helix voltage and the cylinder voltage. The frequency was maintained constant by adjusting both the helix and cylinder voltages to keep the "average" beam voltage constant. Measurements were made at five different frequencies across the band. The measured data are shown in Fig. 9 (next page) where the reduction in power output expressed by $F_3(S)$ is plotted against the parameter

$$S = \left(\frac{\Delta V}{4CV_0}\right)^2, \quad C \cong 0.06 \text{ (for these measurements)}$$

which was introduced by Watkins and Rynn.[15] Since the

[15] D. A. Watkins and N. Rynn, "The effect of velocity distribution on traveling-wave tube gain," *Jour. Appl. Phys.*, vol. 25, pp. 1375–1379; November, 1954.
[16] P. D. Lacy, private communication.

gain parameter $C$ did not vary for these measurements, it could not easily be determined whether the parameter $S$ is the correct one to use for the reduction of efficiency.

The examination of the experimental points led to their presentation on semi-log paper where they are seen to trace a straight line. The magnitude of the reduction down to 10 per cent, although seemingly large, corresponds to a relatively large voltage variation. Thus, at $S=0.71$ the voltage drop across the beam was about 8 v. This is quite large when compared to the average value of 36 v. The straight line, drawn for the best fit, represents an empirical curve showing efficiency reduction.

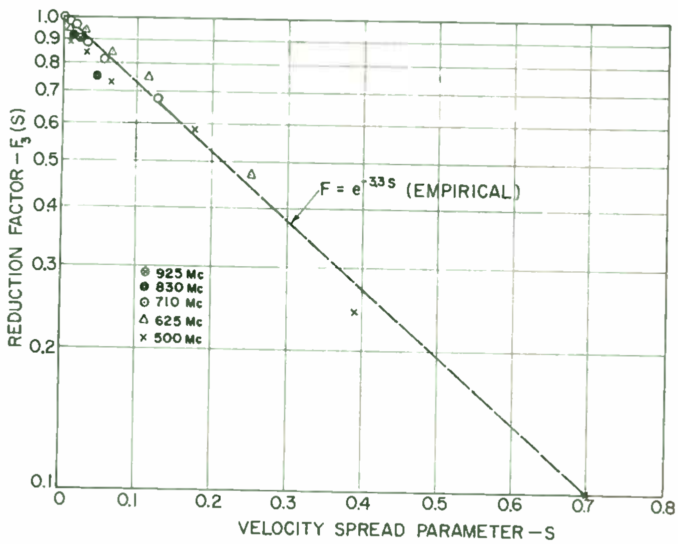


Fig. 9—The experimental measurement of efficiency reduction caused by velocity spread in the 500–1,000 mc backward-wave oscillator. The empirical curve was drawn for the best fit to the experimental data.

## Effect of Reflections on Starting Conditions

The gain expression for small $C$, zero space charge, and no loss is given by the relation

$$\frac{V(L)}{V(0)} = \frac{e^{iy\xi_1}}{(\xi_1 - \xi_2)(\xi_1 - \xi_3)(\xi_1 + b)} + \frac{e^{iy\xi_2}}{(\xi_2 - \xi_1)(\xi_2 - \xi_3)(\xi_2 + b)} + \frac{e^{iy\xi_3}}{(\xi_3 - \xi_1)(\xi_3 - \xi_2)(\xi_3 + b)}. \tag{30}$$

The method used to solve for the start oscillation condition is to set $V(L)=0$ and solve (30) simultaneously with the root equation

$$\xi^3 + b\xi^2 + 1 = 0. \tag{31}$$

A different approach was used to solve for the effect of reflections. The equations were solved as a function of $V(L)/V(0)$ to obtain the starting parameters when $V(L)/V(0)$ is not zero. The quantity $V(L)/V(0)$ is the ratio between the voltage that is applied at the collector end of the transmission line and the output voltage. For convenience let us define the ratio as

$$Re^{i\theta} = \frac{V(L)}{V(0)}. \tag{32}$$

The quantity $R$ is a real number which depends on the matches at the ends of the tube and the loss of the circuit as defined by the relation

$$R = \rho_1\rho_2 e^{-\alpha L}, \tag{33}$$

where $\rho_1$ and $\rho_2$ are the reflection coefficients and $\alpha$ is the attenuation constant. $R$ will be zero when the reflection coefficients are zero or the loss is infinite. The method of solution was to expand about the point $R=0$ for deviations of $y$ and $b$ in the form of a Taylor series:

$$\begin{aligned} Re^{i\theta} &= f(y, b) \\ &= f(1.97, 1.52) + f_y(1.97, 1.52)\Delta y \\ &\quad + f_b(1.97, 1.52)\Delta b. \end{aligned} \tag{34}$$

The first term of the series is identically zero, and only the first-order effects have been included to make the problem soluble. For convenience let us define

$$\begin{aligned} f_y(1.97, 1.52) &= Ae^{i\alpha} \\ f_b(1.97, 1.52) &= Be^{i\beta}. \end{aligned} \tag{35}$$

Then separating (34) into a real and an imaginary equation and solving simultaneously yields the result

$$\Delta y = \frac{R \sin(\beta - \theta)}{A \sin(\beta - \alpha)} \tag{36}$$

$$\Delta b = \frac{R \sin(\alpha - \theta)}{B \sin(\alpha - \beta)}. \tag{37}$$

It has been demonstrated here that if we find the value of $f_y(1.97, 1.52)$ and $f_b(1.97, 1.52)$ then we can write $\Delta y$ and $\Delta b$ in analytical form to give the variation for values of $R>0$. From (36) and (37) we see that $\Delta y$ and $\Delta b$ are sinusoidal functions of $\theta$ with amplitudes which are proportional to the value of $R$. Thus we see that the effect of reflections, at least on the starting conditions, should be sinusoidal in nature and should vary about the mean position which would exist in the absence of the reflections. The evaluation of $f_y(y,b)$ at the point $y=1.97$ and $b=1.52$ where the $\xi$'s are given by (31) yields the result

$$\begin{aligned} A &= 1.53 \\ \alpha &= 223 \text{ degrees} \end{aligned} \tag{38}$$

and the evaluation of $f_b(y,b)$ at the point $y=1.97$, $b=1.52$ yields the result

$$\begin{aligned} B &= 1.045 \\ \beta &= 296.7 \text{ degrees.} \end{aligned} \tag{39}$$

The results given by (38) and (39) may be substituted into (36) and (37) to give the absolute magnitude of $\Delta y$ and $\Delta b$

$$|\Delta y| = 1.68\,R \tag{40}$$

$$|\Delta b| = -1.00\,R. \tag{41}$$

These values represent the mathematical solution of the problem. From (40) we can obtain

$$\frac{|\Delta I_0|}{I_0} = R \tag{42}$$

which is the desired result expressing the change of starting current as a function of $R$. From (36) and (42) we can see that $\Delta I_0/I_0$ is a sinusoidal function with an amplitude $R$ which shows that the starting current $I_0+\Delta I_0$ varies in a sinusoidal manner about a mean position $I_0$ corresponding to the starting current in the absence of reflections.

A similar expression for the frequency variations may be obtained from the definition of $b$ as given below:

$$\frac{|\Delta f|}{f} = 2CR\frac{V_0}{f}\left|\frac{df}{dV_0}\right|, \tag{43}$$

where $C$ is the gain parameter, $V_0$ is the beam voltage, and $df/dV_0$ is the tuning rate. On a helical structure (43) may be written as follows:

$$\frac{|\Delta f|}{f} = CR(1 - ka). \tag{44}$$

It has thus been possible to find analytic relationships for the magnitude of the variations of starting current and frequency as a function of the reflection parameter $R$. Both (42) and (44) have been verified experimentally by measurement on the tube of Fig. 8 with the value of $C$ at start oscillation used in the comparison with (44).

### Effect of Reflections on the Efficiency

It has been observed that poor matches on a backward-wave oscillator produce variations in the power output and hence in the efficiency. In a manner similar to that used to determine the effect of reflections on the starting conditions the effect on the efficiency will be determined in this section. For this case the efficiency must be determined as a function of $\Delta b$ and $\Delta y$.

Let us define the relation

$$G = G_0 + G_b\Delta b + G_y\Delta y, \tag{45}$$

where $G_0$ is the value of $G$ without reflections corresponding to zero space charge, and $G_b$ and $G_y$ are the partial derivatives of $G$ with respect to $b$ and $y$, respectively, evaluated at $y=1.97$ and $b=1.52$. Let us define a correction factor $F$ as follows

$$F_4 = \frac{\eta}{\eta_0} = \frac{|G_0|^2}{|G|^2}. \tag{46}$$

Then substituting from (45) we obtain

$$F_4 = \frac{1}{\left|1 + \frac{2G_b}{G_0}\Delta b + \frac{2G_y}{G_0}\Delta y\right|}. \tag{47}$$

By substituting (36) and (37) into (47) we obtain (48) which expresses $F$ as a function of $R$ and $\theta$

$$F_4 = \frac{1}{\left|1+\frac{2R}{G_0}\left[\frac{G_b}{B}\frac{\sin(\alpha-\theta)}{\sin(\alpha-\beta)}+\frac{G_y}{A}\frac{\sin(\beta-\theta)}{\sin(\beta-\alpha)}\right]\right|}. \tag{48}$$

The maximum or minimum value of $F$ can be found easily when the second term of the denominator is less than unity. For this case it is approximately true that the components which are at right angles with unity are negligible and only the real part of the second term is important. To simplify the expression let us define

$$B_3 = \text{Real Part}\left[\frac{G_b}{G_0}\right] \tag{49}$$

$$A_3 = \text{Real Part}\left[\frac{G_y}{G_0}\right]. \tag{50}$$

The definitions given by (49) and (50) may be substituted into (48) and the correction factor may be expressed in the following manner:

$$F_4 = \frac{1}{1 + \frac{2R}{\sin(\alpha-\beta)}\left[\frac{B_3^2}{B^2} + \frac{A_3^2}{A^2} + 2\frac{A_3B_3}{AB}\cos(\alpha-\beta)\right]\cos(\theta+\theta_0)} \tag{51}$$

where $\theta_0$ is the phase at $\theta=0$ and is unnecessary to the present treatment. The efficiency is seen to vary in a sinusoidal manner about the mean position with the maximum efficiency and the minimum efficiency occurring when $\cos(\theta+\theta_0)=\pm 1$. Since values of $A$ and $B$ are already available, it is only necessary to evaluate $A_3$ and $B_3$. The parameters of (49) and (50) may be evaluated at the point $y=1.97$ and $b=1.52$ to yield the results

$$G_y(1.97, 1.52) = 1.227\,e^{j297.3\text{ degrees}} \tag{52}$$

$$G_b(1.97, 1.52) = 1.00\,e^{j191.5\text{ degrees}} \tag{53}$$

$$G_0(1.97, 1.52) = 1.53\,e^{j318\text{ degrees}} \tag{54}$$

Using (49) and (50) we may then evaluate $A_3$ and $B_3$ by substituting the results contained in (52), (53), and (54). Finally, from (51) we obtain the equation for the reduction factor

$$F_4 \cong 1 + 1.42\,R\cos(\theta+\theta_0). \tag{55}$$

The reduction factor does indeed vary in a sinusoidal manner about the zero reflection case. All of the previous work has been based on the assumption that

$$1.42\,R \ll 1,$$

ereby restricting the region of validity of the result ven by (55). No quantitative experimental verification (55) has been made although qualitative agreement s been observed.

## Conclusion

The results presented in this paper enable the designer a backward-wave oscillator to predict the efficiency d output power at the time of the initial design. Thus ckward-wave oscillators can now be designed not only the basis of whether they will oscillate but also on the sis of how much rf power they will produce. The effect the important parameters which control the efficiency has been presented. The effect of space charge has en considered to be the primary factor and all other ctors have been considered to produce corrections to is basic efficiency. The basic efficiency $\eta_0$ may be obined from Fig. 3 when the value of the space-charge rameter $QC$ is calculated. The correction factors, $F_1$, , and $F_3$ for circuit loss, beam thickness, and velocity read are presented on Figs. 4, 5, 6, and 9, respectively. By making use of these curves the designer can obtain a reasonably accurate value for the efficiency. If the matches are imperfect then the magnitude of the variations in efficiency can be obtained from (55). Eqs. (42) and (43) also permit the designer to calculate the magnitude of the variations in the starting current and the frequency. These results therefore permit the design of backward-wave oscillators with considerably more confidence than was formerly possible.

## Acknowledgment

The research reported in this paper is based on a Ph.D. dissertation submitted to the Department of Electrical engineering at Stanford University. It was supported primarily by a gift to the University from the Hewlett-Packard Company and partly by the U. S. Army Signal Corps, the U. S. Air Force, and the U. S. Navy (Office of Naval Research) under Contract N6onr 251(07). Financial assistance was also received from a Radio Corporation of America Fellowship in Electronics under the National Research Council.

# The Effects of Junction Shape and Surface Recombination on Transistor Current Gain—Part II*

K. F. STRIPP† AND A. R. MOORE†

*Summary*—**Previous work demonstrated the importance of surce recombination and junction shape on the transistor current plification factor $\alpha$ by means of a two-dimensional conducting per analog.[1] This is now extended theoretically and experimentally other cases which are also of practical importance.**

**Exact analytical solutions have been obtained for the collector-base current amplification factor, $\alpha_{cb}$, for plane-parallel (grown nction) transistors of rectangular and round cross section including rface and volume recombination. For the case in which the surface combination velocity $s$ is small and the volume lifetime $\tau$ is large, ese equations reduce to**

$$\alpha_{cb} = \left(\frac{s}{K} + \frac{T}{\tau}\right)^{-1}$$

**ere $K$ and $T$ are geometrical constants simply related to the base dth $W$ and the cross-sectional area. The range of validity of this uation has been investigated.**

**For the more complex geometries usually found in alloy transiss, $\alpha_{cb}$ is found to be of the form:**

$$\alpha_{cb} = \left(\frac{s}{K} + \frac{T}{\tau}\right)^{-1}[1 + F(\tau, s)],$$

**ere $K$ and $T$ are again geometrical constants. $F(\tau, s)$ is a small sitive correction term which goes to zero as $s$ approaches zero and pproaches infinity.**

**Although the constants $K$ and $T$ are not readily evaluated anaically, in most practical cases volume recombination can be negted, and then**

$$\alpha_{cb} = \frac{K}{s}[1 + f(s)].$$

**By means of a three-dimensional electrolytic conductance analog, $K$ and $f(s)$ have been evaluated for various geometries typical of alloy junction transistors. Since $f(s)$ is small, the constant $K$ serves as a geometrical figure of merit against which various emitter-collector configurations can be judged. The most striking result of this study is that for a given minimum junction spacing $W$ and given emitter and collector diameter, the value of $K$, and hence $\alpha_{cb}$, can be varied over a considerable range by changes in emitter and collector penetration. In particular, the highest figure of merit is obtained by combining essentially nonpenetrating emitters (i.e., lying on the wafer surface) with collectors which penetrate as far as necessary to give the required minimum spacing $W$.**

## Introduction

THIS PAPER presents an extension, both theoretical and experimental, of a study reported in a previous paper.[1] In the theoretical portion the general analytical approach to the determination of the fate of injected minority carriers in semiconductor devices, in the steady state, is formulated. It is carried to explicit solution for certain simple geometries, corresponding to transistors of the grown junction type with

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[1] A. R. Moore and J. I. Pankove, "The effect of junction shape and surface recombination on transistor current gain," Proc. IRE, vol. 42, pp. 907–913; June, 1954. Hereafter called Part I.

rectangular or circular cross section. In the case of geometrically more complex structures, such as the alloy junction TA-153 transistor,[2] an implicit perturbation treatment is applied to deduce the general form to which the dependence of current gain on the surface recombination velocity and bulk lifetime must reduce for small values of the former and large values of the latter. These results prove of considerable use in analyzing the experimental findings.

On the experimental side, the conductance analog introduced in Part I is made three-dimensional by an adaptation of the familiar electrolytic tank. In this way it is possible to deduce the transistor current gain $\alpha$ directly from measured tank currents, and field plotting is then unnecessary. This represents a large saving in effort and an increase in accuracy, once the tank has been properly constructed. Presented here are the results of such an analog survey of the TA-153 alloy transistor, carried out to establish the dependence of $\alpha$ on surface recombination velocity and certain geometrical factors in this device. It is found that the shape of the junctions is quite important in this regard; for a given minimum emitter-collector separation, flatness (small penetration) of emitter offers considerable advantage in limiting loss of minority carriers due to surface recombination.

## Formulation of the Problem

The fate of injected minority carriers in semiconductors is of concern in numerous important instances, both from the theoretical and the practical points of view. This paper deals with the particular case of injected minority carriers in junction transistors when the system is in the steady (time independent) state and when the motion of the carriers is diffusion controlled. For simplicity of expression, the text will speak of holes in *p-n-p* devices and will denote by $P$ the excess hole density in the *n*-region. The results will be equally valid in the *n-p-n* case, by suitable interchanging of electrons for holes and use of the applicable diffusion coefficients.

In general, the emitter current of the *p-n-p* transistor shown schematically in Fig. 1 consists of holes injected into the base and electrons flowing from the base into the emitter. Since in alloy transistors the emitter section has very high conductivity compared to the base, the emitter current consists almost entirely of holes.[3] It will be assumed that the number of holes injected per second into the base at the emitter constitutes the entire emitter current $I_e$. Those holes which survive the trip to the collector comprise the collector current $I_c$. Some, however, are lost by volume recombination in the base region. Their total number per second is designated as the volume current, $I_v$. Still others are lost by surface recombination at the free surface of the base region. This total number per second is designated as the surface current, $I_s$. These currents are related by

$$I_e = I_c + I_v + I_s, \tag{1}$$

and the sum of the volume and surface currents constitutes the base current, $I_b$.

$$I_b = I_v + I_s. \tag{2}$$

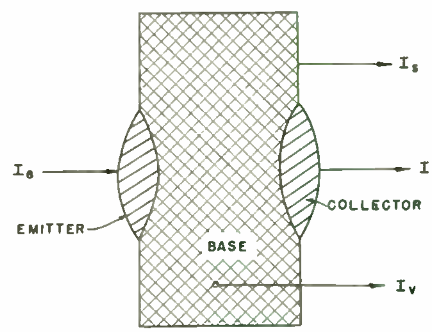



Fig. 1—Schematic representation of an alloy transistor.

Two particular ratios of currents are of interest in transistor performance:

$$\alpha_{ce} = \frac{I_c}{I_e} = 1 - \frac{I_b}{I_e} = \frac{\alpha_{cb}}{1 + \alpha_{cb}} \tag{3}$$

$$\alpha_{cb} = \frac{I_c}{I_b} = \frac{\alpha_{ce}}{1 - \alpha_{ce}}, \tag{4}$$

and it is the purpose here to extend the study, begun in Part I, of the dependence of these quantities on surface recombination velocity, bulk lifetime, and geometry.

In the diffusion controlled steady state, the behavior of the system is contained in the differential equation[4]

$$D\nabla^2 P - \frac{P}{\tau} = 0; \text{ in the base volume,} \tag{5}$$

subject to the boundary conditions:

$$-\overrightarrow{D\nabla P}\cdot\vec{n} = sP; \text{ at the free base surface,} \tag{6}$$

$$P = P_e; \text{ at the emitter junction surface,} \tag{7}$$

$$P = 0; \text{ at the collector junction surface.} \tag{8}$$

Here $D$ is the diffusion constant for holes, $\tau$ the bulk lifetime, $s$ the surface recombination velocity, and $\vec{n}$ the unit surface vector. The desired currents for evaluating (3) and (4) are then certain surface and volume integrals of the solution of these equations, i.e.,

$$I_e = qD\int \left| \nabla P \right| d\sigma; \text{ over emitter junction surface,} \tag{9}$$

$$I_c = qD\int \left| \nabla P \right| d\sigma; \text{ over collector junction surface,} \tag{10}$$

[2] R. R. Law, C. M. Mueller, J. I. Pankove, and L. Armstrong, "A developmental *p-n-p* junction transistor," Proc. IRE, vol. 40, p. 1352; November, 1952.

[3] Moore and Pankove, *op. cit.*, p. 907; and W. Shockley, "Electrons and Holes in Semiconductors," D. Van Nostrand and Co., New York, N. Y., p. 102; 1950.

[4] Moore and Pankove, *loc. cit.;* and Shockley, *op. cit.*, p. 320.

$$I_v = \frac{q}{\tau} \int P dV; \text{ over the base volume,} \tag{11}$$

$$I_s = qs \int P d\sigma; \text{ over free base surface.} \tag{12}$$

Here $q$ is the charge of a hole.

### Solution for Special Cases

Eq. (5) with its boundary conditions (6), (7), and (8) can be solved explicitly for certain geometrically simple cases.[5] Here the computed $\alpha_{ce}$ for two of these is presented, leaving the mathematical details for the Appendices. These geometries correspond to grown (flat) junction types with circular and rectangular cross sections, respectively.

For the circular case, with the emitter-collector spacing $W$ and cross-sectional radius $R$,

$$\alpha_{ce} = 1 - \frac{\sum_n \alpha_n \tanh\left[\frac{W}{2R}\sqrt{n^2 + R^2/D\tau}\right]}{\sum_n \alpha_n \coth\left[\frac{W}{R}\sqrt{n^2 + R^2/D\tau}\right]}; \tag{13}$$

where the sums are over all the successive positive roots of an equation involving the zero*th*-order Bessel function, $J_0$;

$$nJ_0'(n) + \frac{Rs}{D} J_0(n) = 0; \tag{14}$$

and where

$$\alpha_n = \frac{\sqrt{n^2 + R^2/D\tau}}{n^2(n^2 + R^2s^2/D^2)}. \tag{15}$$

Similarly, for the rectangular case, with cross sectional dimensions $2a$ and $2b$,[6]

$$\alpha_{ce} = 1 - \frac{\sum_n \sum_m \alpha_n\alpha_m\sqrt{n^2 + m^2 + 1/D\tau}\tanh\left[\frac{W}{2}\sqrt{n^2 + m^2 + 1/D\tau}\right]}{\sum_n \sum_m \alpha_n\alpha_m\sqrt{n^2 + m^2 + 1/D\tau}\coth\left[W\sqrt{n^2 + m^2 + 1/D\tau}\right]}; \tag{16}$$

where the sums are over all the positive roots of the equations

$$n \tan na = \frac{s}{D}, \qquad m \tan mb = \frac{s}{D}, \tag{17}$$

and where

$$\alpha_n = \frac{\sin^2 na}{n^2a + \frac{s}{D}\cos^2 na}$$

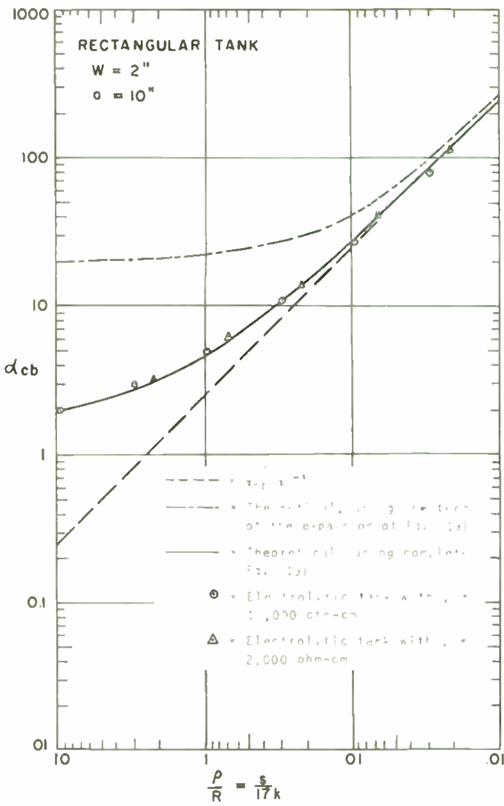



Fig. 2—Comparison of theoretically computed current gain and measurements in the tank of Fig. 3. The abscissa gives surface recombination velocity $s$ in terms of the linear scale-up factor $k$.

$$\alpha_m = \frac{\sin^2 mb}{m^2b + \frac{s}{D}\cos^2 mb}. \tag{18}$$

As a special case of (16), let $s \to 0$ on the face $y = b$, and let $\tau \to \infty$ (no volume recombination). This gives the result for a "two-dimensional" transistor in which volume recombination is negligible:

$$\alpha_{ce} = 1 - \frac{\sum_n n\alpha_n \tanh\left(\frac{nW}{2}\right)}{\sum_n n\alpha_n \coth(nW)}. \tag{19}$$

In the experimental part of this bulletin, the result (19) is compared with measurements in an electrolytic tank. The behavior of this series is depicted in Fig. 2, above, for a particular choice of $W$ and $a$.

### Perturbation Approach

In geometrically complex arrangements, such as the TA-153 alloy junction transistor, explicit and analytical

[5] Geometrically simple means here structures in which emitter, collector, and base surface are each individual co-ordinate surfaces in an orthogonal co-ordinate system.

[6] A mathematically equivalent solution to this problem has been presented by E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, p. 1161; 1954. See also reference 13.

solution has not been achieved.[7] Nevertheless, interesting general results can be deduced for such structures by a somewhat implicit use of the perturbation (iteration) approach. Assume that when there is no volume or surface recombination ($\tau=\infty$, $s=0$) the solution to (5)–(8) is $P^0$. When both recombinations are present, a first approximation, whose validity is related to the largeness of $\tau$ and the smallness of $s$, consists of substituting $P^0$ into (9)–(12) to compute the currents. The integrals are then independent of $\tau$ and $s$, and lead to

$$\alpha_{ce} \approx 1 - \frac{T}{\tau} - \frac{s}{K}, \tag{20}$$

or

$$\alpha_{cb} \approx \left(\frac{T}{\tau} + \frac{s}{K}\right)^{-1}; \tag{21}$$

where

$$T = \frac{\int_v P^0 dV}{D\int_e |\nabla P^0|\, d\sigma} \tag{22}$$

and

$$K^{-1} = \frac{\int_s P^0 d\sigma}{D\int_e |\nabla P^0|\, d\sigma} \tag{23}$$

are certain geometrical factors.

The approximate expressions (20) and (21) can be made formally correct in the following way. The actual hole concentration in the volume and on the surface will always be smaller than $P^0$ when recombination is taking place, since $P^0$ neglects recombination. Hence the insertion of $P^0$ into (11) and (12) progressively overestimates the volume and surface currents as $\tau$ decreases and $s$ increases. Thus it can be expected, e.g., that

$$\alpha_{cb} = \left(\frac{T}{\tau} + \frac{s}{K}\right)^{-1}[1 + F(\tau, s)], \tag{24}$$

where $F(\tau, s)$ is some positive function which vanishes as both $\tau \to \infty$ and $s \to 0$. The corrected $\alpha_{ce}$ is similarly obtained by using (24) in (3).

In simple cases one can evaluate the constants $T$ and $K$ analytically. As a demonstration one can consider the structures for which explicit solutions were obtained in the previous section, i.e., plane parallel junctions of circular and rectangular cross section. The zero-order hole density is

$$P^0(Z) = P_e \frac{W - Z}{W}, \tag{25}$$

where $Z$ is the distance from the emitter. Then the integrals required in (22) and (23) are

$$\int_v P^0 dV = P_e \frac{W}{2} \times \text{cross-sectional area}, \tag{26}$$

$$\int_e |\nabla P^0|\, d\sigma = P_e \frac{1}{W} \times \text{cross-sectional area}, \tag{27}$$

$$\int_s P^0 d\sigma = P_e \frac{W}{2} \times \text{cross-sectional perimeter}, \tag{28}$$

for both these cases. Consequently, the approximate form (20) becomes

$$\alpha_{ce} \approx 1 - \frac{W^2}{2D\tau} - \frac{W^2 s}{DR} \tag{29}$$

for the circular cross section; and

$$\alpha_{ce} \approx 1 - \frac{W^2}{2D\tau} - \frac{W^2 s}{D}\left[\frac{1}{a} + \frac{1}{b}\right] \tag{30}$$

for the rectangular. These expressions are just what one obtains from the series solutions (13) and (16) by neglecting all but the lead terms and by replacing the remaining functions by their small argument power expansions.

In more complicated cases, one can seek to evaluate the various quantities empirically. The remainder of this paper deals primarily with such a study of the TA-153. For the most part, it will be assumed that the bulk lifetime is sufficiently long so that volume recombination is negligible. Then (24) becomes

$$\alpha_{cb} = \frac{K}{s}[1 + F(s)]; \qquad F(s) \to 0,\ s \to 0, \tag{31}$$

and similarly

$$\alpha_{ce} = 1 - \frac{s}{K}\left[\frac{1}{1 + f(s) + s/K}\right]. \tag{32}$$

Both the value of $K$ and the form of $f(s)$ are deduced from analog measurements.

## Electrolytic Tank Analog

In Part I it was shown that the process of diffusion-controlled minority carrier flow in a semiconductor with surface recombination is analogous to field controlled electrical current flow in a conductive medium whose surface is divided into small, isolated segments (tabs), each leaking current to ground through suitable resistors.[8] The analogs constructed at that time were two-dimensional, and in order to deduce the three-dimensional result, it was necessary to plot fields and to weigh radially the computed local currents. Since that time three-dimensional analogs have been constructed by

[7] See, however, J. Laplume "Calcul du courant de recombination en surface dans le transistor à jonction obtenu par fusion," *Compt. Rend.*, vol. 238, pp. 1107–1109; March 8, 1954, for an essay in this direction.

[8] The analogy remains valid when volume recombination is also assumed, if one simulates the volume recombinations by introducing additional bleeding resistors to ground from a suitably fine grid of connections to the interior of the conductor.

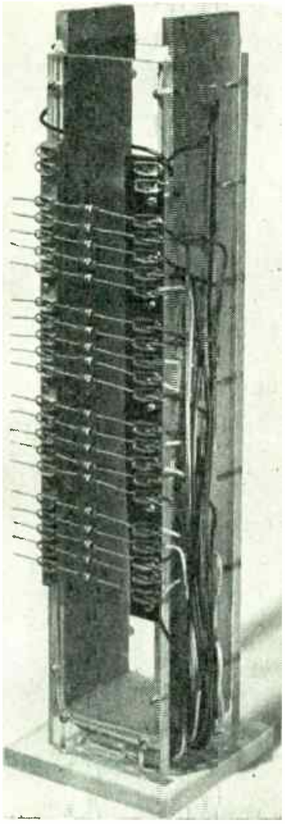


Fig. 3—Photograph of the electrolytic tank with plane-parallel geometry (grown-junction transistor analog).

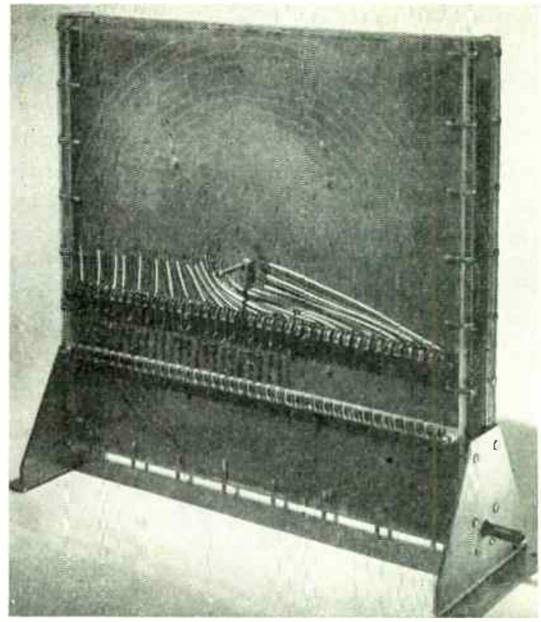


Fig. 4—Photograph of the electrolytic tank with geometry suitable for alloy transistor analog.

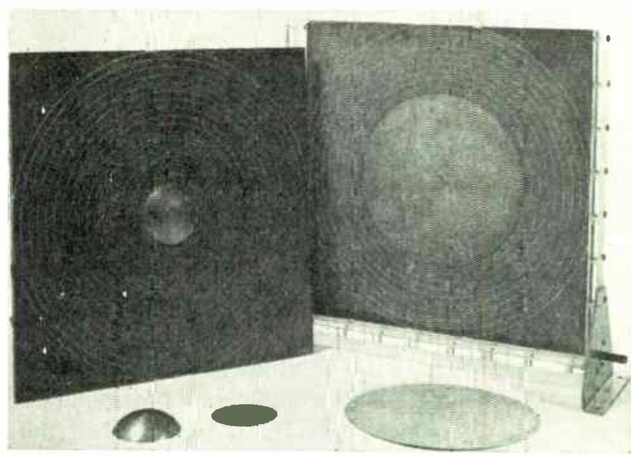


Fig. 5—Interior of the electrolytic tank of Fig. 4, showing the surface recombination rings and extra interchangeable electrodes.

adaptation of the familiar electroytic tank. In these the desired total currents can be measured directly.

The analogy between the diffusion and the conductance devices is given by the equation

$$\frac{s}{D} = k\frac{\rho}{AR}; \tag{33}$$

where $\rho$ is the electrical conductivity of the analog medium, $A$ is the area of a surface tab, $R$ is the bleeder resistance, and $k$ is the linear scale-up factor from the semiconductor. Variation in simulated $s$ can be achieved by changing either $\rho$ or $R$. In covering a wide range of values of $s$, it proves advisable to do both in order to keep the over-all resistance of the analog within reasonable bounds.

To imitate a surface of constant $s$, it is not necessary that all the tabs have the same size, but merely that the product of each tab area by its bleeder resistance have the same value. The analogy (33) is strictly valid only when the dimensions of each surface tab approach zero in any direction in which $P$ changes. In practice, of course, one must be satisfied with making such tab dimensions reasonably small. The tabs need not be small, however, along directions of constant $P$. When these directions are apparent, from symmetry or otherwise, the tabs can be taken as long in them as is desirable. This fact is used in the construction of the present analogs.

Two such tanks have been built and operated; a rectangular one for which the analytical solution is available, and a more complex one which represents typical alloy transistor geometry. These are shown in Figs. 3, 4, and 5; and are described in somewhat more detail in subsequent sections. Both are lucite tanks whose inside dimensions reproduce the outer dimensions of the semiconductor base region. Electrolyte solution in the tank provides the conducting medium. Surface electrodes on the plastic walls were made by an initial spray deposit of silver, which was then built up by copper electroplating and finally given a light goldplate to retard corrosion. The tabs were formed by scribing through the metal plate, and each tab was provided with an external electrical contact by means of a small pin through the plastic. Similar plated surfaces served as emitter and collector electrodes in the rectangular tank. The curved emitter and collector electrodes in the

TA-153 were in the form of detachable inserts, either of solid metal or coated plastic depending on their size.

### Measurement of Currents

Fig. 6 is a schematic diagram of the current measuring circuit used in conjunction with the tank analogs. A 10-kc signal of 2 to 10 volts was applied across $AB$, and $R_c$ and $R_s$ were each adjusted to give null readings against the equal-armed, pure resistive branch $R_1R_2$. The capacitors $C_c$ and $C_s$ were tuned to nullify the reactive components in each of the two tank paths, $EC$ and $ES$. At mutual balance the voltage drop across each path is the same, hence the transistor parameters of interest,

$$\alpha_{cb} = \frac{I_c}{I_s} = \frac{R_s}{R_c} \tag{34}$$

$$\alpha_{ce} = 1 - \frac{I_s}{I_s + I_c} = 1 - \frac{R_c}{R_c + R_s}, \tag{35}$$

are calculable directly from the bridge readings.

Since electrolyte solutions decrease in resistivity by about $2\frac{1}{2}$ per cent per degree C. rise at room temperature,[9] it is necessary either to maintain the tank at a fairly constant temperature or to record the solution temperature at balance and compute the correction. The latter procedure was followed in the present work, the correction being applied to the value of $s$ through (33).

### Rectangular Analog and Results

The rectangular tank, Fig. 3, was constructed for the purpose of gauging the adequacy of such analog measurements by comparison with values which can be computed analytically. The inside dimensions of the device are 10 inches by 2 inches by 2 inches. Emitter and collector are represented by the two plated, vertical faces; their effective size can be varied simply by adjusting the depth of electrolyte in the tank. The tank bottom, which represents the surface of recombination, consists of 20 tabs (0.1 inch wide) cut parallel to the emitter and collector. Each tab is in electrical contact with the insulated terminals of the bleeder resistance plug board on the right side of the device. The terminals on the left are connected to a common bus bar, which represents $S$ in Fig. 6. The illustration shows a set of bleeder resistances in place.

A comparison between measured $\alpha_{cb}$ and that calculated from (19) and (4) is shown in Fig. 2 for a 10-inch depth of electrolyte. These results were taken with two concentrations of solution having resistivities of 2,000 and 10,000 ohm cm, and with bleeder resistors of from 1 $K$ to 330 $K$ ohms. Agreement between the analog measurement and theory is quite satisfactory.

[9] See, for example, M. Dole, "Experimental and Theoretical Electrochemistry," McGraw-Hill Book Co., Inc., New York, N. Y., p. 101; 1935.

### TA-153 Analog

Fig. 4 shows the TA-153 analog in operating condition. As in the rectangular tank, the surface tabs are again led to the insulated terminals of the bleeder plug board on the side of the device and thence through the bleeder resistors to the common terminal bar below. The protruding central bolt serves both to hold the emitter (or collector) in place and to provide electrical contact to these electrodes. Thus it represents $E$ (or $C$) in Fig. 6. An opened view of the tank is provided by

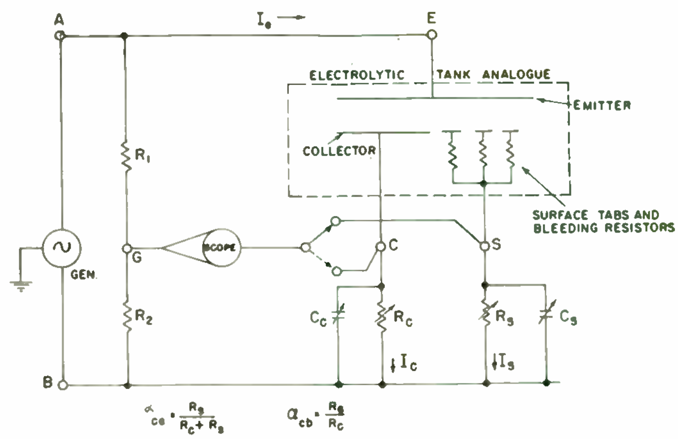



Fig. 6—Circuit diagram for measurement of $\alpha_{ce}$ and $\alpha_{cb}$ with the electrolytic tank analog.

Fig. 5. It shows the emitter and collector inserts in place and also the surface tabs, which could be taken here in the form of concentric rings in view of the cylindrical symmetry. Additional emitter and collector inserts of various curvatures are to be seen in the foreground.

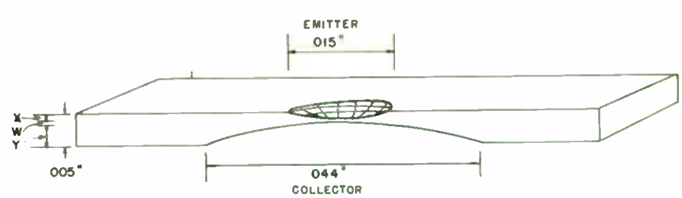



Fig. 7—Cross section through a typical alloy transistor; $X$ and $Y$ measure the maximum penetration of the emitter and collector respectively, while $W$ is the minimum separation.

The nominal dimensions of the TA-153 transistor, depicted in section in Fig. 7, are assumed to be: wafer thickness, 5 mils; emitter diameter, 15 mils; collector diameter, 44 mils; and the emitter and collector junction surfaces are idealized as spherical sections. The analog was constructed at a scale-up of 200:1; hence the inner width of the tank is 1 inch, and the emitter and collector inserts are 3 inches and 8.8 inches in diameter. The inner length and breadth of the tank were arbitrarily fixed at 16 inches. Measurements in Part I established that the carrier concentration at such distances from the cylindrical axis is quite small. For the same reason, recombination on the thin outer surfaces of the wafer was ignored, which means that these tank surfaces were not plated.

The surface tabs on the emitter side were taken as five rings of 0.1-inch width from radius 1.5 inches (emitter edge) to radius 2 inches, then ten rings of 0.2-inch width, and finally ten more of 0.4-inch width. On the collector side nine rings of 0.4-inch width were cut, starting at radius 4.4 inches (collector edge). The insulating gap produced by the scribing tool was 0.002 to 0.004 inch.

Previous work with the two-dimensional analog had demonstrated the importance of maintaining a large ratio of collector-to-emitter diameter. In the present work the geometrical variation studied most extensively was the emitter-and-collector penetration, i.e., the radii of curvature of the junction surfaces. $X$ and $Y$ denote the maximum emitter and collector penetration respectively, and $W$ the minimum emitter-collector separation, Fig. 7. With these conventions established, the device geometry can be specified by the sequence $X:W:Y$; for example, 1:2:2 denotes such a TA-153 structure with 1 mil maximum emitter penetration, 2 mils minimum gap, and 2 mils maximum collector penetration.

The analog was arranged to represent $n$-type germanium, with $D_p$ taken to be 44 cm²/sec. For convenience a standard set of bleeder resistors, $R^*$, was chosen, of such size that the magnitude of $s$ in cm/sec equals half the magnitude of the electrolyte resistivity in ohm-cm:

$$|s| = \left|\frac{\rho}{2}\right|. \tag{36}$$

By (33) this required the $R$ for each ring to satisfy

$$AR^* = 17{,}600 \text{ ohm-cm}^2, \tag{37}$$

$A$ being the ring area. Additional sets of bleeders were prepared so that

$$R_t = tR^*; \qquad t = 2, 5, 10, 20, 50, 100, 200. \tag{38}$$

With these,

$$|s| = \left|\frac{\rho}{2}\right| \times \frac{1}{t}. \tag{39}$$

Measurements were made using all these bleeders in conjunction with two concentrations of electrolyte having resistivities of 20,000 and 3,000 ohm-cm respectively. In this way the range of $s$ up to 10,000 cm/sec was surveyed. The results are presented and discussed in the following sections. They are not limited to $n$-type germanium devices, but can easily be applied to a material with minority carrier mobility $D$ cm²/sec by renormalizing $s$ as follows:

$$s = s(n - G_e) \times \frac{D}{44}. \tag{40}$$

## Results

The dependence of $\alpha_{ce}$ and $\alpha_{cb}$ on $s$ has been measured in this analog on fifteen geometrical variations of the TA-153, comprising all the compatible combinations of integral values of $X$ and $Y$ from 0 to 4 mils. Fig. 8 shows a plot of the experimentally determined values of $\alpha_{ce}$ against ln $s$ for the case, 1:2:2, in both the forward and the reverse directions. Fig. 9 shows ln $\alpha_{cb}$ against ln $s$ for the same structure. Similar plots are obtained in all instances.

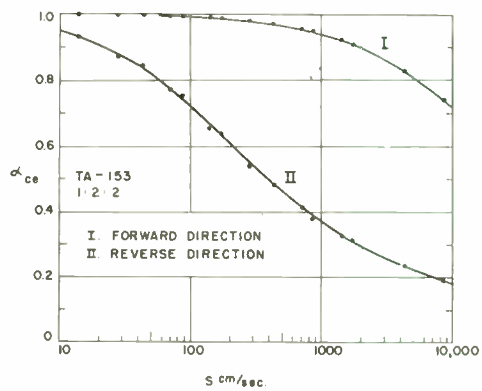



Fig. 8—$\alpha_{ce}$ as a function of $s$ for a typical alloy transistor with $X=1$ mil, $W=2$ mils, and $Y=2$ mils.

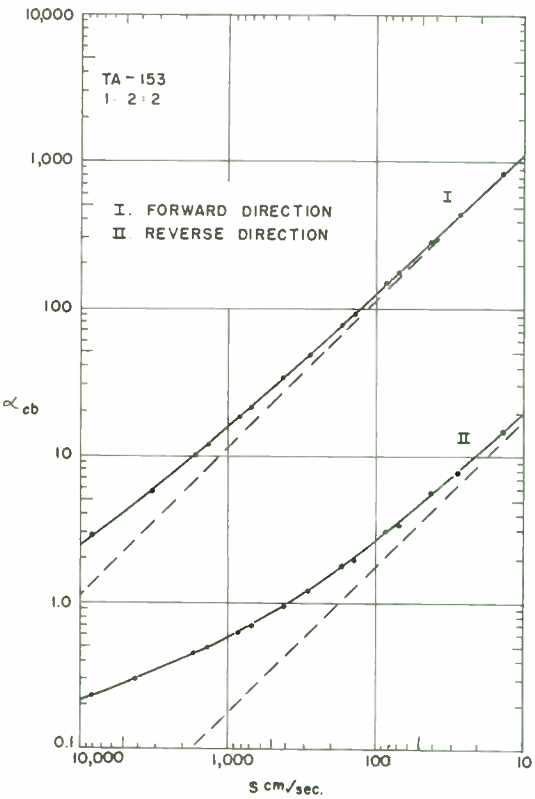



Fig. 9—$\alpha_{cb}$ as a function of $s$ for the same geometry as in Fig. 8. The dashed line shows the linear approximation $\alpha_{cb}=Ks^{-1}$.

Analysis of the results along the lines suggested by perturbation theory, (31), discloses that in this general geometry and up to the maximum value of $s$ investigated, 10,000 cm/sec, the dependence of $\alpha_{cb}$ on $s$ in the forward direction is very adequately represented by an equation of the form:

$$\alpha_{cb} = \frac{K}{s}\,[1 + b\sqrt{s}], \tag{41}$$

where $K$ and $b$ are geometrically dependent. The values of these parameters for the geometries investigated are shown in Table I. A typical comparison between the experimental $\alpha_{cb}$ and those computed from (41), for the case 1:2:2, is contained in Table II.

TABLE I

CURRENT GAIN IN TA-153; $X:W:Y$ (FORWARD)

$$\alpha_{cb} = \frac{K}{s}\,[1 + b\sqrt{s}]$$

$K$

| X \ W | 1 | 2 | 3 | 4 | 5 |
|---|---|---|---|---|---|
| 0 | 49,470 | 19,170 | 11,230 | 7,820 | 4,490 |
| 1 | 22,500 | 11,160 | 8,640 | 5,150 | |
| 2 | 14,900 | 9,410 | 5,580 | | |
| 3 | 10,730 | 6,230 | | | |
| 4 | 7,370 | | | | |

$b$

| X \ W | 1 | 2 | 3 | 4 | 5 |
|---|---|---|---|---|---|
| 0 | 0.01211 | 0.01140 | 0.01290 | 0.01722 | 0.02214 |
| 1 | 0.00939 | 0.01263 | 0.01443 | 0.01635 | |
| 2 | 0.01232 | 0.01430 | 0.01637 | | |
| 3 | 0.01566 | 0.01797 | | | |
| 4 | 0.02016 | | | | |

TABLE II

CURRENT GAIN IN TA-153; 1:2:2 (FORWARD)

$$\alpha_{cb} = \frac{11,160}{s}\,[1 + 0.01263\sqrt{s}]$$

| S | $\alpha_{cb}$ (measured) | $\alpha_{cb}$ (calculated) |
|---|---|---|
| 8,460 | 2.86 | 2.85 |
| 4,260 | 4.79 | 4.78 |
| 1,700 | 9.88 | 10.00 |
| 1,400 | 11.46 | 11.72 |
| 850 | 18.15 | 18.00 |
| 700 | 21.0 | 21.2 |
| 425 | 34.4 | 33.2 |
| 280 | 47.7 | 48.2 |
| 170 | 76.9 | 75.8 |
| 140 | 91.9 | 91.3 |
| 86 | 150 | 145 |
| 70 | 174 | 176 |
| 43 | 287 | 282 |
| 28 | 414 | 410 |
| 14 | 833 | 834 |
| 7 | 1,590 | 1,650 |

For given value of $s$, the bracketed quantity in (41) does not vary greatly from one geometry to another. Thus the geometrical dependence of $\alpha_{cb}$ is essentially contained in $K$; one can take this number as a geometrical figure of merit. For example, with a given emitter shape (given $X$) the size of $K$ increases with decreasing $W$, as is to be expected. In addition, however, $K$ is markedly affected by junction shape; and favorable junction geometry can compensate for considerable disadvantage in $W$. As a particular instance, 1:2:2 is about twice as good as 3:2:0, and is as good as 3:1:0. The general situation with regard to junction geometry can be summed up in the following rule: minority carrier loss through surface recombination in the TA-153 can be held down by keeping the emitter junction as flat as possible (consistent with good junction properties) and by closing the emitter-collector gap by collector penetration.

By means of (41) one can compute the maximum $s$ tolerable for a given $\alpha_{cb}$, when volume recombination is negligible. Table III shows the geometrical dependence of this value of $s$, for the particular case $\alpha_{cb} > 30$.

TABLE III

MAXIMUM TOLERABLE $s$ IN TA-153; $X:W:Y$ (FOR)

$\alpha_{cb} \geqq 30$

| X \ W | 1 | 2 | 3 | 4 | 5 |
|---|---|---|---|---|---|
| 0 | 2,700 | 850 | 480 | 340 | 200 |
| 1 | 970 | 470 | 370 | 210 | |
| 2 | 650 | 400 | 230 | | |
| 3 | 480 | 270 | | | |
| 4 | 340 | | | | |

In the reverse direction, $\alpha_{cb}$ is similarly expressible as

$$\alpha_{cb} = \frac{K}{s}\,[1 + bs^{0.7}]. \tag{42}$$

Table IV contains the values of the observed constants, and Table V (next page) gives a comparison between observed and computed quantities.

TABLE IV

CURRENT GAIN IN TA-153; $X:W:Y$ (REVERSE)

$$\alpha_{cb} = \frac{k}{s}\,[1 + bs^{0.7}]$$

$K$

| X \ W | 1 | 2 | 3 | 4 | 5 |
|---|---|---|---|---|---|
| 0 | 320 | 200 | 145 | 130 | 120 |
| 1 | 295 | 180 | 165 | 135 | |
| 2 | 245 | 200 | 150 | | |
| 3 | 235 | 200 | | | |
| 4 | 235 | | | | |

$b$

| X \ W | 1 | 2 | 3 | 4 | 5 |
|---|---|---|---|---|---|
| 0 | 0.0163 | 0.0139 | 0.0172 | 0.0159 | 0.0175 |
| 1 | 0.0139 | 0.0178 | 0.0154 | 0.0169 | |
| 2 | 0.0176 | 0.0159 | 0.0179 | | |
| 3 | 0.0181 | 0.0164 | | | |
| 4 | 0.0203 | | | | |

TABLE V

CURRENT GAIN IN TA-153; 1:2:2 (REVERSE)

$$\alpha_{cb} = \frac{180}{s}\left[1 + 0.0178\, s^{0.7}\right]$$

| $S$ | $\alpha_{cb}$ (measured) | $\alpha_{cb}$ (calculated) |
|---|---|---|
| 8,460 | 0.235 | 0.233 |
| 4,260 | 0.307 | 0.303 |
| 1,700 | 0.460 | 0.451 |
| 1,400 | 0.484 | 0.494 |
| 850 | 0.609 | 0.638 |
| 700 | 0.686 | 0.708 |
| 425 | 0.928 | 0.953 |
| 280 | 1.17 | 1.24 |
| 170 | 1.75 | 1.74 |
| 140 | 1.91 | 2.01 |
| 86 | 3.07 | 2.94 |
| 70 | 3.32 | 3.47 |
| 43 | 5.61 | 5.24 |
| 28 | 7.61 | 7.61 |
| 14 | 14.7 | 14.5 |
| 7 | 28.8 | 27.5 |

## INTERPRETATION OF GEOMETRICAL DEPENDENCE

From (29) and (30) it may be seen that, for the simple geometries discussed in that section,

$$K \propto \frac{1}{W^2}. \tag{43}$$

In the TA-153 the relation between $K$ and $W$ is more complicated. The perturbation treatment, however, in pointing up the geometrical source of $K$, sheds considerable light on the observed results in Table I.

In the simple geometry the inverse square dependence in (43) arises from two factors. First, the collector current depends on the hole concentration gradient between emitter and collector, and this varies as $W^{-1}$, (27). Second, the surface current depends on the effective free surface area, and this varies with $W$, (28). These factors cannot be changed independently when the cross section is fixed, thus yielding the inverse square dependence of (43).

In the TA-153, $K$ similarly depends on these currents. Here, however, the geometry is of such a nature that considerable independent variation is possible. The collector current depends on an average hole gradient between emitter and collector, and this can be expected to change in some inverse manner with $W$, as in Fig. 10. The relation of surface current to $W$, however, is not direct. As established in Part I, the surface current is confined essentially to a small, effective region of the free surface around the emitter. This region is indicated by $S$ in Fig. 10. The extent of $S$ is determined by competition between the surface and the collector, $C$, for the holes injected near the outer edge of the emitter, $E$. Thus, the surface current is expected to depend largely upon the nearness of $C$ to $S$, and this can change or not with $W$. For example, if $W$ is reduced by moving the emitter to $E'$, essentially only the collector current is changed. Hence in such cases $K$ should vary in some inverse manner with $W$. On the other hand, if $W$ is reduced to a similar extent by changing the collector to $C'$, not only is the collector current changed but also the effective surface, i.e., the nearness of $C$ to $s$. In this case, therefore, a higher order dependence of $K$ on $W$ is expected.

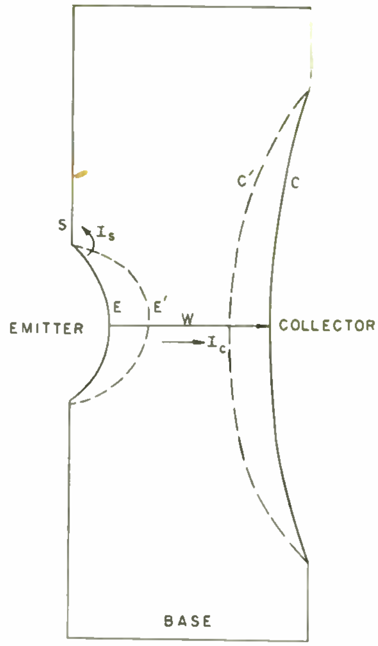



Fig. 10—Cross section through an alloy transistor showing how possible variations in geometry can change the surface currents and $W$ independently.

The values of $K$ given in Table I are plotted vs $W$ in Fig. 11 to show the observed geometrical dependence of this figure of merit in the TA-153. The results are compared there with a grid representing

$$K = \frac{35 \times 10^4}{(W+6)\left[1.35(W+X) - 0.35\right]}, \text{ and } W+X = 5-Y. \tag{44}$$

Now the use of this expression is intended merely to be suggestive, the actual relation between $K$ and geometry being surely more complex. In particular (44) should not be used loosely for any extended extrapolation to smaller $W$. Nevertheless, it is a simple function whose behavior shows good qualitative agreement with the observed results, and one which gives a more mathematical statement to the line of argument presented above. When $W$ is varied with $Y$ constant on this grid, the second term in the denominator of (44) is unchanged, and $K$ varies as $(W+6)^{-1}$. Alternatively, when $W$ is varied with $X$ constant, both terms in the denominator change, and $K$ varies inversely as a quadratic in $W$. Thus the two factors in the denominator simulate respectively the influence of geometry on the collector current and on the effective surface area, $S$.

## VOLUME RECOMBINATION IN TA-153

The analog measurements of the minority carrier loss in the TA-153 apply strictly to a base material of bulk lifetime sufficiently long so that surface recombination completely over-shadows volume recombination. It shall now be shown how one can make a rough estimate of the requirement this puts on $\tau$. For this purpose it is convenient to use the approximate expression for $\alpha_{ce}$, (20). The size of the surface term is known from Table I. The volume term can, in fair approximation, be taken in the form of the volume term in (29), if a suitable choice of an equivalent cross section and emitter-collector spacing is inserted into (26) and (27). For simplicity, take the actual cross-sectional area of the emitter and the minimum spacing $W$. Then,

$$\alpha_{ce} = 1 - \frac{s}{K} - \frac{W^2}{2D\tau}; \tag{45}$$

or, explicitly for $n$-type germanium,

$$\alpha_{ce} = 1 - \frac{s}{K} - \frac{0.075W^2}{\tau}, \tag{46}$$

with $W$ in mils and $\tau$ in $\mu$sec.

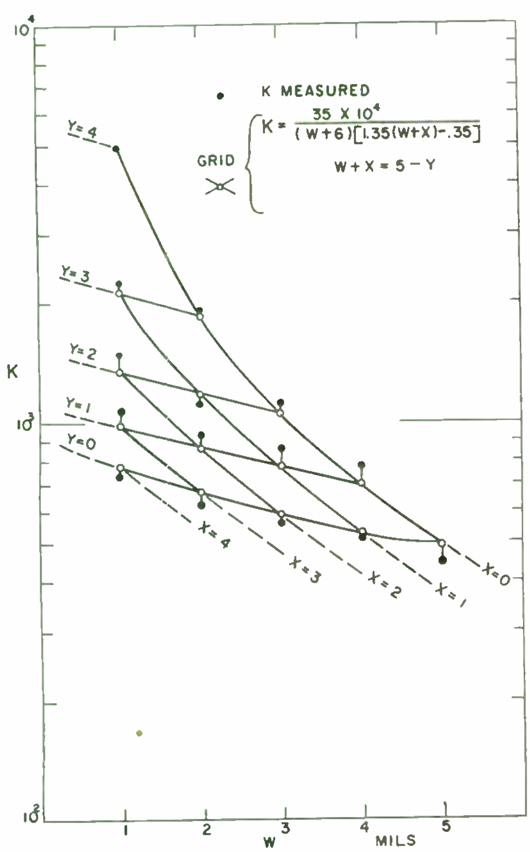



Fig. 11—The dependence of $K$ on $W$ with $X$ or $Y$ held fixed.

The use of a $W^2$ term for the volume recombination in the TA-153 can be justified, in essence, by arguments similar to those used in the previous section. The square of $W$ in the volume term represents again the product of two factors: an increase in $W$ both decreases the velocity of diffusion (decreases the gradient) and increases the path length. These factors, which determine the transit time, vary simultaneously both in the simple geometry and in the TA-153.

As an example of such an estimate of the relative importance of surface and volume terms, consider TA-153; 1:2:2 with $\tau = 100$ $\mu$sec. Insertion of the proper figures into (46) shows the surface loss to be about 10 times the volume loss when $s = 330$, but to be about equal to the volume loss when $s = 33$. Since in most cases practical values of $s$ are in the range of a few hundred to one thousand, the assumption of negligible volume loss is a good one.

## Appendix A

Consider the grown (flat) junction type with emitter-collector separation $W$ and circular cross section of radius $R$. A cylindrical co-ordinate system is used, with the $z$-axis perpendicular to the junction faces; and for convenience reduced units with $R$ as the unit length are introduced; abbreviating:

$$g = \frac{W}{R} \qquad h = \frac{Rs}{D} \tag{47}$$

Then (5)–(8) become

$$\frac{1}{r}\frac{\partial}{\partial r}\left(r\frac{\partial P}{\partial r}\right) + \frac{\partial^2 P}{\partial Z^2} - \frac{R^2 P}{D\tau} = 0 \tag{48}$$

$$\frac{\partial P}{\partial r} = -hP; \qquad r = 1, \qquad 0 < Z < g \tag{49}$$

$$P = P_e; \qquad r \leqq 1, Z = 0 \tag{50}$$

$$P = 0; \qquad r \leqq 1, Z = g. \tag{51}$$

This set of equations can now be solved by any of the standard methods; here it was chosen to apply transform techniques. By means of the Hankel Transform,[10]

$$\overline{P}(n, Z) = \int_0^1 P(r, Z) r J_0(nr) dr, \tag{52}$$

defined in terms of the zero order Bessel function $J_0$, (48) is converted to

$$\frac{\partial^2 \overline{P}}{\partial Z^2} - \left(n^2 + \frac{R^2}{D\tau}\right)\overline{P} = 0; \tag{53}$$

provided that by (49), $n$ is chosen such that

$$nJ_0'(n) + hJ_0(n) = 0. \tag{54}$$

Conditions (50) and (51) become

$$\overline{P}(n, 0) = P_e \frac{J_1(n)}{n} \tag{55}$$

$$\overline{P}(n, Z) = 0. \tag{56}$$

The solution of (53) which satisfies (55) and (56) is readily found to be

$$\overline{P}(n, Z) = P_e \frac{J_1(n) \sinh\left[(g - Z)\sqrt{n^2 + R^2/D_\tau}\right]}{\sinh\left[g\sqrt{n2 + R^2/D_\tau}\right]}. \tag{57}$$

[10] See, for example, C. J. Tranter, "Integral Transforms in Mathematical Physics, Methuen, Ltd., London, Eng., p. 88; 1951.

Then inversion[11] gives the required hole density:

$$P(r, Z) = 2 \sum_n \frac{h}{n^2 + h^2} \frac{nJ_0(nr)}{J_0(n)J_1(n)} \overline{P}(n, Z), \quad (58)$$

where the sum is over the successive positive roots of (54).

The currents required by (9) and (10) are

$$I_e = 2\pi \int_0^1 r \left.\frac{\partial P}{\partial Z}\right|_{Z=0} dr \quad (59)$$

and

$$I_c = 2\pi \int_0^1 r \left.\frac{\partial P}{\partial Z}\right|_{Z=g} dr, \quad (60)$$

and when these operations are performed on (58) and the results substituted into (3), (13) results.

## Appendix B

Consider the grown (flat) junction type with emitter-collector separation $W$ and rectangular cross section of semi-dimensions $a$ and $b$. A rectangular co-ordinate system with the $z$-axis perpendicular to the junction faces is used. Then (5)–(8) become

$$\left(\frac{\partial^2}{\partial x^2} + \frac{\partial^2}{\partial y^2} + \frac{\partial^2}{\partial z^2}\right) P - \frac{P}{D\tau} = 0 \quad (61)$$

$$\frac{\partial P}{\partial x} = \frac{-s}{D} P; \quad x = a, 0 < y < b, 0 < Z < W$$

$$\frac{\partial P}{\partial x} = 0; \quad x = 0, 0 < y < b, 0 < Z < W \quad (62)$$

$$\frac{\partial P}{\partial y} = -\frac{s}{D} P; \quad 0 < x < a, y = b, 0 < Z < W$$

$$\frac{\partial P}{\partial y} = -\frac{s}{D} P; \quad 0 < x < a, y = 0; \quad 0 < Z < W \quad (63)$$

$$P = P_e; \quad 0 < x < a, 0 < y < b, Z = 0 \quad (64)$$

$$P = 0; \quad 0 < x < a, 0 < y < b, Z = W. \quad (65)$$

By means of the double Cosine Transform[12]

$$\overline{\overline{P}}(n, m, Z) = \int_0^a dx \int_0^b dy P(x, y, Z) \cos nx \cos my. \quad (66)$$

Eq. 61 is converted to

$$\frac{\partial^2 \overline{\overline{P}}}{\partial Z^2} - \left(n^2 + m^2 + \frac{1}{D\tau}\right) P = 0, \quad (67)$$

provided that by (62) and (63) $n$ and $m$ are chosen such that

$$n \tan na = \frac{s}{D}$$
$$m \tan mb = \frac{s}{D}. \quad (68)$$

Conditions (64) and (65) become

$$\overline{\overline{P}}(n, m, 0) = P_e \frac{\sin na}{n} \frac{\sin ma}{m} \quad (69)$$

$$\overline{\overline{P}}(n, m, W) = 0. \quad (70)$$

The solution of (67) which satisfies (69) and (70) is

$$\overline{\overline{P}}(n, m, Z) = P_e \frac{\sin na}{n} \frac{\sin mb}{m} \times \frac{\sinh\, [W - Z]\sqrt{n^2 + m^2 + 1/D\tau}}{\sinh\, [W\sqrt{n^2 + m^2 + 1/D\tau}}. \quad (71)$$

Then inversion gives[13]

$$P(x, y, Z) = 4 \sum_n \sum_m \frac{n}{na + (s \cos^2 na/nD} \times \frac{m}{mb + (s \cos^2 mb/mD} \cos nx \cos my \overline{\overline{P}}(n, m, Z) \quad (72)$$

where sums are over the successive positive roots of (68).

The currents required by (9) and (10) are

$$I_e = \int_0^a dx \int_0^b dy \left.\frac{\partial P}{\partial Z}\right|_{Z=0} \quad (73)$$

and

$$I_c = \int_0^a dx \int_0^b dy \left.\frac{\partial P}{\partial Z}\right|_{Z=W}, \quad (74)$$

and when these operations are performed on (72) and the results substituted into (3), (16) results.

When $s \to 0$ on the surface $y = b$, one need only consider the first term in the $m$ series. As $s \to 0$, this first value of $m$ behaves as

$$m = \sqrt{\frac{s}{bD}}. \quad (75)$$

Then, allowing $\tau \to \infty$ as well, (72) reduces to

$$P(x, y, Z) = 2 P_e \sum_n \frac{\sin na \cos nx}{na + s \cos^2 na} \frac{\sinh\, [n(W-Z)]}{\sinh\, [nW]}, \quad (76)$$

and (19) is obtained by using this hole density in (37) and (74) and thence in (3).

[11] *Ibid.*, p. 89.
[12] *Ibid.*, p. 86.

[13] This solution is closely allied to one for a sample with square cross section and infinite length given by W. Shockley, *Bell. Sys. Tech. Jour.*, vol. 28, pp. 480–483; July, 1949.

# Further Analysis of Transmission-Line Directional Couplers*

R. C. KNECHTLI†

*Summary*—**The conditions of infinite directivity for transmission-line directional couplers are derived in a general and rigorous way. Exact expressions valid for any degree of coupling are found.**

**The case of small coupling and the case of a matched primary line are considered as particular cases of this general analysis. In the case of small coupling, the condition $Z_{11}Z_{44}=1$ and $Z_{22}Z_{33}=1$ given by W. L. Firestone in a recent paper are found to be correct; it is shown also that these simple conditions do not apply anymore when the coupling becomes large.**

IN HIS PAPER,[1] Firestone derived a condition of infinite directivity for transmission-line directional couplers. He expressed this condition in the notation defined further on as follows:

$$Z_{22}Z_{33} = 1 \quad \text{and} \quad Z_{11}Z_{44} = 1. \tag{1}$$

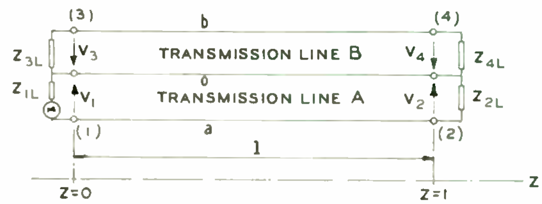
It can be shown that conditions (1) are not the most general conditions for infinite directivity, but rather they apply only for the special case of weak coupling. The general conditions for infinite directivity will be derived here.

A rigorous analysis, based directly on Maxwell's equations, of a system of $n$-parallel cylindrical conductors of arbitrary cross section, was published by the author (in joint authorship) several years ago.[2] As an application of this analysis, the case of a system of three conductors was investigated. (Such a system may consist of two coupled single wire transmission lines above a common ground or within a common cylindrical ground conductor.) It was found at that time (as Firestone has shown by other methods in his paper) that such a system has the properties of a directional coupler, if properly terminated. We shall base our present derivations upon this analysis.

While Firestone first analyzes a system of four parallel conductors and treats the system of three conductors as a particular case of the system of four conductors, for the sake of simplicity we shall limit our present discussion to the system of three conductors; in doing so we may bear in mind that a similar discussion is also possible for a system of four conductors. Further, let us refer from now on to Firestone's paper as (I) and to our paper as (II).

Let us consider the system shown in Fig. 1 with the notation indicated on this figure (a practical design of such a system for vhf or uhf is shown in Fig. 2). This system may be considered as the junction of four transmission lines coupled over the length $l$ by the two transmission lines $A$ and $B$, the lines $A$ and $B$ consisting respectively of the conductors $a-0$ and $b-0$.

Let $z$ be the co-ordinate in the direction parallel to the transmission lines $A$ and $B$. Let be $z=0$ at the terminals (1) and (3) and $z=l$ at the terminals (2) and (4), the lines $A$ and $B$ being coupled only in the interval $(0, l)$.




$l$ = Distance between terminals (1)–(2) and (3)–(4)
$Z_{L1}$ to $Z_{L4}$ = Impedances seen at terminals (1) to (4)
Conductors $(a)$ and $(o)$ = Transmission line $A$
Conductors $(b)$ and $(o)$ = Transmission line $B$
Conductor $(o)$ = common ground or common cylindrical envelope for lines $A$ and $B$
$C_{ao}$ = Capacity per unit length between conductors $(a)$ and $(o)$
$C_{bo}$ = Capacity per unit length between conductors $(b)$ and $(o)$
$C_{ab}$ = Capacity per unit length between conductors $(a)$ and $(b)$
($C_{ab}$ = coupling capacity)

Fig. 1—Coupled transmission lines.

Considering the propagation of TEM waves only, we have at any given plane perpendicular to the conductors ($z$ = constant) uniquely defined voltages $V_a$ on line $A$ between the conductors $(a)-(0)$, and $V_b$ on line $B$ between the conductors $(b)-(0)$.

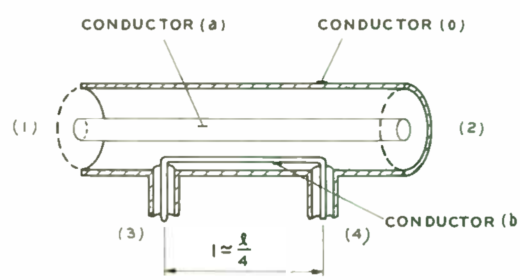



Fig. 2—Transmission-line directional coupler for uhf.

According to (II) [matrix equation (15)] $V_a$ and $V_b$ can be expressed as a sum of forward and backward traveling waves:

$$\left.\begin{aligned} V_a &= V_a' e^{j(\omega t-\beta z)} + V_a'' e^{j(\omega t+\beta z)} \\ V_b &= V_b' e^{j(\omega t-\beta z)} + V_b'' e^{j(\omega t+\beta z)} \end{aligned}\right\}, \tag{2}$$

where

$$V_a', V_a'', V_b' \quad \text{and} \quad V_b''$$

are constants and can be complex.

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[1] W. L. Firestone, "Analysis of transmission-line directional couplers," Proc. IRE. vol. 42, pp. 1529–1538; October, 1954.
[2] H. J. von Baeyer and R. C. Knechtli, "Ueber die Behandlung von Mehrleitersystemen mit TEM Wellen bei hohen Frequenzen," *Zeit. für Angew. Math. und Phys.*, vol. 3, pp. 271–286; 1952.

Let us define the complex reflection coefficients $\Gamma_{a1}$ at terminals (1) of line $A$ and $\Gamma_{b3}$ at terminals (3) of line $B$ as follows:

$$\left.\begin{aligned}\Gamma_{a1} &= \frac{V_a''}{V_a'}\\ \Gamma_{b3} &= \frac{V_b''}{V_b'}\end{aligned}\right\}. \quad (3)$$

According to (2), (3) and with the notation of Fig. 1, the voltages $V_3$ and $V_4$ at the terminals (3) and (4) may be expressed as follows:

$$V_3 = (V_b' + V_b'')\cdot e^{j\omega t} = (1 + \Gamma_{b3})V_b'e^{j\omega t}$$

$$V_4 = (V_b'e^{-j\beta l} + V_b''e^{+j\beta l})e^{j\omega t} = (e^{-j\beta l} + \Gamma_{b3}e^{+j\beta l})V_b'e^{j\omega t}.$$

Let us define the directivity $D_1$ of the system, when it is excited at terminals (1), by:

$$D_1 = 20 \log \left|\frac{V_3}{V_4}\right|. \quad (4)$$

The expressions for $V_3$ and $V_4$ show that, according to (4)

$$D_1 = 20 \log \left|\frac{1 + \Gamma_{b3}}{e^{-j\beta l} + \Gamma_{b3}e^{+j\beta l}}\right|. \quad (5)$$

This last relation yields for the condition of infinite directivity:

$$\Gamma_{b3} = -e^{-2j\beta l}. \quad (6)$$

We can derive from (18) of (II) a general and rigorous expression for $\Gamma_{b3}$. In order to do so let the following quantities be defined in agreement with the notation of Fig. 1:

$$Z_{a0} = \frac{\sqrt{\epsilon\mu}}{C_{0a} + C_{ab}} = \text{characteristic impedance of line } A$$

$$Z_{0b} = \frac{\sqrt{\epsilon\mu}}{C_{b0} + C_{ab}} = \text{characteristic impedance of line } B$$

(by introducing the coefficient $1/\sqrt{\epsilon\mu} = v =$ phase velocity of the TEM waves, the inductivities per unit length of the conductors considered are eliminated from our expression).

Let us further define, according to the notation of Fig. 1 and in agreement with the notation used by Firestone:

$Z_{1L}$ = impedance seen at terminals (1)

$Z_{2L}$ = impedance seen at terminals (2)

$Z_{3L}$ = impedance seen at terminals (3)

$Z_{4L}$ = impedance seen at terminals (4)

$$Z_{11} = \frac{Z_{1L}}{Z_{0a}} \qquad Z_{22} = \frac{Z_{2L}}{Z_{0a}}$$

$$Z_{33} = \frac{Z_{3L}}{Z_{0b}} \qquad Z_{44} = \frac{Z_{4L}}{Z_{0b}}.$$

Using the notation just defined we obtain from (18) of (II) the following expression for $\Gamma_{b3}$:

$$\Gamma_{b3} = \frac{(1+Z_{33})(\Gamma_{a1}e^{j\beta l} - e^{-j\beta l})Z_{44} - (1-Z_{44})(1-\Gamma_{a1})Z_{33}e^{-j\beta l}}{(1+Z_{44})(1-\Gamma_{a1})Z_{33}e^{j\beta l} - (1-Z_{33})(\Gamma_{a1}e^{j\beta l} - e^{-j\beta l})Z_{44}}. \quad (7)$$

(It will be noticed that when $Z_{33} = Z_{44}$, (7) becomes identical with (22) of (II), if the proper change of notation is performed. This had also to be expected.)

Introducing (7) into (6) one obtains for the condition of infinite directivity the following expression:

$$\Gamma_{a_1}e^{2j\beta l} = -\frac{Z_{33} - 1}{Z_{33} + 1}. \quad (8)$$

For some applications, it may be more convenient to express the condition of infinite directivity in terms of impedances rather than of reflection coefficients. For this purpose we express $\Gamma_{a1}$ as a function of the normalized terminating impedances $Z_{11}$ to $Z_{44}$, and of a coefficient of coupling $K$ which will be defined.

For $\Gamma_{a1}$ we have, from (28) of (II):

$$\Gamma_{a1} = \frac{\Gamma_{10} - K^2 \dfrac{Z_{22}}{1 + Z_{22}} e^{-2j\beta l}}{1 - \dfrac{K^2}{2}\cdot\dfrac{Z_{22}}{1 + Z_{22}}(1 + e^{-2j\beta l})} \quad (9)$$

with

$$\Gamma_{10} = \frac{Z_{22} - 1}{Z_{22} + 1} e^{-2j\beta l} \quad (9a)$$

$$K = \frac{C_{ab}}{\sqrt{(C_{a0} + C_{ab})(C_{b0} + C_{ab})}} \leq 1. \quad (9b)$$

The reflection coefficient $\Gamma_{10}$ would exist on line $A$ at terminal (1) in the case of zero-coupling ($C_{ab} \to 0$).

By means of (8) and (9) the condition of infinite directivity can finally be expressed as follows:

$$Z_{33} = \frac{\dfrac{1}{Z_{22}} + \dfrac{K^2}{4}(1 - e^{-2j\beta l})}{1 - \dfrac{K^2}{4}(3 + e^{-2j\beta l})}. \quad (8a)$$

Eq. (8a) is equivalent to (8).

If we excite the system of Fig. 1 at the terminals (2) and define the directivity $D_2$ by:

$$D_2 = 20 \log \left|\frac{V_4}{V_3}\right|$$

the condition for infinite directivity $D_2$ is found, in analogy to (8a) to be:

$$Z_{44} = \frac{\dfrac{1}{Z_{11}} + \dfrac{K^2}{4}(1 - e^{-2j\beta l})}{1 - \dfrac{K^2}{4}(3 + e^{-2j\beta l})}. \quad (8b)$$

The relations (8a) and (8b) show that infinite directivity can be obtained when the primary line (line $A$ in our case) is mismatched as well as when it is matched. (We call line $A$ "matched" when $\Gamma_{a1}=0$.) Having found the general condition for infinite directivity [(8a) and (8b)] let us consider some particular cases of practical interest.

### Case of Small Coupling

Let us define "small coupling" by the condition:

$$K^2 \ll 1. \tag{10}$$

In this case, the conditions of infinite directivity given by (8a) and (8b) become:

$$\left.\begin{array}{l}\lim Z_{22}Z_{33} = 1 \\ \lim Z_{11}Z_{44} = 1\end{array}\right\} \quad \text{for} \quad K^2 \to 0. \tag{11}$$

Comparing (11) (which has been rigorously derived) with the conditions $Z_{22}Z_{33}=1$ and $Z_{11}Z_{44}=1$ given by Firestone for infinite directivity, we see that these last conditions hold only in the case of weak coupling. Although the case of weak coupling is of considerable practical importance [e.g. when a transmission-line directional coupler is used for the measurement of swr, in the way suggested in (II)], cases of strong coupling may also be important; in such cases, the conditions (11) for infinite directivity are not valid any more, and (8a) and (8b) have to be used instead.

### Case of Matched Lines

Let us determine under which conditions the reflection coefficient on line $A$ is zero and the directivity of the system is infinite.

From (9) we find the condition for $\Gamma_{a1}=0$ to be:

$$Z_{22} = \frac{1}{1-K^2}. \tag{12}$$

From (8) we find the condition of infinite directivity ($D_1=\infty$) when $\Gamma_{a1}=0$ to be:

$$Z_{33} = 1. \tag{13}$$

One notices in (12) how the coupling $K$ affects the matching condition of line $A$.

As a conclusion of this analysis we may summarize our results as follows:

1. Infinite directivity may be obtained with transmission-line directional couplers for mismatched as well as for matched lines. The condition of infinite directivity in its general form is given by the relations (8a) and (8b).
2. In the case of weak coupling, the conditions of infinite directivity reduce to the conditions $Z_{22}Z_{33}=1$ and $Z_{11}Z_{44}=1$.
3. In the case of matched lines and infinite directivity, ($\Gamma_{a1}=0$ and $D_1=\infty$) the matching impedance of the primary line is a function of the coupling coefficient [cf. (13a)].

# Phase Stabilization of Microwave Oscillators*

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**Summary—A circuit has been developed with which microwave oscillators may be phase-locked to weak but stable reference signals. The circuit was operated with S-band oscillators (707B klystron; 2C37 triode oscillator) and a 2K50 K-band klystron. It is possible to lock a microwave oscillator directly or through a cascade of such circuits to a quartz-stabilized oscillator. The statistical theory of random noise is used to obtain an analysis of the stabilizing effect of the circuit, and the power spectrum of the stabilized microwave source is calculated. The scheme can also be applied in divider operation. Modifications are discussed. A modified circuit that uses carrier-suppressed modulation of the reference signal has also been realized. In another circuit, the oscillator frequency is converted by means of a stable reference, and compared with a second reference that can be of low frequency and tunable. These latter circuits allow elimination of the excess noise introduced by crystal diodes. In the original straight dc circuit this noise cannot be eliminated, but calculation shows that its influence on the output power spectrum is very small.**

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### Introduction

THIS PAPER will discuss the phase stabilization of microwave oscillators. It should be clearly understood at the outset that phase stabilization is quite distinct from frequency stabilization in the conventional form. A frequency discriminator with a $(f_0-f)^{-1}$ control circuit would give essentially a phase-stabilization type of control, but such a $(f_0-f)^{-1}$ control is neither realizable physically nor defined analytically for the operating region, i.e., for $f=f_0$. If instead of a frequency discriminator a phase discriminator is used at the outset, all necessary components are realizable.

Note also that frequency stabilization allows one to establish a frequency to an accuracy which is constant with time. Phase stabilization establishes a mean frequency with an accuracy directly proportional to the locking time. The interest then in phase stabilizing microwave oscillators is to realize the transference of frequency stability from one frequency region to another with any desired precision.

Recent developments in the techniques of molecular beam measurements and microwave spectroscopy[1] make it possible to observe substances in a state where they absorb electromagnetic energy at one or several extremely sharply defined microwave frequencies. A substance in such a state is therefore analogous to a cavity of very high $Q$ ($10^7$ or better) with a persistently accurate resonance frequency. In order to measure the center frequency of one of these resonances it is desirable to have microwave oscillators whose output power is as monochromatic as possible.

At lower frequencies, an oscillator controlled by a quartz crystal may be used to generate a signal with very high stability for a period of hours or days. This low frequency can be multiplied by means of vacuum tube or silicon diode multipliers. Conventional multiplication usually yields a high-frequency spectrum that is not monochromatic but has sidebands, arising from lower frequency modulations, that remain because of the finite selectivity of the circuits. Furthermore, since multipliers with a gain of less than one (silicon diodes, for example) introduce additional noise into the spectrum, it is not desirable to multiply a frequency by more than a factor of 10 in these diodes.

In this paper we describe a stabilization circuit that allows a microwave oscillator to be locked to a harmonic of a stable reference oscillator. Through iteration of this process, the stability of a quartz-controlled oscillator can, essentially, be transferred to a K-band oscillator (23,040 mc). Description of the experiment is followed by an analysis of the stability and performance of a phase-locking circuit.

## Phase Stabilization of a K-Band Oscillator

The circuit that has been successfully used to stabilize the frequency of a 2K50 klystron is shown in Fig. 1. The circuit consists, essentially, of a single, absolutely stable feedback loop. Any phase modulation in the klystron is detected in the phase discriminator: the resulting signal is amplified in the differential dc amplifier and applied to the repeller of the klystron to produce an opposite and stabilizing phase modulation.

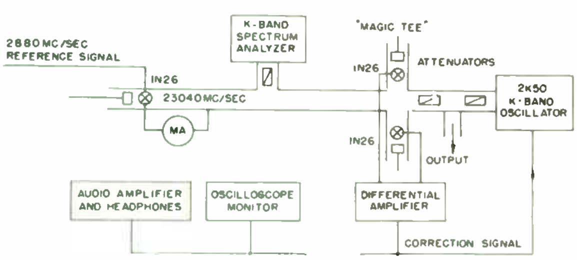


Fig. 1—Block diagram of phase-stabilizing circuit for K-band oscillator.

### *Phase Discriminator*

The reference signal and oscillator are introduced through the noncoupling arms of a waveguide hybrid junction, or "magic tee." The signals that arrive at the detector crystals on the two remaining arms, 1 and 2, $\epsilon_{S1}\sin(\omega_c t+\eta_1)$ and $\epsilon_{S2}\sin(\omega_c t+\eta_2)$ are shown by the vector diagram in Fig. 2.

These fields are the sum of a signal coming from the reference source, $\epsilon_{R1}\sin\omega_c t$ and $\epsilon_{R2}\sin\omega_c t$, and a signal coming from the oscillator, $\epsilon_{01}(\sin\omega_c t+\phi_1)$ and $\epsilon_{02}(\sin\omega_c t+\phi_2)$.

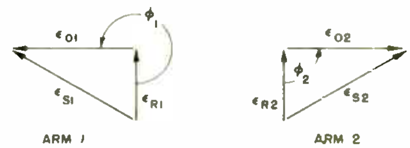


Fig. 2—Vector diagram of signals in hybrid junction.

From symmetry properties of the magic tee it is seen that $\phi_1=\phi_2+\pi$. Since detected power $P_D$ in a silicon diode as a function of input power $P_i$ is given by[2]

$$P_D = SP_i^2 \quad (S = \text{conversion gain per watt} \approx 10^4\ (\text{watts})^{-1})$$

it follows that the detected output signal is:

$$V_j = -2\sqrt{RSP_{0j}P_{Rj}}\cos\phi_j + \sqrt{RS}(P_{0j}+P_{Rj}),\ j = 1, 2.$$

Here, $P_{0j}$, $P_{Rj}$ are the power of oscillator and reference signal in each arm; $V_j$ is the phase discriminator detected output of each of the crystals. It can be seen that if $P_0$ and $P_R$ are divided equally between arms 1 and 2, then $V_1-V_2$ is independent of variations of $P_0$ and $P_R$ for $\phi_j\approx\pi/2$. Amplitude modulation of the two signals is therefore proportional to the balanced-out control signal, and hence second order, being negligible in the limit of small control signal. Since the insensitivity to amplitude variation allows discriminator to be operated at a high power level in spite of the small $P_R$, crystals can be operated in a region of good conversion gain. For $P_0=200\ \mu$ watts, $R=100$ ohms, $P_R=8\ \mu$ watts, we expect a differential output of $g_p=0.04$ volt per radian.

As indicated in Fig. 1, this output is amplified in a differential dc amplifier. A cross-coupled circuit[3] was used for this purpose; the actual circuit is shown in Fig. 3. The circuit uses all readily available techniques to achieve stable, hum-free operation. The heaters of the 12AX7 tubes are fed in series from the negative power supply. A K-band spectrum analyzer, an oscilloscope, and headphones are used to tune the oscillator to the reference signal. Once a klystron is within about 1,000 cps of the reference, it will phase-lock itself automatically. Since the 2K50 klystron is microphonic, good sound isolation is essential. The experiment was, therefore, carried out in an anechoic chamber. But, any good acoustic isolation for the klystron should be sufficient.

The reference signal was supplied from a very stable, cavity-tuned, planar triode S-band oscillator. This oscillator, in turn, was locked by an analogous circuit to the tenth harmonic of the output of the M.I.T. frequency standard.[4] The correction signal was applied to the plate of the oscillator triode.

For the S-band oscillator, a klystron also could be

[1] M. W. P. Strandberg and H. Dreicer, "Doppler line-width reduction," *Phys. Rev.*, vol. 94, pp. 1393–1394; June, 1954.

[2] M. W. P. Strandberg, "Microwave Spectroscopy," Methuen and Co., London, Eng.; 1954.

[3] J. N. van Scoyoc and G. F. Warnke, "A d-c amplifier with cross-coupled input," *Electronics*, vol. 23, pp. 104–107; February, 1950.

[4] C. G. Montgomery, "Technique of Microwave Measurements," M.I.T. Radiation Lab. Ser., McGraw Hill Book Co., Inc., New York, N. Y., vol. 11, pp. 347–375; 1947.

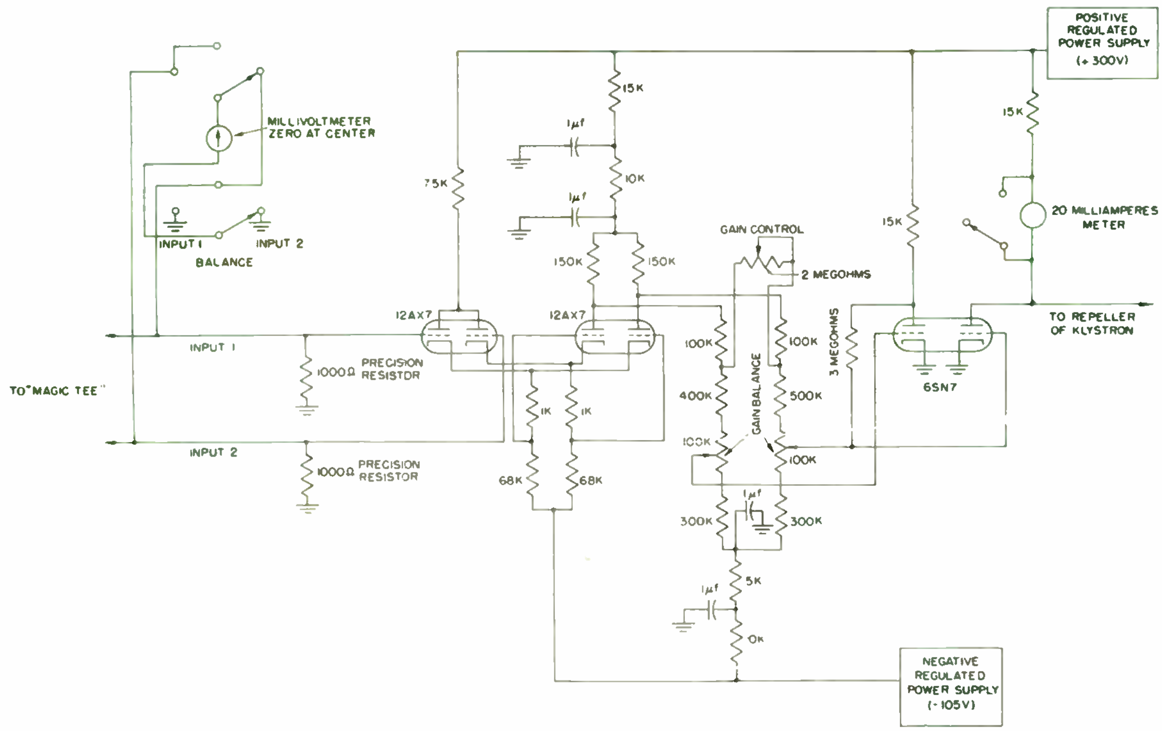


Fig. 3—Cross-coupled dc amplifier.

used; the 707B klystron was phase-locked with this same equipment to the M.I.T. frequency standard.

### Analysis of the Phase-Stabilizing Feedback Loop

The power spectrum of the output of a klystron, or any conventional oscillator, is not a single sharp line. There are three reasons. First, the klystron puts out a noise band as broad as the pass band of the loaded plate cavity. This noise may be thought of as simple diode noise.[2] Second, variations of the supply voltages on the klystron electrodes impress a frequency modulation on the carrier. Third, microphonic pickup also causes frequency modulation in the klystron through the relative physical motion of the frequency-determining elements of the oscillator.

The frequency-modulated output is written

$$S = S_0 \sin \left\{ \omega_c t + \int \omega(t) dt \right\}, \tag{1}$$

where $\omega(t)$ is the frequency modulation, and $\zeta(t) = \int \omega(t) dt$ is the phase modulation. The frequency change produced by voltage variations on an oscillator electrode may be written as

$$\frac{\partial \omega}{\partial m} = \beta \text{ radians sec}^{-1}/\text{volt}, \tag{2a}$$

where $m(t)$ is the random part of electrode voltage. The acoustical pickup might be caused by a variation of the distance of the grids in the gap of the plate cavity or by the vibration of the repeller perpendicular to the tube axis. The effect of this motion may, in general, be expressed as

$$\frac{\partial \omega}{\partial d} = \epsilon \text{ radians sec}^{-1}/\text{cm}, \tag{2b}$$

where $d(t)$ is a characteristic distance in the physical frequency determining circuit. Both $m(t)$ and $d(t)$ are assumed to represent random noise having a normal amplitude probability distribution. Hence, $\omega(t)$ can be represented as[5]

$$\omega(t) = \sum_{n=1}^{N} c_n \cos (\Omega_n t - Z_n) \tag{3a}$$

$$\overline{\omega^2(t)} = \frac{1}{2} \sum_{n=1}^{N} c_n^2 = \int_b^r W(\Omega) d\Omega, \tag{3b}$$

where $W(\Omega)$ is the frequency modulation power in radians sec$^{-1}$, and $b$ and $r$, the lower and upper cut-off frequencies, will be discussed later [see (8a) and (12)]. We would like to know the power spectrum of $S$ with the modulation (3a). The Fourier spectrum of a carrier, frequency-modulated by several independent sine waves, has been calculated by Crosby.[6] He found sidebands displaced by $\Omega_n$ from the carrier with amplitude $1/2(c_n/\Omega_n)$, and cross-modulation bands of higher order in $c_n/\Omega_n$. The energy of the modulated carrier is concentrated within either twice maximum frequency deviation or twice modulating frequency, whichever is greater.

Thus we have two different cases to consider, $(c_n/\Omega_n) > 1$ and $(c_n/\Omega_n) \ll 1$. In the first case we expect to find the power distributed within a band of width $[\overline{\omega^2(t)}]^{1/2}$ around $\omega_c$. This assumption was verified in an experiment in which a noise voltage of $10^5$ cps bandwidth and of known rms voltage was applied to the repeller of a klystron whose output was observed in a

[5] S. O. Rice, "Mathematic Analysis of Random Noise,—Noise and Stochastic Processes," Dover Publications, Inc., N. Y., N. Y.; 1954.

[6] M. G. Crosby, "Carrier and side frequency relations with multi-tone frequency or phase modulation," *RCA Rev.*, vol. 3, July, 1938.

spectrum analyzer. In the second case, the cross modulation was neglected, and we find

$$S = S_0 \left\{ \sin \omega_c t + \sum_1^N \frac{c_n}{2\Omega_n} \langle \sin [(\omega_c + \Omega_n)t - Z_n] \right.$$

$$\left. - \sin [(\omega_c - \Omega_n)t + Z_n] \rangle \right\}, \tag{4}$$

with the power spectrum

$$p(\Omega)d\Omega = \frac{1}{2} S_0^2 \frac{W(\Omega)}{\Omega^2} d\Omega. \tag{5}$$

We now wish to calculate the effect of the stabilizing feedback loop on the power spectrum. In order to obtain the open loop gain, $\mu(\Omega_n)$, we consider the $n$th term in (4). This term is the result of a phase modulation, $\zeta(\Omega_n)$. This modulation, present in the output of the oscillator, gives rise to a voltage output, $G_P \cdot \zeta(\Omega_n)$ from the phase discriminator. This output in turn is amplified by a factor of $\mu_D$ in the dc amplifier and converted according to (2a) into frequency modulation. The latter process can also be described as phase modulation, with a gain of $(\beta/i\Omega)$ radians/volt. Hence,

$$\mu(\Omega) = G_P \mu_D \beta \frac{1}{i\Omega} = \frac{r}{i\Omega} \quad \text{with } r \equiv G_P \mu_D \beta. \tag{6}$$

If the feedback loop is now closed, the $n$th term will be reduced by a factor $[1/(1-\mu(\Omega))]$, and the power spectrum of the stabilized oscillator, if $r$ is assumed to be real, is given by

$$p_s(\Omega)d\Omega = \frac{1}{2} S_0^2 \frac{W(\Omega)d\Omega}{r^2 + \Omega^2}. \tag{7}$$

where $p_s(\Omega)$ is the stabilized noise power in watts/radian sec$^{-1}$. With an estimate of $W(\Omega)$ and $r$, the order of magnitude of the residual noise left in the stabilized system may be calculated. From (6) and typical numbers $G_P = 0.04$ volts/radian; $\mu_D = 2$; $\beta = 10^7$ radians sec$^{-1}$/volt, $r$ may be computed as $r = 8 \cdot 10^6$ radians sec$^{-1}$. This leads to a stabilization cut-off frequency $\Omega_c/2\pi$, where $|\mu(\Omega_c)| = 1$, of

$$\frac{\Omega_c}{2\pi} = \nu_c = 130 \text{ kc}. \tag{8a}$$

We can estimate $W(\Omega)$ from (3b). If we assume it in a first approximation to be independent of $\Omega$ and use an empirical value of $\overline{\Omega^2(t)} = 10^8$ (radian sec$^{-1}$)$^2$ for the unstabilized klystron, and assume that these deviations are the result of noise up to $\Omega_c$ we compute

$$W(\Omega) = 100 \text{ radians sec}^{-1}, \quad \text{for} \quad \Omega < \Omega_c.$$

Hence

$$p_s(\Omega) = 8 \cdot 10^{-11} P_c \text{ in the pass band.} \tag{8b}$$

We estimate[2]

$$p_{\text{diode}} \doteq 10^{-16} P_c. \tag{8c}$$

By integrating (7) we find

$$P_{st} = 100 P_c \int^{r} \frac{d\Omega}{r^2 + \Omega^2} = \frac{100 P_c}{r} \arctan 1 \doteq 10^{-4} P_c, \tag{8d}$$

where $P_c$ is the carrier power in watts; $p_{\text{diode}}$ is the power density resulting from diode noise in the klystron in watts/radian sec$^{-1}$; and $P_{st}$ is the total noise output of the stabilized circuit, up to $\nu_c$, in watts.

The phase angle $\zeta$ of the stabilized signal still has a Gaussian amplitude probability distribution.

$$P(\zeta) = \frac{1}{(2\pi\overline{\zeta^2})^{1/2}} \exp(-\zeta^2/2\overline{\zeta^2})$$

$$\overline{\zeta^2} = \frac{P_{st}}{P_c} = 10^{-4}, \tag{9}$$

where $(\overline{\zeta^2})^{1/2}$ is the rms value of the phase variation, computed for the stabilized system.

Following conclusions are suggested by our results:

1. Eqs. (8b) and (8d) show that the noise power in the whole spectrum of the stabilized oscillator is far below the carrier. This justifies the assumption of the validity of (4).

2. Eqs. (8b) and (8c) show that between $\omega_i - 2\pi\nu_c$ and $\omega_i + 2\pi\nu_c$ the noise power produced by the klystron diode noise is negligible, compared with the frequency-modulation noise of the stabilized output. Although an ordinary broadband amplitude-modulation detector is insensitive to frequency modulation, so the diode noise is dominant, the high-$Q$ experiments described in the introduction will detect the frequency modulation noise.

3. Eq. (8a) shows that the stabilization loop has a pass band of only 130 kc. The dc amplifier will show no phase shift in this band; therefore, the loop will have a phase shift of $\pi/2$ and will be absolutely stable.

4. Eq. (9) shows that the rms phase shift is roughly 1 degree and that a shift bigger than $\pi/4$ is quite impossible. Hence, the phase discriminator operates in the linear region of its discriminator characteristic.

5. The stabilization changes the klystron power spectrum that was originally spread over a finite width (of approximately 100 cps) into an impulse function and a very low and broad noise band. This result rests on the assumption that the reference signal is monochromatic. In practice, the spectrum width of the stabilized oscillator will be reduced to the width of the reference signal.

## Modifications of the Stabilizing Circuit

However obvious it may seem, it does appear worthwhile to point out that this phase-locking circuitry may also be used to make a divider of particular use in the microwave region where no other kind exists. The operation of the circuit has been discussed from the point of view of locking a microwave oscillator to the harmonics of a lower frequency oscillator. However, the correction signal may also be applied to the lower frequency oscillator to transfer to it the stability of the high-frequency oscillator. This divider type of operation would be quite useful, i.e., for the general utilization of output of molecular microwave oscillator frequency standards.[7,8]

[7] H. Lyons, "Spectral lines as frequency standards," *Ann. N. Y. Acad. Sci.*, vol. 55, art. 5; 1952.

[8] J. P. Gordon, H. J. Zeiger, and C. H. Townes, "Molecular microwave oscillator and new hyperfine structure in the microwave spectrum of NH$_3$," *Phys. Rev.*, vol. 95, pp. 282–284; 1954.

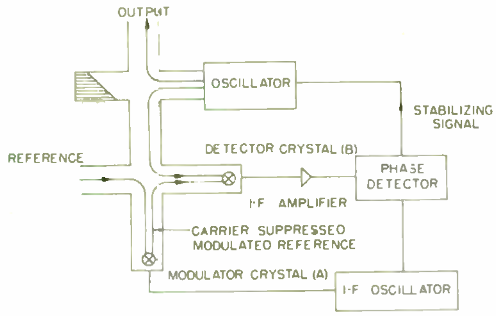


Fig. 4—Stabilizing circuit using carrier-suppressed modulation.

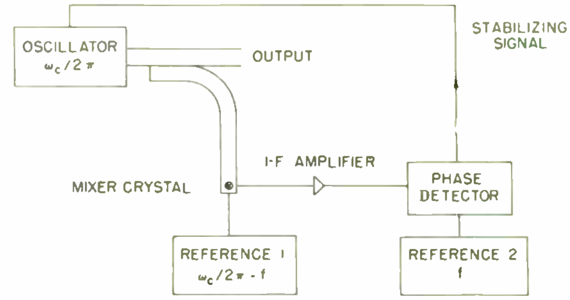


Fig. 5—Stabilizing circuit using second reference.

Figs. 4 and 5 show two of the many modifications of the feedback loop. These two forms allow a discussion that is sufficiently general that it can be applied to variations of the basic circuit.

In Fig. 4 the hybrid junction is used to apply carrier-suppressed modulation to the reference signal (in crystal $A$) and to combine it with part of the oscillator output. Detection of the resultant signal in crystal $B$ is followed, after amplification, by a second detection in a phase detector whose output is the stabilizing signal.

The whole stabilization problem may be transformed to any convenient frequency, of course, by converting the oscillator signal with a stable reference. This may be accomplished in the manner shown in Fig. 5. Here, the oscillator is converted by the reference to a frequency of $f$ cps. The converted signal is amplified and compared with a second reference in a phase detector whose output serves again as the stabilizing signal. The relative stability of the second reference can be worse than that of the first reference by the ratio of their frequencies. Thus the second reference may be obtained from a tunable source so that the frequency of the stable oscillator can be varied even if the first reference is fixed.

In the circuits of both Fig. 4 and Fig. 5 use is made of IF power coming from a detector crystal; in the original circuit dc power coming from two detector crystals was used. This is of interest in connection with the problem of detector noise. If a crystal diode (1N26) rectifies a small signal $P_c$, a noise power density $p_k(\Omega)$ will appear in excess of the thermal noise:[2]

$$p_k = \frac{10^{15} kTP_c^2}{\Omega} \text{ watts/radian sec}^{-1}. \tag{10}$$

If an IF carrier is generated in the diode, a similar noise power spectrum is found, distributed around the carrier as it was before around the dc carrier, in accordance with (10).[9,10] Whether this power results from both frequency modulation and amplitude modulation or from amplitude modulation alone is not, at present, decided. It would seem that the latter is more likely. If this is true we have a means of minimizing the effect of the crystal noise on the stabilizing signal. This AM crystal noise may be cancelled by using a phase detector that is insensitive to amplitude modulation; for example, one that is adjusted to work with zero output at equilibrium. If the oscillator is stabilized with the dc circuit the influence of the crystal noise on the output power spectrum cannot be balanced out because it arises from two independent sources—the two detecting crystals. Since this appears offhand to be a serious fault of the dc stabilization, we conclude with a calculation of influence of crystal noise on stabilized oscillator spectrum.

A noise voltage $s_N(\Omega)d\Omega$ introduced at the discriminator (see Fig. 1) is reduced by $1/(1-\mu(\Omega))$ by feedback, and appears as phase modulation $\mu_D \beta s_N(\Omega)d\Omega/i\Omega(1-\mu)$ at the output. The noise power

$$p_N(\Omega)d\Omega = Rp_kP_c \frac{\mu_D^2\beta^2}{\Omega^2 + r^2} d\Omega \tag{11}$$

is calculated in a fashion similar to $p_s(\Omega)$ in (7). We find that $p_N(\Omega)$ exceeds $p_s(\Omega)$ only when

$$\Omega < \frac{2R10^{16} kTP_c^2 \mu_D^2 \beta^2}{W(\Omega)} = 350 \text{ radians sec}^{-1}$$

if we set $p_N(\Omega) = p_s(\Omega)$ and assume $P_c = 10^{-4}$ watts. This means that only in a band of 400 cps around the stabilized signal the noise introduced by the crystal exceeds $p_s(\Omega)$. Furthermore, even 1 cps from the carrier, $p_N(\Omega) < 10^{-7}P_c$. The contribution of the total crystal noise to phase modulation noise $P_{Nt}$ is $\int_t^r p_N(\Omega)d\Omega$. We get

$$P_{Nt} = \frac{1}{2}\, 10^{16} kTRP_c^3 \mu_D^2 \beta^2 \frac{1}{r^2} \ln\left[\frac{1}{2}\left(1 + \frac{r^2}{b^2}\right)\right]. \tag{12}$$

For $b = 10^{-4}$, the lowest frequency observable within an hour, we obtain $P_{Nt} = 6 \cdot 10^{-7} P_c$. Comparing (12) with (8d) it may be seen that $P_{Nt}$ is far less than $P_{st}$. Eq. (9) indicates that the probability of saturating the discriminator has not increased. This shows that the stability of the output signal is not significantly affected by crystal noise. The circuit of Fig. 4 has therefore no special advantage. Its realization has been tried with some success, but it is difficult to set all the adjustments properly. The circuit of Fig. 5 makes the oscillator more flexible frequency-wise but it does require a broad band IF amplifier. The original dc circuit is found to be simple and quite effective.

[9] M. W. P. Strandberg, H. R. Johnson, and J. R. Eshbach, "Apparatus for microwave spectroscopy," *Rev. Sci. Instr.*, vol. 25, pp. 776–792; August, 1954.

[10] M. Strieby, "Transistors: Circuit Noise Problems," Quart. Prog. Rep., M.I.T. Res. Lab. Elec., pp. 110–113; July 15, 1953.

# Correspondence

## Kompfner Dip Conditions*

Kompfner[1] has described operation of a traveling-wave tube amplifier so that zero output occurs for nonzero input. This occurs for a particular value of beam voltage and current; measurement of these quantities enables exact calculation of circuit phase velocity and impedance. Kompfner presents calculations based on the assumptions of a long tube, and zero space charge.

Because of the wide use of Kompfner's technique, it appears justifiable to generalize his theory through use of Pierce's three-wave theory.[2] This will remove the assumptions that the tube be long and that space charge be negligible, but leave the assumption $C \ll 1$. The notation of Pierce is used throughout.

TABLE I

CONDITIONS FOR THE KOMPFNER DIP

$QC=0$

| $L$ | $CN$ | $(\beta-\beta_e)l$ | $b$ | $d$ |
|---|---|---|---|---|
| 0 | .3141 | −3.0040 | −1.522 | 0 |
| 3.201 | .2931 | −2.9116 | −1.581 | .2 |
| 6.017 | .2755 | −2.8369 | −1.639 | .4 |
| 8.527 | .2603 | −2.7721 | −1.695 | .6 |
| 10.79 | .2471 | −2.7169 | −1.750 | .8 |
| 12.85 | .2354 | −2.6697 | −1.805 | 1.0 |

$QC=0.2$

| $L$ | $CN$ | $(\beta-\beta_e)l$ | $b$ | $QC/CN$ | $H$ |
|---|---|---|---|---|---|
| 0 | .3363 | −3.1780 | −1.504 | 0.5947 | 1.890 |
| 3.391 | .3105 | −3.0492 | −1.563 | 0.6441 | 1.745 |
| 6.318 | .2893 | −2.9449 | −1.620 | 0.6913 | 1.626 |
| 8.898 | .2716 | −2.8620 | −1.677 | 0.7364 | 1.526 |
| 11.20 | .2565 | −2.7916 | −1.732 | 0.7797 | 1.441 |
| 13.29 | .2434 | −2.7332 | −1.787 | 0.8217 | 1.368 |

$d=0$

| $QC/CN$ | $CN$ | $(\beta-\beta_e)l$ | $b$ | $QC$ | $H$ |
|---|---|---|---|---|---|
| 0 | .3141 | −3.004 | −1.522 | 0 | 0 |
| 0.5947 | .3363 | −3.178 | −1.504 | .2 | 1.890 |
| 0.7280 | .3434 | −3.239 | −1.501 | .25 | 2.158 |
| 1.2531 | .3990 | −3.843 | −1.533 | .5 | 3.545 |
| 1.3803 | .4347 | −4.438 | −1.625 | .6 | 4.231 |
| 1.6098 | .4659 | −5.368 | −1.834 | .75 | 5.070 |
| 2.0354 | .4913 | −6.396 | −2.072 | 1.00 | 6.174 |
| 2.3697 | .5275 | −7.549 | −2.278 | 1.25 | 7.410 |
| 2.7164 | .5522 | −8.672 | −2.499 | 1.5 | 8.500 |

$L$ is the total circuit loss in db, $l$ is the physical length of the active part of the circuit and beam, and $H=2\pi CN\sqrt{4QC}$.

Application of a small signal of voltage $V$ to the tube input sets up three waves of incremental propagation constants $\delta_1$, $\delta_2$, and $\delta_3$ which are the roots of (7.14) of Pierce (which contains a misprint),[3]

$$\delta^2 = \frac{1}{(-b+jd+j\delta)} - 4QC. \qquad (1)$$

The total voltage amplitudes of these three waves are given by (9.4) of Pierce[4] with

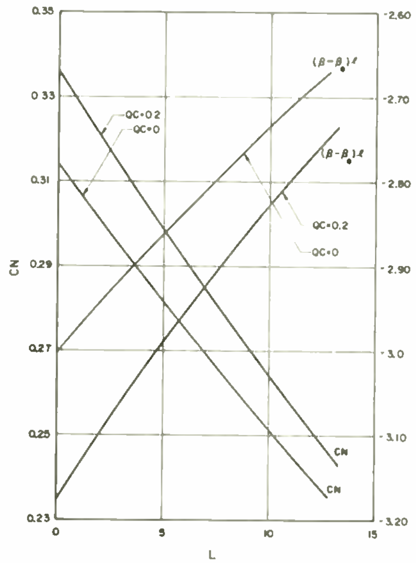


Fig. 1—Plot of conditions for the Kompfner dip.

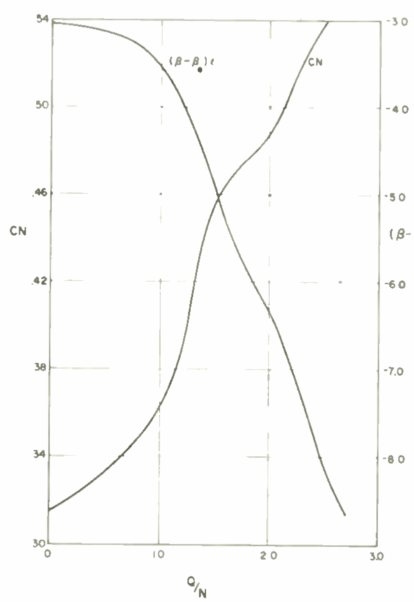


Fig. 2—Plot of conditions for the Kompfner dip. Circuit loss $L=0$.

$i=v=0$, namely

$$V_1 = \frac{V\delta_1^2}{(\delta_1-\delta_2)(\delta_1-\delta_3)} \qquad (2)$$

and cyclical permutations. The total voltage at $z=0$ is equal to the circuit voltage, so $V(0)$ is the actual input circuit voltage; the circuit voltage at $z$ is related to the total voltage at $z$ by (7.17) of Pierce.[5] With (1) this can be written

$$\frac{V_{c1}}{V_1} = \frac{\delta_1^2+4QC}{\delta_1^2} \qquad (3)$$

and cyclical permutations. In (3) we have added to Pierce's notation numerical subscripts needed to distinguish the three component waves. Thus, the total circuit voltage at $z$ is

$$\frac{V_{c1}+V_{c2}+V_{c3}}{V}\exp(j2\pi N) = \frac{(\delta_1^2+4QC)\exp(2\pi CN\delta_1)}{(\delta_1-\delta_2)(\delta_1-\delta_3)} + \frac{(\delta_2^2+4QC)\exp(2\pi CN\delta_2)}{(\delta_2-\delta_3)(\delta_2-\delta_1)} + \frac{(\delta_3^2+4QC)\exp(2\pi CN\delta_3)}{(\delta_3-\delta_1)(\delta_3-\delta_2)}. \qquad (4)$$

Now using Pierce's (15) of appendix 7,[6] namely

$$d = 0.0183L/CN, \qquad (5)$$

we can compute the real roots $CN$ and $b$ of the complex equation

$$\frac{V_{c1}+V_{c2}+V_{c3}}{V} = 0, \qquad (6)$$

where $L$ and $QC$ are parameters. The root of lowest $CN$ is tabulated in Table I, and plotted in graphical form in Figs. 1 and 2.

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* Received by the IRE, February 23, 1955.

[1] R. Kompfner, "On the operation of the traveling-wave tube at low level," *Jour. Brit. IRE*, vol. 10, pp. 283–289; August–September, 1950.

[2] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., New York, 1950.

[3] *Ibid.*, p. 113.

[4] *Ibid.*, p. 133.

[5] *Ibid.*, p. 114.

[6] *Ibid.*, p. 255.

## Correction

J. H. Crysdale, one of the authors of the discussion on "Large Reduction of VHF Transmission Loss and Fading by the Presence of a Mountain Obstacle in Beyond-Line-of-Sight Paths," which appeared on pages 627–628 of the May, 1955 issue of the PROCEEDINGS OF THE IRE, has brought the following correction to the attention of the editors.

The phase of the second term in eq. (1) is

$$(f_2+\psi_1-\phi_1)$$

and not

$$(f_2-\psi_1-\phi_1).$$

The figures accompanying the discussion have been interchanged.

The multiplying factor in footnote 2 is

$$\exp(-j\zeta_1)$$

and not

$$\exp(-j\zeta').$$

## Frequency Stable LC Oscillators*

I would like to comment on the above paper by J. K. Clapp.[1] Mr. Clapp admits that the linear analysis by Edson and others indicates that the series-tuned oscillator is no more stable than a high $C$ Colpitts oscillator having the same circuit $Q$ and the same impedances presented to the tube. He then brings up the effect of harmonic components and resorts to linear theory to attempt to prove that the series-tuned circuit is superior.

In this attempt he has made two basic errors. (1) He substitutes an equivalent circuit (Fig. 4) which is a special case pertaining only to resonant operation. Because $Rg=-Rs$, only three operating conditions are possible: a) resonance where we have either zero voltage or infinite current, b) below resonant frequency where the series impedance is pure capacitive reactance and the current leads the voltage by 90 degrees, c) above resonant frequency where the series impedance is pure inductive reactance and the current lags the voltage by 90 degrees. (2) The generator phase angle $\phi$ represented by (30) has no meaning. The phase relation between the current through the generator and its terminal voltage is entirely determined by the load connected to the terminals. The phase relation between the current through a generator and its internally generated voltage is determined by the sum of the impedances of the generator and the load. Had Mr. Clapp included the load impedance into his calculation he would have found that $df/d\phi$ for Fig. 4 was a discontinous function.

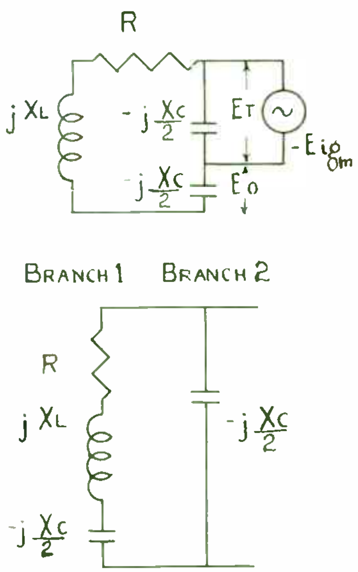


Fig. 1.

It should be clear from the above that (40) is not valid because it was not based on valid assumptions. An analysis of an equivalent circuit substituting a constant current generator for the tube will indicate that the function $df/d\phi$ is the same for the Colpitts and for the series-tuned circuit for equal $Q$'s and equal impedances presented to the tube. The calculation of points on a Nyquist diagram which follows shows that the two circuits are entirely equivalent at resonance, near resonance, and at the second harmonic of resonant frequency (Fig. 1).

$$Z_1 = R + j\left(X_L - \frac{X_C}{2}\right)$$

$$Y_1 = \frac{1}{R + j\left(X_L - \frac{X_C}{2}\right)}$$

$$Y_2 = \frac{1}{-j\frac{X_C}{2}}$$

$$Y_T = Y_1 + Y_2 = \frac{1}{R + j\left(X_L - \frac{X_C}{2}\right)} - \frac{1}{\frac{X_C}{2}}$$

$$Y_T = \frac{j\frac{X_C}{2} - \left[R + j\left(X_L - \frac{X_C}{2}\right)\right]}{j\frac{X_C}{2}\left[R + j\left(X_L - \frac{X_C}{2}\right)\right]}$$

$$Z_T = -\frac{j\frac{X_C}{2}\left[R + j\left(X_L - \frac{X_C}{2}\right)\right]}{R + j(X_L - X_C)}$$

$$i = -Ei\,\mathrm{gm}$$

$$E_T = iZ_T$$

$$= \frac{Ei\,\mathrm{gm}\,j\frac{X_C}{2}\left[R + j\left(X_L - \frac{X_C}{2}\right)\right]}{R + j(X_L - X_C)}$$

$$I_1 = E_T Y_1$$

$$= \frac{Ei\,\mathrm{gm}\,j\frac{X_C}{2}\left[R + j\left(X_L - \frac{X_C}{2}\right)\right]}{R + j(X_L - X_C)} \cdot \frac{1}{R + j\left(X_L - \frac{X_C}{2}\right)}$$

$$I_1 = \frac{Ei\,\mathrm{gm}\,j\frac{X_C}{2}}{R + j(X_L - X_C)}$$

$$E_0 = -I_1 j\frac{X_C}{2} = \frac{Ei\,\mathrm{gm}\frac{X_C^2}{4}}{R + j(X_L - X_C)}$$

$$= \frac{Ei\,\mathrm{gm}\,X_C^2}{4[R + j(X_L - X_C)]}$$

$$u\beta = \frac{E_0}{Ei} = \frac{\mathrm{gm}\,X_C^2}{4[R + j(R_L - X_C)]}.$$

For a practical case let $gm = 10^{-3}$ at

$$\omega_0 + X_C = 20\Omega, \quad X_L = 20\Omega, \quad R = .1\Omega$$

$$u\beta = \frac{10^{-3} \times 400}{4 \times .1} = 1$$

at

$$\omega_0 + \tfrac{1}{10}\%, \quad X_C = 19.98\Omega, \quad X_L = 20.02\Omega, \quad R = :1\Omega$$

$$u\beta = \frac{10^{-3} \times 400}{4(.1 + j.04)} = \frac{.4}{.4 + j.16}$$

$$= .862 - j.344$$

at

$$2\omega_0, \quad X_L = 10\Omega, \quad X_L = 40\Omega, \quad R = .1\Omega$$

$$u\beta = \frac{10^{-3} \times 100}{4(.1 + j30)} = \frac{.1}{.4 + j120} = \frac{1}{4 + j1200}$$

by a similar development for the series tuned oscillator (Fig. 2)

$$u\beta = \frac{\mathrm{gm}\,X_C^2}{4[R + j(X_L - X_{CV} - X_C)]}.$$

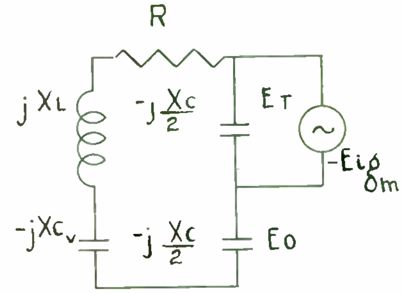


Fig. 2.

Let

$$\mathrm{gm} = 10^{-3}$$

at

$$\omega_0, X_L = 500\Omega, X_{CV} = 400\Omega, X_C = 100\Omega, R = 2.5\Omega$$

$$u\beta = \frac{10^{-3} \times 10^4}{4(2.5)} = 1$$

at

$$\omega_0 = \tfrac{1}{10}\%, \quad X_C = 99.9\Omega, \quad X_{CV} = 399.6\Omega,$$

$$X_L = 500.5\Omega, \quad R = 2.5\Omega$$

$$u\beta = \frac{10^{-3} \times 10^4}{10 + j4} = \frac{100 - j40}{116}$$

$$= .862 - j.344$$

at

$$2\omega_0, \quad X_C = 50\Omega, \quad X_{CV} = 200\Omega,$$

$$X_L = 1000\Omega, \quad R = 2.5\Omega$$

$$u\beta = \frac{10^{-3} \times 2500}{4(2.5 + j750)} = \frac{2.5}{10 + j3000}$$

$$= \frac{1}{4 + j1200}.$$

This should end all contention that the series-tuned circuit has any electrical advantage over any other, so that any choice of oscillator circuits may be made on the basis of practical advantages. On this basis it would seem that a circuit which is useful over a 2:1 or 2.5:1 frequency ratio would in most cases be preferable to one which is useful over a 1.2:1 range.

Actual experiments have borne out the above theoretically derived proof that the series-tuned oscillator is basically no more stable than other configurations.

The opinions herein are those of the writer and are not to be construed as representing the views of the Navy Department of the Naval Service at large.

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* Received by the IRE, August 26, 1954.
[1] Proc. IRE, vol. 42, pp. 1295–1300; August, 1954.

## Rebuttal[2]

Commander Bernard questions the equivalent circuit of Fig. 4 of this paper and states that it is applicable as a "special case pertaining only to resonant operation." Frankly, that is exactly what it is intended to represent. Apparently, he loses sight of the significance of the generator resistance, $Rg = R_s$, as representing a voltage rise, which balances the drop in the circuit-series resistance, $R_s$, at *resonance* and *at the stable current of operation*. The voltage is certainly not zero, and the current is certainly not infinite. The only possible mode of operation is at series resonance, hence discussion of operation above or below resonance is misdirected. The entire development was based on circuit as shown, i.e. with load connected.

The paper states, and it is reiterated, that the author "admits that the series-tuned oscillator is no more stable than a high-C Colpitts having the same circuit $Q$ and the same impedances presented to the tube," and a mathematical development is presented to prove the point again. This point is but of minor academic interest.

If the references of the paper are consulted, particularly reference 8 to Prof. Edson's book, it would be appreciated that over a wide range of frequencies up to some tens of megacycles, in a high-$C$ circuit, presenting the same impedances to the tube, it is difficult or impossible to realize the same $Q$. On this practical consideration, the series-tuned circuit will always have higher frequency stability than high-$C$ Colpitts.

A further practical consideration is that in a low-impedance high-$C$ circuit the tuning range is severely restricted, if usually available components of small volume and cost are employed.

As Prof. Edson sums it up: "In the Colpitts oscillator, the reactances required for optimum stability are often impracticably small; and an attempt to realize the calculated values is frustrated by poor values of $Q$, impracticably large variable condensers, and other similar limitations."

One of the main objects of the author's paper was to present circuits having frequency ratios greater than 1.2:1 (the ratio of the circuit of the first paper[3]) while maintaining the advantages of the series-tuned oscillator. Since the paper describes circuits of up to 2.5:1 frequency ratio, the objection to the series-tuned oscillator on the basis of frequency ratio cannot be sustained.

Commander Bernard's experiments to prove the series-tuned oscillator no more stable than other oscillators must have been conducted at frequencies high enough to make the advantages of the series-tuned oscillator begin to deteriorate. It would be interesting to know how equality of the $Q$'s and equality of impedance levels presented to the tube were established. The experiences of others, as given by references 1 and 4, indicate that improvements in stability of from 10 to 100 times over conventional circuits are readily obtained.

G. G. Gouriet's and the author's procedure in taking the phase shift of the fundamental, due to harmonic distortion, as equivalent to the nonlinear effect, was based on Llewellyn's work, referred to in reference 9. The method used by Gouriet, and followed by the author, for evaluating the effect of the reactive component, and the conclusion that the effect is reduced by increasing the inductance of the oscillator circuit are not correct. This problem will be considered in a later communication. More importantly, this work pointed out the deleterious effect of disturbances, such as unwanted feedback from amplifiers, which had not been appreciated previously.

The circuit in question is amenable in design and adjustment and gives satisfactory performance in many fields, ranging from standard frequency crystal oscillators, through many varieties of laboratory and testing oscillators, to master oscillators for transmitters.

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[2] Received by the IRE, January 25, 1955.
[3] J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," Proc. IRE, vol. 36, pp. 356–358; March, 1948.

## Characteristic Impedance of Air-Spaced Strip Transmission Line*

The authors of a recent paper[1] describe an improved mathematical technique for determining the characteristic impedance of a TEM wave in an air-spaced transmission line of the form illustrated in Fig. 1. The method leads to results which differ considerably from those previously published,[2] the latter being based on approximate solutions originated by Maxwell, Palmer and others. The difference is most marked for relatively high values of impedance (e.g. $Z_0 > 80\Omega$).

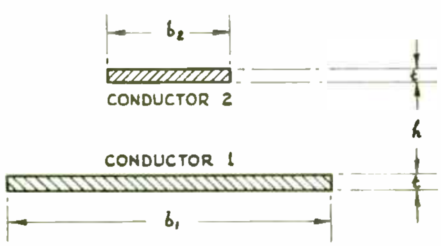


Fig. 1—Cross section of air-spaced microstrip line.

As part of a general program on strip-line techniques we have, at these laboratories, carried out a fairly comprehensive analysis of the properties of different strip-line geometrics with the aid of an electrolytic tank. The tank employed was of the double-layer type,[3] thus ensuring a good approximation to an infinite medium in the transverse plane. Instrumentation difficulties, however, limited the over-all accuracy (in the case of impedance measurements) to an estimated $\pm 3$ per cent.

A few of these results, restricted to the region of interest, are presented in Fig. 2. The thickness of the conductors is fixed at $t = b_2/32$ throughout, but repetition of some of the measurements for $t = b_2/64$ and $t = b_2/96$ showed that within this range, at any rate, the impedance is largely independent of strip thickness. (The change in conductor loss is, however, appreciable.) It will be noted that they are in excellent agreement with the numerical results calculated by Black and Higgins,[1] and hence confirm the inadequacy of the earlier formulations. The agreement would seem to be more than a coincidence and I therefore feel that the authors deserve considerable praise for their hard-won achievement.

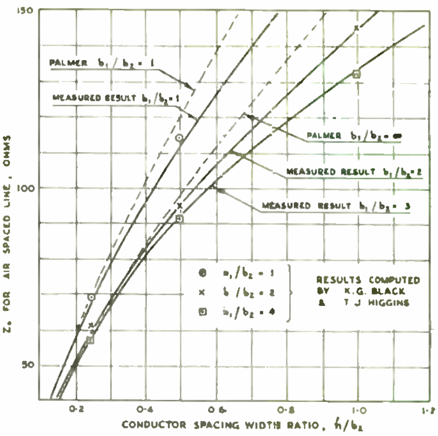


Fig. 2—Curves of characteristic impedance for air-spaced microstrip line.

The sandwich (or tri-plate) line has also been studied and good agreement has likewise been obtained with published results.[4] In both cases considerable data has been collected in regard to conductor losses, capacitative coupling between adjacent strip conductors (considerably lower in the case of the tri-plate line), and the effect on attenuation of changes to the shape of the conductor edge (found to be negligibly small). It is hoped to include much of this data in a forthcoming publication.

In conclusion it should be stressed that Assadourian and Rimai[2] never claimed anything other than an approximate solution. It has always been clear that a rigorous solution to the air-spaced line would be of rather academic interest only, since practical microstrip lines usually involve a dielectric supporting slab. In the latter case the boundary conditions require (as a minimum) that there exist a longitudinal component of the electric vector. Sciegienny and Schetzen, whose work is mentioned briefly,[5] made some progress towards a rigorous solution, and, for an idealized type of slab supported line, identified a dominant mode of the EH type. However, in discussions with them in 1953 I was led to understand that attempts to obtain an explicit solution had not been successful, and that numerical computations of wavelength, results of which are quoted,[5] had proved very tedious.

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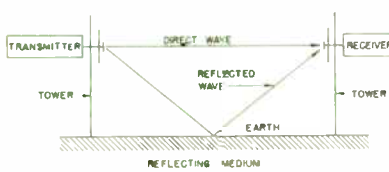
* Received by the IRE, February 28, 1955.
[1] K. G. Black and T. J. Higgins, "Rigorous Determination of the Parameters of Microstrip Transmission Lines," Symposium on Microwave Strip Circuits, Tufts College, Mass.; October 11–12, 1954.
[2] F. Assadourian and E. Rimai, "Simplified theory of microstrip transmission systems," Proc. IRE, vol. 40, pp. 1651–1657; 1952.
[3] A. R. Boothroyd, E. C. Cherry, and R. Makar, "An electrolytic tank for the measurement of the steady-state response, transient response, and allied properties of networks," *Jour. IEE*, part 1, vol. 96, p. 163; May, 1949.
[4] S. B. Cohn, "Characteristic impedance of the shielded-strip transmission line," Trans. IRE, vol. MTT-2, pp. 52–57; July, 1954.
[5] J. Sciegienny and M. Schetzen, "Theoretical Analysis of a Strip Transmission System," Part XI, Quart. Prog. Rep., Res. Lab. Elec., MIT; April 15, 1953.

## Reflection Coefficients of Irregular Terrain at 10 Cm*

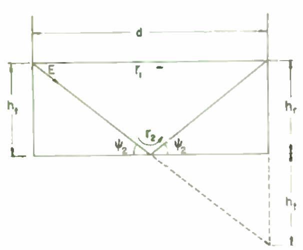
A recent paper on the above subject by K. Bullington[1] reminds the present authors of some relevant data accumulated while they were members of the Research Laboratories of Sperry Gyroscope Co. at Garden City, New York.

The work of Bullington was concerned with gross reflection of radio waves encountered in point-to-point relay transmission. Because the separation between points of transmission and reception is large, the reflection takes place from a large area of the earth, usually being quite inhomogeneous. In order to understand this and related problems better, we chose to study reflections from sample regions, carefully selected to be homogeneous in character. It is believed that the behavior of reflections from large regions can best be understood by studying homogeneous samples of the various types of surfaces that are likely to be encountered. Our experiments were performed at a wavelength of 10 cm and included the study of salt and fresh water, dry sand and soil, moist sand, dry soil with vegetation, and ice. The bulk of the data was accumulated in 1943, on various sites on Long Island, New York.

The method of measuring the reflection coefficient of a surface employed in this investigation is illustrated in Fig. 1. A suitable radio transmitter and a receiver were supported on portable towers above the surface to be studied as shown. The direct and reflected waves arrive at the receiver after traveling the paths indicated. The intensity of the field at the receiver depends upon the vector sum of the two waves. The relative phase of the direct and reflected waves may be changed by changing the height of the receiver. By observing the character of the resulting interference pattern as a function of height of the receiver, the strength of reflected waves can be determined in the usual way, making it possible to determine the magnitude of the reflection coefficient.



(a)



(b)

Fig. 1 (a) Arrangement of apparatus in making measurements of reflection coefficient. (b) Idealized geometry of the experiment.

* Received by the IRE, January 17, 1955.

[1] K. Bullington, "Reflection coefficient of irregular terrain," PROC. IRE, vol. 42, pp. 1258–1262; August, 1954.

The symbols used in the following discussion are as follows: $d$ is the horizontal separation between the transmitter and the receiver, $r_1$ is the length of path of the direct wave, $r_2$ is the length of the path of the reflected wave, $h_t$ is the height of the transmitter above plain earth, $h_r$ is the height of the receiver, and $\psi_2$ is the angle of incidence of the reflected wave.

Using the above notation and assuming that $d \gg \lambda$, $h_t = h_r \gg \lambda$, and $r_1 = d$, the reflection coefficient $R$ can be shown to be

$$R = \frac{1}{\cos \psi_2} \frac{1 + \dfrac{E_{\max}}{E_{\min}}}{1 - \dfrac{E_{\max}}{E_{\min}}},$$

where $E_{\max}$ and $E_{\min}$ are the maximum and minimum values observed at the receiver as it is moved vertically through the interference pattern. In practice, one may not always be able to satisfy the approximations used in arriving at the above formula, and more elaborate calculations become essential. Moreover, the transmitting and receiving antennas are partially directional, thus changing the relative strength of the direct and reflected waves. All of these details were accounted for in our work by the expedient of using specially prepared nomographs. The reflection coefficients for the various types of earth are presented in Figs. 2 to 13. An attempt was made in each case to correlate these data with theory. Whenever possible, values of dielectric constant and conductivity were obtained from existing sources and appropriate curves were computed using the well-known methods.[2]

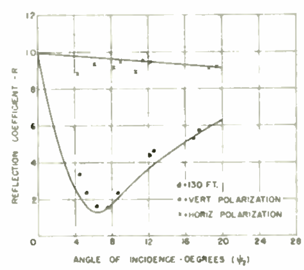



Fig. 2—Reflection coefficient for salt water, Merrick Canal. Surface: Sea water covered with estimated 1 inch ripples. Points—experimental; theoretical curves drawn for $\epsilon = 69$, $\sigma = 6.5 \times 10^{-11}$ emu.

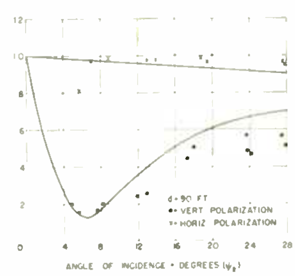



Fig. 3—Reflection coefficient for salt water, Short Beach. Surface: Tidal flat covered with 18 inches of sea, 2 inch ripples. Points—experimental; theoretical curves drawn for $\epsilon = 69$, $\sigma = 6.5 \times 10^{-11}$ emu.

[2] C. R. Burroughs, "Radio propagation over plain earth—field strength curves," *Bell Sys. Tech. Jour.* vol. 16, pp. 45–75; January, 1937.

F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Co., Inc., New York; pp. 700–707.

K. A. Norton, "Ground Wave Propagation," Federal Communication Commission, Document 47475.

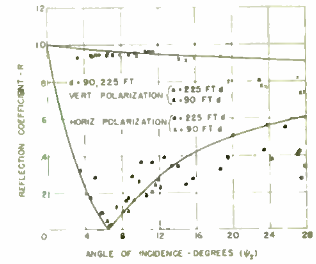



Fig. 4—Reflection coefficient for fresh water, Kenyon Farm. Surface: Fresh water pond, smooth. Theoretical curves plotted for $\epsilon = 80$.

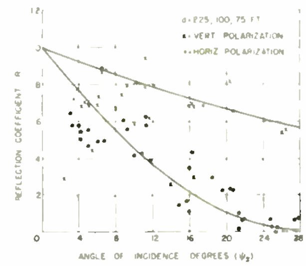



Fig. 5—Reflection coefficient for dry sand, Oak Beach. Surface: Small dry sand hillocks, some vegetation. Theoretical curves drawn for $\epsilon = 4$.

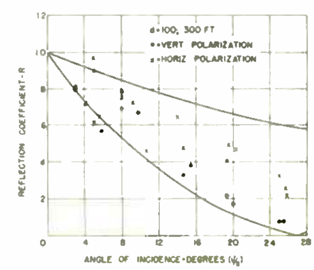



Fig. 6—Reflection coefficient for harrowed field, New York State Agricultural Institute. Surface: Harrowed field, clay-sand soil, lumps 1 to 2 inches in diameter. Theoretical curves drawn for $\epsilon = 4$.

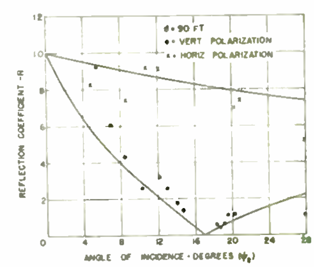



Fig. 7—Reflection coefficient of a tidal flat at low tide, Short Beach. Surface: Tidal flat with some organic material. Theoretical curves drawn for $\epsilon = 10$.

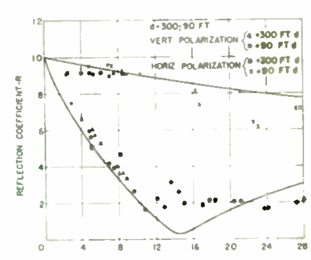



Fig. 8—Reflection coefficient for moist sand, Short Beach. Surface: Moist sand, some algae, very smooth. Theoretical curves drawn for $\epsilon = 15$.

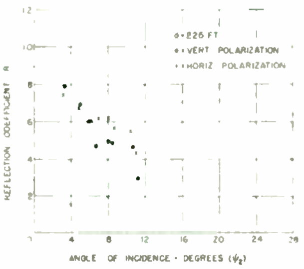



Fig. 9—Reflection coefficient of dry soil, Hicksville Airport. Surface: Rolling field, dry soil, grass 4 inches long.

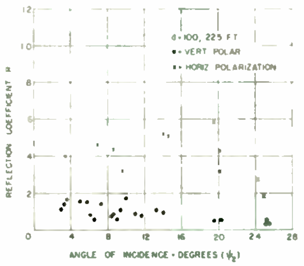



Fig. 10—Reflection coefficient of a grass covered field, Garden City, New York. Surface: Slightly rolling, grass 4 to 18 inches high, dry.

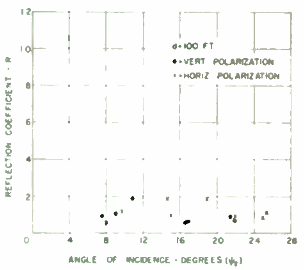



Fig. 11—Reflection coefficient of an agricultural field, New York State Agricultural Institute. Surface: Beet field covered with weeds.

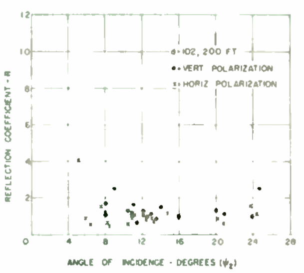



Fig. 12—Reflection coefficient of brush covered terrain, New York State Agricultural Institute. Surface: Growth of pine trees 3 to 10 feet tall, bushes, weeds, and gravelly soil.

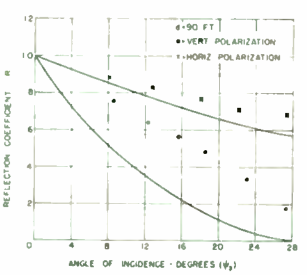



Fig. 13—Reflection coefficients of ice, Lake Tiorati, New York. Surface: Smooth ice, 22 inches thick. Theoretical curves drawn for $\epsilon=4$.

Too accurate an agreement cannot be expected as the dielectric constant and conductivity are not directly applicable to the heterogeneous material encountered in our experiments.

Our results show, as would be expected from optical reasoning, that earth, which is rough when compared to $\lambda/2$, does not produce specular reflection and the reflection coefficient has no meaning. For example, plowed fields produce almost complete scattering. Earth covered with vertical grass may produce specular reflection for horizontal polarization, and scattering for vertical polarization. In the case of smooth and isotropic surfaces, the conventional theory of reflection coefficient applies, provided that the dielectric constant and conductivity of the surface are known. By fitting curves to the experimental data presented above, values of dielectric constant and conductivity have been determined and appear to be in fair agreement with known or estimated values of these constants as obtained by other means.

We wish to acknowledge the help of Messrs. W. Frost, J. Singer, and M. Dickerson for help in accumulating the above information and to V. R. Learned for numerous suggestions. We also appreciate the encouragement and help of Dr. W. T. Cooke, and the permission of Sperry Gyroscope Co. to publish this information.

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## Note on Helix Propagation*

It has often been stated that all space harmonic components of a helix mode have the same group velocity.[1] By a tacit assumption that the familiar relation among power flow, energy storage, and group velocity is valid for the separate space harmonics, it has been implied that all space harmonics carry power in the same direction. This writer thinks that this implication is erroneous because the above relation is not a valid one for the separate space harmonics. The usual proof[2] fails because these fields separately carry complex power into the cylindrical surface in which the helix lies. Helical-line cut-off properties can be clarified if one is not bound by the concept to which this writer takes exception.

Brief evidence in support of the writer's position is shown by examination of power flow conditions as the high frequency cut-off point of the delay-line mode (Sensiper's $\beta_0$ mode) is approached. Here the impedance (total axial power flow divided by square of current) drops rapidly and is finally zero at cutoff, where the field is a complete standing wave. It is proposed that the decrease in total power flow is not the result of all powers in space harmonic fields decreasing, but rather by a subtraction of the power in the $-1$ space harmonic field from the aggregate power in the remaining fields. The $-1$ field carries power in the backward direction, which is in the same direction as its phase velocity. As cutoff is approached, this field spreads radially as its phase velocity approaches the velocity of light; it becomes increasingly dominant and finally cancels the forward power to produce cutoff. This reasoning applies to both the shielded and unshielded helical lines and tends to support earlier evidence[3,4] that cutoff is not simply the result of a radiation condition, which is peculiar to the unshielded helix only.

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* Received by the IRE, March 14, 1955.

[1] See, for example, S. Sensiper, "Electromagnetic wave propagation on helical structures (a review and survey of recent progress)," Proc. IRE, vol. 43, pp. 149–161; February, 1955.

[2] S. Sensiper, "Electromagnetic Wave Propagation on Helical Conductors," M.I.T. Res. Lab. Elec. Tech. Rep. No. 194, eq. (106); May, 1951.

[3] L. Stark, "The lower modes of a concentric line having a helical inner conductor," *Jour. Appl. Phys.*, vol. 25, pp. 1155–1162; September, 1954.

[4] J. R. Pierce and P. K. Tien, "Coupling of modes in helixes," Proc. IRE, vol. 42, pp. 1389–1396; September, 1954.

## Design Considerations of Junction Transistors at Higher Frequencies*

The lowest frequency for which the T-network obtained in this paper[1] remains correct is governed by the validity of the approximation of the frequency variable $s_p$:

$$s_p = \frac{W_0}{(D_p\tau_p)^{1/2}}(1+j\omega\tau_p)^{1/2} \approx \frac{W_0}{\sqrt{D_p}}(j\omega)^{1/2}. \quad (1)$$

By using the exact expression above for $s_p$, one can extend the applicability of the T-network down to zero frequency. Retaining the previous expressions for $z_A$, $z_B$, and $z_C$, in terms of $z_{a0}$ and $z_{b0}$, and using for the collector capacitor impedance on a $2b$-ohm impedance level

$$Z_2 = \frac{2b}{C_c s} = \frac{2}{K s_p^2 - A}, \quad (2)$$

where $s=j\omega$, $A=C_c/b\tau_p$, and $K=D_pC_c/bW_0^2$, one now obtains for the T-arm impedances $z_A$, $z_B$, and $z_C$,

$$z_A = \frac{2\left(K-\dfrac{A}{s_p^2}\right)+\dfrac{\tanh s_p/2}{s_p/2}}{\left(Ks_p-\dfrac{A}{s_p}\right)\coth s_p+1}, \quad (3)$$

$$z_B = \frac{\dfrac{\tanh s_p/2}{s_p/2}}{\left(Ks_p-\dfrac{A}{s_p}\right)\coth s_p+1}, \quad (4)$$

and

$$z_C = \frac{\dfrac{2}{s_p \sinh s_p}}{\left(Ks_p-\dfrac{A}{s_p}\right)\coth s_p+1}. \quad (5)$$

* Received by the IRE, March 16, 1955.

[1] H. Statz, E. A. Guillemin, and R. A. Pucel, Proc. IRE, vol. 42, pp. 1620–1628; November 1954.

The poles of these functions are obtained as the solutions of the equation

$$Ks_p - \frac{A}{s_p} + \tanh s_p = 0; \qquad (6)$$

or, with $s_p = ju$, from the equation

$$\tan u = -Ku - \frac{A}{u}. \qquad (7)$$

For the typical values $C_c = 100$ $\mu\mu$f, $b = 2\times 10^{-5}$, $\tau_p = 10^{-4}$, as given in the article, $A \approx 0.05$; thus, one may ignore the term $A/u$ in comparison with $Ku$ and obtain the same solutions for $u$:

$$u_\nu = \frac{\nu\pi}{2} \quad \text{for} \quad \nu = 1, 3, 5, \cdots. \qquad (8)$$

Using the same definition of the frequency variable $p$,

$$p = \frac{4}{\pi^2}\frac{W_0^2}{D_p}s \approx \frac{s}{\omega_\alpha}, \qquad (9)$$

and introducing the definition

$$d = \frac{4}{\pi^2}\frac{W_0^2}{D_p\tau_p} \approx \frac{1}{\omega_\alpha \tau_p}, \qquad (10)$$

one obtains

$$s_p = \frac{\pi}{2}\sqrt{p+d} = ju. \qquad (11)$$

Hence the poles can be expressed in terms of $p$ as

$$p_\nu = -\nu^2 - d \quad \text{for} \quad \nu = 1, 3, 5, \cdots. \qquad (12)$$

However, since $d$ is of the order of $10^{-4}$, then

$$p_\nu \approx -\nu^2 \quad \text{for} \quad \nu = 1, 3, 5, \cdots, \qquad (13)$$

which coincide with the previous results.

The residues in these poles, which may be obtained following the pattern outlined in the article, are:

$$\text{residues of } z_A \approx \frac{16}{\pi^2}, \qquad (14)$$

$$\text{residues of } z_B \approx \frac{32(-1)^{(\nu-1)/2}}{K\pi^3\nu}, \qquad (15)$$

$$\text{residues of } z_C \approx \frac{32(-1)^{(\nu+1)/2}}{K\pi^3\nu}, \qquad (16)$$

since $A/K \ll 1$ and

$$\left|\frac{\tan \nu\pi/4}{\nu\pi/4}\right| \ll 2K \text{ for } \nu = 1, 3, 5, \cdots.$$

These residues are the same as those obtained previously.

However, *$z_C$ no longer has a pole at $s_p = 0$ (or $p = 0$)*. For as $s_p \to 0$,

$$z_C = \frac{2}{(Ks_p^2 - A)\cosh s_p + s_p \sinh s_p} \to \frac{2}{(Ks_p^2 - A) + s_p^2}.$$

Thus, $z_C$ is finite at the origin, but has a pole at

$$s_p^2 \approx \frac{A}{K+1}, \qquad (17)$$

or, since

$$s_p = \frac{\pi}{2}\sqrt{p+d} \quad \text{and} \quad A = \frac{\pi^2}{4}Kd,$$

this pole is also given by $p = -d/K$. For this pole,

$$z_C \to \frac{8}{K\pi^2}\frac{1}{p + d/K},$$

and its residue there is $8/K\pi^2$, which is the same as that obtained in the article for the pole of $z_C$ at the origin.

This residue in conjunction with the new pole position will appear in the equivalent circuit as a resistor shunted across the series capacitor in the $z_C$-arm of the T-network, Fig. 9 of the article. The net result is that the new T-network will permit a base current to flow even at zero frequency—which is more in accordance with physical facts. This is the major change effected by the use of the exact expression for $s_p$. It can be shown that $z_A$ and $z_B$ also have simple poles at $p = -d/K$ with residues $-d/K$ and $+d/K$, respectively.

The infinite partial fraction expansions for $z_A$, $z_B$, and $z_C$ corresponding to those obtained in the article are:

$$z_A = \frac{16}{\pi^2}\left[\frac{-\pi^2}{16}\frac{d/K}{p+d/K} + \frac{1}{p+1} + \frac{1}{p+9} + \frac{1}{p+25} + \cdots\right], \qquad (18)$$

$$z_B = \frac{32}{K\pi^3}\left[\frac{\pi^3}{32}\frac{d}{p+d/K} + \frac{1}{p+1} - \frac{1/3}{p+9} + \frac{1/5}{p+25} - \cdots\right], \qquad (19)$$

$$z_C = \frac{32}{K\pi^3}\left[\frac{\pi/4}{p+d/K} - \frac{1}{p+1} + \frac{1/3}{p+9} - \frac{1/5}{p+25} + \cdots\right]. \qquad (20)$$

Note that for $d = 0$ (which corresponds to approximating $s_p$ by $W_0(j\omega)^{1/2}/\sqrt{D_p}$), these expressions coincide with those obtained in the article. Using the exact expressions for $z_A$, $z_B$, and $z_C$,

$$z_A(0) \approx 1; \qquad z_B(0) \approx 1;$$

$$z_C(0) \approx 8/\pi^2 d = 2D_p\tau_p/W_0^2,$$

for $0 < d \ll 1$ and $K \gg 1$.

If the infinite expansions are terminated after a finite number of terms, the approximations will be improved if the approximate net effect of the abandoned terms is taken into account in the range of $p$-values considered.

For example, if terms up to and including the pole at $p = -25$ are retained, for the range $0 \leqq p \leqq j6$, the terms dropped are essentially resistive and nearly equal to their respective values at $p = 0$. Consequently, one may represent their combined effect in each series by a constant evaluated at the origin. These constants are the same as those obtained in the article; thus, the modified circuit will revert to that of the article when $d = 0$.

The normalized equivalent circuit for the T-network corresponding to Fig. 9 of the article is shown here in Fig. 10. The portion enclosed by the dotted lines extends the applicability of the circuit obtained in the article to radian frequencies below $1/\tau_p$. This circuit may be denormalized by use of Table I of the article.[1]

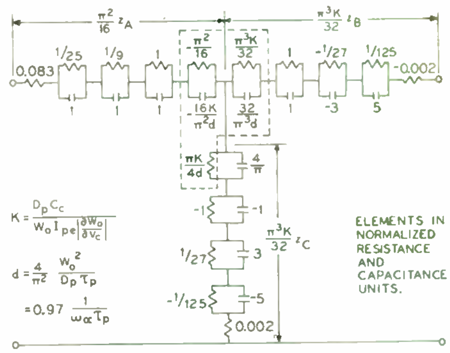


Fig. 1—Explicit form of the T-network applicable to low frequencies (four terms in the expansion).

In conclusion it is informative to observe that this equivalent circuit describes the transistor correctly at zero frequency. For this purpose the current through the collector (which is considered to be shorted) is compared with the current through the base. From the circuit the ratio of base to collector current is simply equal to $z_B/z_C \approx (\pi^3K/32)/(\pi K/4d) = W_0^2/2D_p\tau_p$ since $K \gg 1$ and $0 < d \ll 1$.

It is also very simple to calculate the ratio directly from the physical properties of the transistor. The emitter current is approximately equal to the collector current $I_c \approx I_e = AqD_p \text{ grad } p \approx AqD_p p_e/W_0$, where $A$ is the emitter area, $p$ is the concentration of holes, and $p_e$ is the concentration of holes at the emitter side of the base. The base current, which is equal to the total recombination current, is given by $(q/\tau_p)\int p dv \approx (Aqp_eW_0)/2\tau_p)$. The ratio of the emitter current to the base current is $W_0^2/(2D_p\tau_p)$, which agrees with the above result. The fact that the emitter efficiency is not equal to unity was not considered, but it is included in the complete circuit of Fig. 3 of the article.

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## "yrneh"*

The writer is indebted to Professor True McLean for having called to his attention the true origin of the term "yrneh."[1] It appears that the term was originated in 1910 by the famed Professor Vladimir Karapetoff[2] of Cornell University.

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* Received by the IRE, April 8, 1955.

[1] H. Stockman, "On reciprocal inductance," Proc. IRE, vol. 43, p. 341; March, 1955.

[2] V. Karapetoff, "The Magnetic Circuit," McGraw-Hill Book Co., New York, N. Y., p. 10; 1911.

## On Reciprocal Inductance*

I wish to concur with Mr. Baghdady[1] in his suggestion of the term "inertiance" for reciprocal inductance. A further suggestion is the use of the term "erny" for the units of "inertiance" is a sort of tribute to Ernst Guillemin of M.I.T., who has done so much to popularize the use of reciprocal inductance in his courses on "Guillementary Circuit Analaysis."

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* Received by the IRE, December 30, 1954.
[1] Proc. I.R.E., p. 1807; December, 1954.

## Optimum Patterns for Endfire Arrays*

In a recent paper, DuHamel[1] has described a method of synthesizing an equal-minor-lobe, or Tchebycheff, directivity pattern for an endfire linear array. This method is based on a generalization of the synthesis procedure suggested by H. J. Riblet[2] in extending the work of C. L. Dolph[3] for the broadside array to include the case of arrays having an element spacing less than a half wavelength. Dolph's original method yields the optimum pattern only for element spacings greater than a half wavelength, whereas Riblet's method permits an optimum pattern to be specified for any element spacing. However, Riblet's method is applicable as such to arrays having an *odd* number of elements only, whereas Dolph's method may be used for even or odd numbers of elements.

The purpose of this communication is to point out that the method originally described by Dolph may be applied directly to the case of the endfire array. An optimum pattern may be obtained for the endfire array for *any* element spacing, in contrast with the case of the broadside array. (It should be noted, however, that element spacings greater than a half wavelength normally are not used in an endfire array in order to avoid extra major lobes in the pattern.[4]) There are two advantages of using Dolph's method. First, the same method may be applied directly to arrays having even or odd numbers of elements. Second, the resulting equations for determining the relative currents for the elements (excitation coefficients) are of somewhat simpler form than the corresponding equations derived by DuHamel from Riblet's method.

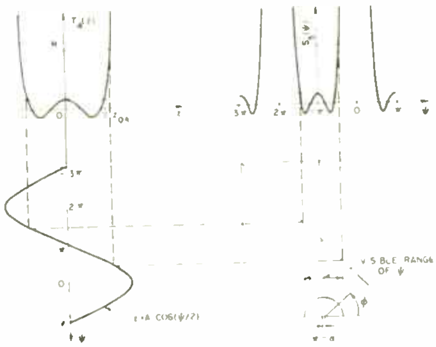


Fig. 1—Graphical construction of optimum directivity pattern $S_5(\psi)$ from fourth-degree Tchebycheff polynomial $T_4(z)$ for five-element endfire array with $d/\lambda = \frac{1}{4}$.

Application of Dolph's method to the endfire array is illustrated in Fig. 1 for the five-element array used by DuHamel (cf. his Fig. 5). Notation used here is the same as that in DuHamel's paper. The space factor $S_5(\psi)$ expressed as a function of the auxiliary variable $\psi$ is determined from the *fourth*-degree Tchebycheff polynomial $T_4(z)$ by means of a linear transformation $z = A \cos(\psi/2)$. (DuHamel determines his space factor from the *second*-degree polynomial $T_2(z)$ with $z = a \cos \psi + b$.) In either case, the physical directivity pattern, expressed as a function of $\phi$, is described by the space factor $S_5(\psi)$ over a limited (visible) range of $\psi$. As shown in Fig. 1, the unknown constants $A$ and $\alpha$ are determined by the two conditions that for $\phi = 0$, $|S(\psi)| = R$, or $z = z_0$, while for $\phi = \pi$ $|S(\psi)| = 1$ or $z = -1$.

Further details of this method cannot be included here because of space limitations. However, a more complete description, including numerical examples for the seven-element array used by DuHamel and for a *four*-element array, plus a description of an alternative method of overdesigning a supergain antenna, has been submitted for publication to the *Transactions of the IRE*, Professional Group on Antennas and Propagation.[5] It also might be noted that the bidirectional array considered by DuHamel may be treated by a similar adaptation of Dolph's method.[6]

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* Original manuscript received by the IRE, August 11, 1953; revised manuscript received, May 12, 1955.
[1] R. H. DuHamel, Proc. I.R.E., vol. 41, pp. 652–659; May, 1953.
[2] H. J. Riblet, "Discussion on A current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," Proc. I.R.E., vol. 35, pp. 489–492; May, 1947.
[3] C. L. Dolph, "A current distribution for broadside arrays which optimizes the relationship between beam width and side-lobe level," Proc. I.R.E., vol. 34, pp. 335–348; June, 1946.
[4] This was noted by DuHamel, *loc. cit.*, p. 655.
[5] R. L. Pritchard, "Discussion on optimum patterns for endfire arrays," *Trans. IRE*, vol. AP-3, pp. 40–43; January, 1955.
[6] R. L. Pritchard, "Optimum directivity patterns for linear point arrays," *Jour. Acous. Soc. Amer.*, vol. 25, pp. 890–891; September, 1953.

## The Unit for Frequency*

The Proceedings uses cycle, kc, and mc as units of frequency, whereas the Radiation Laboratory Series, for example, uses cps, kc/sec, and Mc/sec. The conflict between the desire for convenience and the desire to retain the usual meaning of the word "cycle" is perhaps epitomized by the following sentence (Proc. IRE, vol. 42, p. 1372; Sept., 1954, top of the page). "The light is chopped by a 90-cycle-per-second sector disc, and the ac photoresponse is measured by a 90-cycle amplifier-detector."

A happy solution would be more widespread use of the term "hertz," meaning cycle-per-second. Thus the units of frequency would be hertz (or hz), khz, and Mhz. A rate of sweeping frequency could have the unit Mhz/sec, instead of mc/sec or Mc/sec². Noise power density could have the unit watts/Mhz, instead of watts/mc or watts/(Mc/sec).

The term hertz is listed in "Electronics Dictionary" by Cook and Markus, McGraw-Hill Book Company, 1945; "German Military Dictionary," TM 30-506; "Antennas" by Kraus, McGraw-Hill Book Company, 1950; and "Science," December 24, 1954.

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* Received by the IRE, January 4, 1955.

# Contributors

C. A. Aldridge was born in Canandaigua, N. Y., on May 2, 1922. He was employed from 1940 to 1944 by the Consolidated Machine Tool Corp. and in cooperative programs studied mechanical engineering from 1940 to 1942 at the Rochester Institute of Technology, and electrical engineering from 1942 to 1944 at the University of Rochester. From 1944 to 1946 he was with the Navy at the Naval Research Laboratories and in the South Pacific.

C. A. Aldridge

Mr. Aldridge received his B.S. degree in physics in June, 1950 from Syracuse University. He completed a graduate year in physics and then joined the Electronics Laboratory in June, 1951.

Since then, he has been engaged in development work in the fields of colored television studio equipment, electromechanical simulators, and transistor circuitry, and is continuing his graduate work at Syracuse University.

B. P. Bogert was born on September 26, 1923, in Waltham, Mass. He received the degree of B.S. in physics in February, 1944, the M.S. degree in mathematics in 1946 and a Ph.D. in mathematics in 1948 from Massachusetts Institute of Technology.

During 1944 and 1945, he was a staff member of the M.I.T. Radiation Laboratory. In 1948, he joined the technical staff of the Bell Telephone Laboratories, Murray Hill, and has been primarily concerned with research in physical acoustics, and in narrow band speech transmission.

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He is a member of Sigma Xi, and a Fellow of the Acoustical Society of America.

W. F. Chow (M'53–SM'53) was born in Shanghai, China, on June 7, 1923. After receiving the B.S. degree in electrical engineering from Ta Tung University in 1945, he joined the Chapei Power Co. In 1948 he came to the United States receiving his M.S. degree in electrical engineering in 1949 and his Ph.D. in electrical engineering in 1952, both from the University of Minnesota, while he was serving there as a teaching assistant.

W. F. Chow

Dr. Chow joined the General Electric Co. in 1952. He is engaged in the research and development of transistor circuitry.

Dr. Chow is a member of Eta Kappa Nu and Sigma Xi.

B. F. C. Cooper (M'47) was born in England in 1917. He received the degrees of Bachelor of Science in 1939 and Bachelor of Engineering in 1941 from the University of Sydney. He has been a member of the research staff of the Division of Radiophysics, Commonwealth Scientific and Industrial Research Organization since 1940.

B. F. C. Cooper

During World War II he worked on many developmental aspects of ground and airborne radar, and in the immediate post-war period he was responsible for the design of an airborne distance measuring equipment. In 1946–47 he spent a period with the National Research Council of Canada, where he developed an airborne ground-profile recorder.

Since returning to Australia he has developed instrumentation for rain-physics research, and also a magnetic drum storage system for the C.S.I.R.O. Mk I Computer. At present he leads a group working on transistor electronics.

Martin Peter was born July 12, 1928, in Switzerland. He received the diploma with distinction in physics and mathematics from the Eidgenössische Technische Hochschule in Zurich in 1952.

M. Peter

Mr. Peter was awarded the Kern prize for his thesis on colloidal ferroelectrics.

Mr. Peter subsequently came to the United States and entered the Massachusetts Institute of Technology, where he is a candidate for the Ph.D. degree in the Department of Physics. While studying for his degree, he is also a research assistant in the Research Laboratory of Electronics.

Mr. Peter is a member of the American Physical Society and is an associate of Sigma Xi.

For a photograph and biography of M. W. P. Strandberg, see page 756 of the June, 1955 issue of the Proceedings of the IRE.

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R. W. Grow

From 1949 to 1951, Dr. Grow was employed as an electronic scientist at the U. S. Naval Research Laboratory, where he worked successively in the fields of radar countermeasures and nuclear physics, participating in the 1951 Atomic Weapons Tests. In 1951 he became associated with the Electronics Research Laboratory at Stanford University, where he has been engaged in microwave tube research specializing in traveling-wave tubes and backward-wave oscillators. At present he is a Research Associate at the Applied Electronics Laboratory.

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E. Keonjian (M'50–SM'52) was born in Tiflis, Russia. He received his B.S. and M.S. degrees from the Leningrad Institute of Electrical Engineering in 1932.

After graduation, Mr. Keonjian joined the Leningrad Central Radio-laboratory and in 1935 was transferred to the Leningrad Research Institute of Electronics as a senior engineer. In 1940, he taught electrical communication at Leningrad Institute of Electrical Engineering.

E. Keonjian

In 1947, Mr. Keonjian came to this country and joined Westinghouse Electric Corporation. In 1949, he was appointed as a lecturer in electrical communication at the City College of New York. Since 1951, Mr. Keonjian has been a member of the engineering staff of General Electric, where he is engaged in the development work concerned with transistor circuitry.

Mr. Keonjian is an author of numerous works on electronics, published here and abroad. He is a member of the Research Society of America, holds a professional engineering license in the State of New York, and is a co-author of the book, "Principles of Transistor Circuits."

R. C. Knechtli was born in Geneva, Switzerland, on August 14, 1927. He received his M.S. degree in electrical engineering in 1950, from the Swiss Federal Institute of Technology (E.T.H.). He was employed by Brown Boveri and Company (Baden, Switzerland), from 1951 to 1953 where he engaged in research on microwave circuits and microwave tubes. From 1951 to 1952, he was a research assistant at the Massachusetts Institute of Technology, where he also worked on microwave tubes.

R. C. KNECHTLI

Since 1953, he has been a research engineer with RCA Laboratories, at Princeton, New Jersey.

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J. D. Kraus (A'32–M'43–SM'43–F'54) was born at Ann Arbor, Mich., on June 28, 1910. He attended the University of Michigan receiving the B.S. degree in 1930, M.S. in 1931, and Ph.D. degree in physics in 1933. During the next few years he was active on industrial noise reduction problems and in nuclear research. From 1938 to 1940 Dr. Kraus was an antenna consultant. From 1940 to 1943, he was physicist and division head at the U. S. Naval Ordnance Laboratory, Washington, D. C. and between 1943 to 1946 was research associate and group leader at N.D.R.C.'s Radio Research Laboratory at Harvard University.

J. D. KRAUS

In 1946 Dr. Kraus joined the staff of the Ohio State University where he is now Professor of Electrical Engineering and Director of the Radio Observatory. Professor Kraus is the author of books on antennas and electromagnetic theory.

He is a member of the American Astronomical Society and the American Physical Society.

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J. G. Linvill (A'49) was born on August 8, 1919, in Polo, Mo. He received an A.B. degree from William Jewell College in 1941, and continued his studies at Massachusetts Institute of Technology, where he was awarded the S.B. in 1943, S.M. in 1945, and Sc.D. in 1949, all in electrical engineering. While at M.I.T. he was a member of the faculty, serving as assistant professor in electrical engineering from 1949 to 1951. At the same time, he was a consultant to Sylvania Electrical Products.

J. G. LINVILL

In 1951, Mr. Linvill joined Bell Telephone Laboratories, where he worked on active network problems involving applications of transistors as the active element. Since March of this year, he has been Associate Professor of electrical engineering at Stanford University.

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

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For a photograph and biography of J. R. Macdonald, see pages 1571–1572 of the October, 1954 issue of the PROCEEDINGS OF THE IRE.

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S. Matt (A'53) was born in Cleveland, Ohio, on September 3, 1923. He received the Bachelor of Science degree in electrical engineering from Ohio University in 1944. Following graduation, he served with the U. S. Army Air Force. He returned to Ohio University in 1946 to teach in the Department of Electrical Engineering.

S. MATT

In 1947, he entered California Institute of Technology and obtained his M.S. degree in electrical engineering in 1948. He remained there two years as a research assistant in the High Voltage Laboratory. He then entered Ohio State University and received his Ph.D. degree in 1953.

Dr. Matt was a research assistant and then an instructor while at the Ohio State University. At present, he is at the General Electric Advanced Electronics Center at Ithaca, N. Y.

Dr. Matt is a member of Sigma Xi and Pi Mu Epsilon.

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A. R. Moore was born in New York, N. Y., on January 14, 1923. He received the B.S. degree in chemistry in 1942 from the Polytechnic Institute of Brooklyn. He worked on phototube and thyratron development at RCA Victor in Harrison, N. J. and Lancaster, Pa., from 1942 to 1945. In 1945 he entered Cornell University, and received the Ph.D. in physics in 1949, specializing in physics of solids. During his last two years at Cornell he was an RCA Fellow. He joined the RCA Laboratories Division at Princeton, N. J. in 1949, where he has worked on semiconductor physics.

A. R. MOORE

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R. F. Rutz (A'51) was born in Alton, Ill., on February 9, 1919. He received the B.A. degree in 1941 from Shurtleff College, Alton, Ill., and the M.S degree in physics in 1947 from the State University of Iowa.

Mr. Rutz joined the Sandia Laboratory of the Los Alamos Scientific Research Laboratory, subsequently Sandia Corporation, at Albuquerque, N. M. in 1948, where he worked in the Electronics Research Department. In 1951 he joined the Research Laboratory of the International Business Machines Corporation at Poughkeepsie, N. Y., where he is currently working on transistor research.

R. F. RUTZ

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

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A. P. Stern (A'51) was born on July 20, 1925, in Budapest, Hungary. He studied at the Universities of Budapest and Lausanne and at the Swiss Federal Institute of Technology in Zurich, where he acquired a Master's degree in electrical engineering in 1948.

A. P. STERN

From 1948 to 1951, Mr. Stern was engaged in research and development work in the field of gaseous discharges in Switzerland. In 1950, he became Instructor for illumination engineering and photometry at the Swiss Federal Institute of Technology.

Mr. Stern came to the United States in 1951 and joined the staff of the General Electric Company's Electronics Laboratory in Syracuse, N. Y. At the present time, Mr. Stern is supervisor of solid state circuit development in the Electronics Laboratory.

Mr. Stern is a member of the Scientific Research Society of America.

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K. F. Stripp was born on January 26, 1920, in Union City, N. J. He obtained a B.S. degree in chemistry from the Polytechnic Institute of Brooklyn in 1950. He was teaching assistant in the chemistry department at California Institute of Technology from 1950 to 1951 and obtained his Ph.D. in physical chemistry from Yale in 1953. He spent three summers at RCA Laboratories and had been associated with them on a full-time basis since July 1, 1953.

K. F. STRIPP

Dr. Stripp was a member of the American Chemical Society, AAAS, American Institute of Chemists, Sigma Xi, and Phi Lambda Upsilon. He died suddenly on July 10, 1954.

J. J. Suran (A'52) was born in New York, N. Y., on January 11, 1926. After having served for three years with the U. S. Army during World War II, he received the B.S.E.E. degree from Columbia University in 1949 and continued graduate studies there and at the Illinois Institute of Technology.

J. J. SURAN

From 1949 to 1952, Mr. Suran was employed in control systems design and development by J. W. Meaker and Co., and in FM communications research and development by Motorola, Inc. Since 1952, he has been a member of the Electronics Laboratory of the General Electric Company.

Mr. Suran is a co-author of the book, "Principles of Transistor Circuits." He has a professional engineering license in New York State and is a member of AIEE.

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A. W. Warner (M'52) was born in Sewickley, Pa. in 1915. He received the B.A. degree, with a major in physics, from the University of Delaware in 1940 and the M.S. degree in physics from the University of Maryland in 1942.

In the same year Mr. Warner was a member of the faculty of Lehigh University, leaving in July to join the Western Electric Company, where he worked on the development of crystal-unit test equipment. In 1943 Mr. Warner became a member of the technical staff of Bell Telephone Laboratories, and is engaged in the design of high-frequency plated crystal units.

A. W. WARNER

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For a photograph and biography of D. A. Watkins, see page 106 of the January, 1955 issue of the PROCEEDINGS OF THE IRE.

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H. A. Wheeler (A'27–M'28–F'35) was born in St. Paul, Minn., on May 10, 1903. He received the B.S. degree in physics from George Washington University in 1925. From 1925 to 1928 he studied in the physics department of The Johns Hopkins University, and lectured there during 1926 and 1927. He was employed as a laboratory assistant in the radio section of the National Bureau of Standards in 1921, leaving in 1923 to assist Professor Hazeltine and later to join the Hazeltine Corporation in 1924. He was in charge of their Bayside laboratory from 1930 to 1937, where he later became vice-president and chief consulting engineer.

H. A. WHEELER

He has specialized in the design of radio receivers (including FM and TV), the theory of communication networks, radar (including IFF during World War II), antennas, and microwave equipment.

In 1946 Mr. Wheeler opened his own consulting office in Great Neck, N. Y. He is now also president of Wheeler Laboratories, Inc. From 1950 to date, he has been serving part-time as consultant to the Office of Secretary of Defense in the fields of guided missiles and electronics.

Mr. Wheeler is a Fellow of the American Institute of Electrical Engineers, an Associate of the Institution of Electrical Engineers, and a member of Sigma Xi. He received the Morris Liebmann Memorial Prize in 1940.

# IRE News and Radio Notes

## SECOND SYMPOSIUM ON VACUUM TECHNOLOGY INVITES PAPERS

The Committee on Vacuum Techniques, Incorporated invites the submission of papers at the Second Symposium on Vacuum Technology to be held at Mellon Institute in Pittsburgh, October 13–15. Those interested should write to Rudy Koehler, Committee on Vacuum Techniques, Inc., Box 1282, Boston 9, Massachusetts.

The program will deal with equipment, instrumentation, developments in vacuum technology, standards, nomenclature, methods and techniques, and vacuum systems applications and processes

## M.I.T. GIVES SUMMER COURSE ON NOISE IN ELECTRON DEVICES

Plans for a two-week Special Summer Program on *Noise in Electron Devices* at the Massachusetts Institute of Technology have been announced by Ernest H. Huntress, Director of the M.I.T. Summer Session. The program will be held from July 18 through July 29 and is planned primarily for those who are or plan to become research workers in the field.

L. D. Smullin, of the Research Laboratory of Electronics at M.I.T., and H. A. Haus, Assistant Professor in the Department of Electrical Engineering, will direct the program. Other members of the M.I.T. staff who will lecture include P. Elias, Assistant Professor of Electrical Engineering; Y. W. Lee, Associate Professor of Electrical Engineering; and L. Tisza, Associate Professor of Physics.

Guest lecturers will include D. O. North and R. W. Peter, both of the David Sarnoff Research Center; C. F. Quate and T. E. Talpey, of the Bell Telephone Laboratories; and A. van der Ziel, Professor of Electrical Engineering at the University of Minnesota.

Full details and application blanks for this Special Summer Program may be obtained from the Summer Session office, Room 7-103, Massachusetts Institute of Technology, Cambridge 39, Massachusetts.

## INSTRUMENTATION CONFERENCE TO MEET IN ATLANTA

The Professional Group on Instrumentation and the Atlanta Section will sponsor an Instrumentation Conference and Exhibit at the Biltmore Hotel in Atlanta, November 28 through November 30.

The theme of the conference is "Data Handling." Prospective authors of papers are invited to submit abstracts of 200 words or less on data gathering, processing, utilization, and processing systems not later than September 1. Titles and abstracts should be addressed to B. J. Dasher, School of Electrical Engineering, Georgia Institute of Technology, Atlanta, Georgia.

## AUDIO ENGINEERING SOCIETY WILL MEET OCTOBER 12 TO 15 IN N. Y.

"Practicality" will be the theme of the 1955 Convention of the Audio Engineering Society Twenty-five, which is scheduled for October 12 to 15. Sessions will be held in New York at the Hotel New Yorker and will be concurrent with the Audio Fair.

According to Richard H. Ranger, president of Rangertone, Incorporated and program chairman for the event, the convention will include panel discussions on transistors, amplifier design and tape recording. Their purpose, he said, will be to bring out the "right and easy way" to handle each type of equipment. The agenda will also include theoretical and scientific papers. The annual banquet of the society will be held on the evening of October 12.

Col. Ranger, who is Executive Vice-President of the society this year, is being aided by Effingham Kettleman of RCA.

## FINAL CALL FOR PGED PAPERS

The PGED, which will holds its First Annual Technical Meeting in Washington, D. C., October 24 and 25, is now making its final call for papers.

The meeting will include three parallel sessions: Solid State Devices, Microwave Tubes, and Non-Microwave Tubes.

## Nominations for 1956 Officers

At its May 14, 1955 meeting, the IRE Board of Directors received the recommendations of the Nominations Committee and the reports of the Regional Committees for officers and directors for 1956. They are:

*President, 1956*—A. V. Loughren
*Vice-President, 1956*—Herre Rinia
*Director-at-Large, 1956–1958 (two to be elected)* L. V. Berkner, E. W. Herold, T. A. Hunter, J. R. Whinnery
*Regional Directors, 1956–1957 (one to be elected in each Region)*
*Region 1*—C. R. Burrows, H. F. Dart, L. B. Grew
*Region 3*—J. G. Brainerd, L. R. Quarles
*Region 5*—J. J. Gershon, R. E. Moe
*Region 7*—L. E. French, C. F. Wolcott

According to Article VI, Section 1, of the IRE Constitution, nominations by petition for any of the above offices may be made by letter to the Board of Directors, giving name of proposed candidate and office for which it is desired he be nominated. For acceptance a letter of petition must reach the executive office before noon on August 12, 1955, and shall be signed by at least 100 voting members qualified to vote for the office of the candidate nominated.

## Vehicular Communications Paper Deadline Announced for the First of August

The Professional Group on Vehicular Communications will hold its Sixth Annual Meeting September 26 and 27 at the Multnomah Hotel, Portland, Oregon.

Deadline for papers is August 1; title of paper, abstract, full name and address should be submitted to Newton Monk, Bell Telephone Labs., 463 West Street, New York 14, N. Y.

## Art in Electronics Competition Announced for WESCON Show

The West Coast Electronic Manufacturers Association is sponsoring a competition in art as applied to electronics during the 1955 WESCON Show which will be held in San Francisco, California, August 24–26.

Employees of electronic manufacturers and their families are eligible to compete for cash and vacation trips in a contest planned to augment the association's annual scholarship program. Entries will be exhibited at WESCON and auctioned to the highest bidders. Proceeds from the sale will go to the WCEMA Scholarship Fund.

Purpose of the contest, according to Paul M. Cook and Mrs. Jan Smith, co-chairmen for the event, is to stimulate an awareness of the visually attractive materials, components and designs of electronics and their possible use as objects of art.

Trip prizes will be awarded in each of four categories: photography, painting and sculpture, decorative accessories, and jewelry. All entries must be created from component parts commonly used in electronics. A prize of $250.00 will go to the creator of the object bringing the largest price in the auction.

Complete details and entry blanks for the competition are available from Berk Baker, Eitel-McCullough, Incorporated, 798 San Mateo Avenue, San Bruno, California. They may also be obtained at the WCEMA office, 339 South Robertson Boulevard, Beverly Hills, California.

## Benjamin Bauer and Kenneth Goff Honored by PG on Audio

Under a plan approved last year by the Administrative Committee of the Professional Group on Audio, awards were presented during Audio Sessions at the National Convention to Kenneth E. Goff and B. B. Bauer. Mr. Bauer, of Shure Brothers, Incorporated, received $200 "in recognition of many excellent audio papers appearing in IRE publications over a period of years." Mr. Goff, who is associated with the Acoustics Laboratory at M.I.T., received $100 as the author, under 30 years old, "of an especially meritorious paper dealing with a subject related to audio appearing in any IRE publication." The name of Mr. Goff's paper was *The Development of a Variable Time Delay* and appeared in the 1953 Convention Record of the Institute of Radio Engineers.

Kenneth W. Goff

Mr. Goff graduated in 1950 from West Virginia University with the B.S. degree in electrical engineering. Upon graduation, he joined the M.I.T. Acoustics Laboratory and entered the M.I.T. Graduate School. In accepting the audio award, Mr. Goff said that the "tremendous challenge of the unsolved problems in the field of Audio, together with the opportunities for organization and individual recognition made possible by the PGA, combine to present a very attractive picture to those of us who are just beginning our work in electrical engineering."

B. B. Bauer

Mr. Bauer received the E.E. degree in 1937 from the University of Cincinnati. He has also received the Industrial Electrical Engineering degree from Pratt Institute and has attended the University of Chicago and Illinois Institute of Technology. Mr. Bauer suggested, in accepting his award, that the money be used for work among PGA Student Members. "I suggest," he said, "the creation of a fund, which could be augmented from time to time by PGA proper, and used for annual Student Awards, for meritorious papers on subjects connected with Audio Technology."

## Calendar of Coming Events

SRI and Nat. Ind. Conf. Board Symposium on Electronics in Automatic Production, Sheraton-Palace, San Francisco, Calif., Aug. 22–23

URSI Symposium on Solar Eclipses and the Ionosphere, Royal Society, Burlington House, London, England, Aug. 22–24

**IRE-West Coast Electronic Manufacturers' Association WESCON, Civic Auditorium, San Francisco, California, Aug. 24–26**

**Emporium Section Sixteenth Annual Summer Seminar, Emporium, Pa., August 26–28**

**IRE-ISA Tenth Annual Instrument Conference, Shrine Auditorium, Los Angeles, Calif., Sept. 12–16**

Association for Computing Machinery, Annual Meeting, Moore School of Electrical Engineering, U. of Pa., Sept. 14–16

**IRE Professional Group on Nuclear Science—Second Annual Meeting, Oak Ridge National Labs., Oak Ridge, Tenn., Sept. 14–17**

**IRE Cedar Rapids Section Symposium on Automation, Cedar Rapids, Ia., Sept. 17**

RETMA Automation Symposium, U. of Pennsylvania, Philadelphia, Pa., Sept. 26–27

**PG on Vehicular Communications Sixth Annual Meeting, Multnomah Hotel, Portland, Ore., Sept. 26–27**

IMSA Annual Convention, Hotel Seneca, Rochester, N. Y., Sept. 26–29

International Analogy Computation Meeting, Société Belge des Ingenieurs des Télécommunications et d'Electronique, Brussels, Belgium, Sept. 27–Oct. 1.

**IRE-AIEE Conference on Industrial Electronics, Rackham Memorial Building, Detroit, Michigan, Sept. 28–29**

**National Electronics Conference, Hotel Sherman, Chicago, Ill., October 3–5**

Audio Engineering Society Convention, Hotel New Yorker, New York City, Oct. 12–15

Second Symposium on Vacuum Technology, Mellon Inst., Pittsburgh, Pa., Oct. 13–15

**IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., Oct. 17–19**

Eighth Annual Gaseous Electronics Conference, General Electric Res. Lab., Schenectady, N.Y., Oct. 20–22

**PG on Electron Devices Annual Technical Meeting, Shoreham Hotel, Washington, D. C., Oct. 24–25**

**IRE East Coast Conference on Aeronautical and Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md., Oct. 31–Nov. 1**

Symposium on Applied Solar Energy, Westward Ho Hotel, Phoenix, Ariz., Nov. 1–5

**IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel Statler, Boston, Nov. 7–9**

**IRE-AIEE-ISA Electrical Techniques in Medicine and Biology, Shoreham Hotel, Washington, D. C., Nov. 14–16**

**PGI and Atlanta Section Data Processing Symposium, Hotel Biltmore, Atlanta, Ga., Nov. 28–30**

# LONG ISLAND SECTION HONORS FOUR IRE FELLOW RECIPIENTS AT SPECIAL AWARD CEREMONIES

## Citations Presented to Loughlin, Learned, Gaffney, and Dunning

Fellow citations are presented members by John Dyer as Pres. Ryder watches. Left to right are B. B. Loughlin, V. R. Learned, F. J. Gaffney, O. M. Dunning, J. N. Dyer, and John Ryder.

A special award meeting was held by the Long Island Section on March 20 to honor four Section members who were made Fellows of the IRE. John Dyer, Regional Director, presented the award citations and President John Ryder made the principal address. After the award presentations, a cocktail party was held in the main ballroom of the Garden City Hotel.

The four Long Island Section members who received the awards were Orville M. Dunning, Francis J. Gaffney, Vincent R. Learned, and Bernard B. Loughlin. Mr. Dunning, a member of the Board of Directors and Vice-President in Charge of Engineering at Hazeltine Corporation, received the Fellow award "... for his contributions to the field of sound recording and his effective organization of engineering effort." Vice-President for Engineering at Marion Electrical Instrument Company, Mr. Gaffney was made a Fellow "... for his contributions to the field of electrical Measurements." "... for his contributions to research and development of microwave electron tubes," Dr. Learned, of Sperry Gyroscope, received the Fellow award. A Consulting engineer with Hazeltine Corporation, Mr. Loughlin was made a Fellow "... for his contributions to color television, frequency modulation, and superregeneration."

### John R. Pierce Elected to National Academy of Sciences

John R. Pierce, Editor of IRE and Director of Electronics Research at Bell Telephone Laboratories, was elected to the National Academy of Sciences at its 92nd annual meeting held in Washington recently.

The National Academy of Sciences, a private, non-profit organization, serves as an adviser to the Federal Government in scientific matters and acts in the furtherance of science for the general welfare. The membership of the academy numbers approximately 500 in the physical and biological fields.

In addition to Dr. Pierce, IRE members who have been elected to the academy include M. J. Kelly, J. B. Fisk, W. Shockley, L. V. Berkner and F. E. Terman.

### Program for Applied Solar Energy Symposium Announced

A preliminary program has been announced for the World Symposium on Applied Solar Energy to be held in Phoenix, November 1 through 5. Sponsors of the meeting are the Association for Applied Solar Energy, Stanford Research Institute, and the University of Arizona.

Among the papers scheduled are: *The Sun's Energy*, Farrington Daniels, University of Wisconsin; *Survey of the Domestic Uses of Solar Energy*, H. C. Hottel, Massachusetts Institute of Technology; *Space Cooling With Solar Energy*, George O. G. Löf, Denver, Colorado; *Food and Fuel from Solar Energy*, F. A. Brooks, University of California; *Chlorella for Animal Food*, Jack Meyers, University of Texas; *Engineering for Algae Culture*, A. W. Fisher, Jr., Arthur D. Little, Inc., Cambridge, Mass.; and *Solar Energy Utilization by Higher Plants*, Paul C. Mangelsdorf, Harvard University. Maria Telkes, New York University, *Solar Stills;* R. C. Jordan, University of Minnesota, *Mechanical Energy from Solar Energy;* Paul Erlandson, Southwest Research Institute, *Direct Conversion of Solar Energy;* L. J. Heidt, Massachusetts Institute of Technology, *Converting Solar to Chemical Energy;* G. L. Pearson, Bell Telephone Laboratories, *Photovoltaic P-N Couples.* J. E. Hobson, director of Stanford Research, will lead a discussion on "The Economics of Solar Energy" at the outset of the meeting.

Contributions from abroad will be made by: Felix Trombe, Laboratoire de l'Energie Solaire, Paris. *High-Temperature Furnaces;* R. N. Morse, Commonwealth Scientific and Industrial Research Organization, Australia, *Solar Water Heaters;* and Hiroshi Tamiya, Tokugawa Institute for Biological Research, Tokyo, *Chlorella for Food.*

General chairman for the symposium is Lewis W. Douglas, of the Southern Arizona Bank and Trust Company, Vice-Chairman and program director is Merritt L. Kastens, Assistant Director of Stanford Research Institute. Headquarters for the symposium during its planning phase are located in Suite 204, Mayer-Heard Building, Phoenix.

### PGNS Will Meet in September

The Second National Annual Meeting of the PG on Nuclear Science will be held in Oak Ridge, Tenn., September 14–16, with the Oak Ridge Chapter as host.

Persons wishing to present papers are invited to submit, before July 15, 200-word abstracts to H. E. Banta, Papers Committee, Oak Ridge National Laboratory, Box P, Oak Ridge, Tenn. Since papers will be accepted in 15, 30, and 45 minute categories, specification of the time required is requested.

## Professional Group News

### Eight New Chapters Approved

At its meeting of April 3, the IRE Executive Committee officially approved the following chapters: Los Angeles Chapter, PG on Automatic Control; Los Angeles Chapter, PG on Reliability and Quality Control; Chicago Chapter, PG on Communications Systems; Pittsburgh Chapter, PG on Electronic Computers; Atlanta Chapter, PG on Instrumentation.

On May 3 three chapters were approved by the Executive Committee. They were: San Francisco Chapter, PG on Audio; Twin Cities Chapter, PG on Automatic Control; Long Island Chapter, PG on Instrumentation.

At an earlier meeting the Northwest Florida Subsection was made a full Section and the Fort Huachuca Subsection was established in the Phoenix Section.

## Technical Committee Notes

The **Antennas and Waveguides** Committee met at IRE Headquarters on April 20 with P. H. Smith presiding. The committee approved the *Proposed Standard on Antennas and Waveguides: Definitions for Waveguide Components* for submission to the Standards Committee. Subcommittee 2.4 on Waveguide and Waveguide Component Measurements presented drafts of "Methods of Measurement" comprising discussions on Measurements of Phase Shift, Measurement of Power Handling Capacity, and Measurement of Q. A good deal of discussion followed, with the suggestion that this subject be re-examined by the subcommittee.

The **Antennas and Waveguides** Committee met at IRE Headquarters on March 9 with P. H. Smith presiding. The committee reviewed the *Proposed Standards on Antennas and Waveguides, Definitions for Waveguide Components*. W. E. Waller reported on the Proposed Standards: Waveguide and Waveguide Component Measurements.

D. E. Maxwell presided at the **Audio Techniques** Committee meeting at IRE Headquarters on April 21. L. D. Runkle, Chairman of Subcommittee 3.1 on Definitions, reported that after two meetings the subcommittee has adopted and tentatively approved 27 definitions. The committee reviewed the *Proposed Standards on Methods of Measurement of Gain, Loss, Amplification, Attenuation and Frequency Response*, which they expect to complete at their next meeting.

D. E. Maxwell presided at the **Audio Techniques** Committee meeting at IRE Headquarters on March 24. L. D. Runkle, Chairman of Definitions Subcommittee 3.1, reported that his subcommittee is reviewing a list of proposed definitions. R. C. Moody, Chairman of the West Coast Subcommittee 3.3, reported that his subcommittee is working on a Proposed Standard on the Measurement of Intermodulation Distortion, which they hope to finish by the end of the year. The committee reviewed the proposed Standards on Methods of Measurement of Gain, Loss, Amplification, Attenuation and Frequency Response.

The **Electronic Computers** Committee met at IRE Headquarters on March 24 with Robert Serrell presiding. After discussion of the work of the Definitions Subcommittees, Mr. Brown made and Dr. Haynes seconded the following motion, which was unanimously approved: "The Electronic Computers Committee requests that its Subcommittees 8.4 and 8.5 make plans to distribute glossaries of new terms being considered by them. The glossaries should be sent with the PGEC Transactions and include all desirable explanations of the terms."

The **Facsimile** Committee met at IRE Headquarters on April 22 with Chairman H. Burkhard presiding. The committee discussed the Facsimile Test Chart which they are preparing. The following definitions were approved: electrostatic recording and magnetic recording.

The **Facsimile** Committee met at the Times Annex on March 18 and H. Burkhard presided. The committee reviewed their tentative definitions of terms. The following terms were proposed as possibly requiring definitions: xerographic recording, electrostatic recording, magnetic recording, ferrographic recording.

The **Information Theory and Modulation Systems** Committee met at IRE Headquarters on March 9 with J. G. Kreer, Jr. presiding. Dale Pollack was appointed chairman of the Modulations Systems Subcommittee and Peter Elias was appointed chairman of the Information Theory Subcommittee. The following terms were referred to the subcommittee on Information Theory for study: binary digit, message coding versus symbol coding, systematic versus unsystematic coding, and corrector and characteristic as used in coding.

H. R. Mimno presided at the **Navigation Aids** Committee meeting at IRE Headquarters on April 22. The committee commenced its exploratory examination of the proposed measurements standard on the VHF Omnirange. The discussion covered the *Introduction, Description and Characteristics to be Measured*, together with the introductory paragraphs on *Specialized VOR Test Equipment*. Mr. Moskowitz recorded certain editorial changes arrived at by general agreement, and noted additional suggestions to be considered by an editorial group.

The **Piezoelectric Crystals** Committee met at IRE Headquarters on March 21. W. P. Mason presided. W. L. Bond discussed nomenclature used at the recent meeting of the committee of the International Crystallographic Congress. Three proposals on methods of determining the piezoelectric, elastic, and dielectric constants of crystals and the parameters of piezoelectric vibrators were discussed.

Ernst Weber presided at the meeting of the **Standards** Committee at the Jade Room of the Waldorf-Astoria Hotel on March 24. Dr. Weber explained IRE standardization procedure to new members, and A. G. Jensen described the functions of the Standards Coordinator. There was a review of the past year's work and an announcement of future plans. Mr. Jensen introduced the new members of the Standards Committee. E. A. Laport expressed the appreciation of the group to the recent chairmen, Dr. Brainerd, Mr. Jensen, and Dr. Weber, for assuming the responsibility of leadership of the committee. He stated that the committee had been fortunate in having capable chairmen to lead the group.

The **Standards Committee** met at IRE Headquarters on April 7 with Chairman E. Weber presiding. A report of the work of the Ad Hoc Committee on Spurious Radiation was submitted by R. F. Shea, with a recommendation that this committee be dissolved. This motion was unanimously approved. The formation of a Nuclear Techniques Technical Committee with G. A. Morton as chairman was unanimously approved. No action was taken on the *Proposed Standards on Pulses: Methods of Measurement of Pulse Quantities* or the *Proposed Definitions on Induction and Dielectric Heating* since there were no representatives of the sponsoring committees present.

The **Video Techniques** Committee met at IRE Headquarters on March 31 with W. J. Poch presiding. Mr. Jones reported on the recent activities of the Subcommittee on Video Transmission. The problem of specifying the exact conditions for taking measurements of differential gain and differential phase were discussed at length. Dr. Athey reported that he was making progress in collecting a list of standard terms in the field of kinescope recording. The following definitions were approved by the committee: flyback, linearity control, camera tube, contrast ratio, geometric distortion, dc restorer, dc transmission, return interval, return trace, raster, retrace interval, brightness control, nominal line width, progressive scanning.

## Sixth PGUE Group Administrative Committee Meeting March 22 During IRE Convention

PGUE officers are, *left to right:* Morris Kenny, retiring Secretary; Julia Herrick, Vice-Chairman; Amor Lane, Chairman of Membership Committee and retiring Group Chairman; Oskar Mattiat, Editor of *PGUE Transactions;* Morton Fagen, Chairman; Frank Massa, Chairman of Nominations Committee and retiring Administrative Committee member; Julius Bernstein, Treasurer and Convention Record Committee Chairman. Also elected at the same meeting were W. A. Andersen, C. M. Harris, and T. M. Lambert, Administrative Committee; W. J. Fry, Papers Study and Review Committee; and Donald Berlincourt, Associate Editor.

# 1955 Western Electronic Show and Convention

## Tentative Program
## San Francisco, California
## August 24–26

The 1955 Western Electric Show and Convention will meet in San Francisco on August 24, 25 and 26. This year there will be 570 exhibits representing more than 600 producers. Co-sponsored by the West Coast Manufacturers' Association and the San Francisco and Los Angeles Section of the Institute of Radio Engineers, representing the Seventh Region, WESCON will be attended by more than 20,000 visitors.

A United Airlines "Airlift" has been arranged to transport part of the visitors who will attend. The Airlift was arranged by Noel E. Porter of the Hewlett Packard Company and WESCON Chairman, and Mal Mobley, Jr., Business Manager. In addition to the many United schedules, special Mainliner flights will be arranged exclusively for WESCON visitors and exhibitors from major cities to San Francisco. United Airlines has arranged to handle reservations on its own line or any other scheduled airline. Confirmation will be made directly by United or through any local airline office or travel agent. Show and Convention officials urged that reservations be made as soon as possible in order to be assured of the best flights and schedules.

The Technical Program will consist of twenty-four sessions and over 100 papers. This specialized program has been closely integrated by the program committee which is made up of IRE members, coordinating both Professional Groups and Section activities.

### Wednesday Morning

#### Solid State Devices

*Transistors Today*, J. A. Morton
*Large Signal Semi-Conductor Devices*, John Saby
*High-Frequency Power Gain of Junction Transistors*, R. L. Pritchard
*Recent Developments in Germanium Alloy Junctions*, C. W. Mueller
*A New High-Ambient Transistor*, R. R. Rutherford and J. J. Bowe

#### Information Theory

*Limiting Frequency-Modulation Spectra*, N. Blachman
*The Definition of a General Metric of Information*, N. Abramson
*An Analysis of Optimum Sequential Detectors*, J. J. Bussgang and D. Middleton
*Analysis of Automatic Bias Control for Threshold Detectors*, E. Ackerlind
*Generating a Gaussian Sample*, S. Stein and J. E. Storer
*Proof of the Sampling Theorem for Stationary Processes*, A. Rosenbloom and J. Heilfron

#### Reliability and Quality Control

*Engineering and Testing for Reliability*, H. G. Romig
*Parts Versus Systems: The Reliability Dilemma*, David A. Hill
*An Effective Reliability Program Based Upon "A Triad for Design Reliability,"* F. E. Dreste
*A Basic Study of the Effects of Operating and Environmental Factors on Electron Tube Reliability*, W. S. Bowie
*Surface Contamination of Dielectric Materials*, Saul Chaikin

#### Propagation

*An Explanation of Fading in Microwave Relay Systems*, H. Magnuski
*Some Notes on Propagation over a Spherical Earth*, S. J. Fricker
*Radio Power Received via Tropospheric Scattering*, A. Waterman
*Atmospheric Attenuation of Microwave Radiation*, G. R. Marner
*Theory of Deviative Absorption in the F2 Layer and its Relation to Temperature*, R. Gallet
*Symposium on Industrial Electronics and Nuclear Engineering*

### Wednesday Afternoon

#### Broadcast and TV Receivers

*A Thin Cathode Ray Tube*, William R. Aiken
*Beam Focusing and Deflection in the Aiken Tube*, R. Madey
*Radiation Measurements at VHF and UHF*, A. B. Glenn
*An Experimental Automobile Receiver Employing Transistors*, L. A. Freedman, T. O. Stanley and D. D. Holmes
*High-Efficiency, Unipotential Post Focus, Tri-Color Picture Tube*, Wilfrid F. Niklas

#### Circuit Theory I—Transistors and Blocking Oscillators

*Advantages of Direct Coupled Transistor Amplifiers*, Richard Hurley
*Junction Transistor Blocking Oscillators*, J. G. Linville
*The Design of Blocking Oscillators as Fast Pulse Regenerators*, F. K. Bowers
*Stability of Multi-Mode Oscillating Systems*, R. W. De Grasse
*(Additional paper to be announced.)*
*Electronic Instrumentation in Aircraft—Joint Symposium of the Professional Group on Aeronautical and Navigational Electronics and the Institute of Aeronautical Sciences*
*Experiments with Radio Controlled, Dynamically Similar Models*, E. G. Stout
*Role of Electronics in Engineering Flight Testing*, W. L. Howland
*Instrumentation for Rocket Engine Testing*, R. F. Gompertz
*(Additional papers to be announced.)*

#### Antennas I

*Recent Developments in Microwave Antennas*, L. C. Van Atta
*Printed Surface Wave Antennas*, H. W. Cooper
*Circularly-Polarized Slot Radiators*, A. J. Simmons
*Radiation from Ferrite-Loaded Slot Radiators*, D. J. Angelakos and M. Korman
*A Large Aperture Differential Polarization Antenna for Radio Astronomy Use*, V. H. Goerke and O. D. Remmler

#### Instrumentation

*Beamplexer-High Speed Channel Multiplexing Unit*, H. Moss and S. Kuchinsky
*A Stable Diode Chopper Circuit*, H. Patton
*A Completely Automatic Impedance Plotter*, J. R. Vinding
*A Broadband Microwave Frequency Meter*, P. H. Vartanian and J. L. Melchor
*An Expanded Scale Frequency Meter*, Duane Marshall
*Measurement of Time Varying Frequencies*, Martin Graham

### Thursday Morning

#### Electronic Component Parts

*Design and Properties of High Voltage Glass Capacitors*, G. P. Smith
*Characteristics of Modular Electronic Components*, W. G. James
*Simple Electronic Transformer Design*, R. Lee
*Measurement of Parameters Controlling Pulse Front Response of Transformers*, P. R. Gillette, K. Oshima and R. M. Rowe
*Development of MIL-T-27-A: Transformers and Reactors*, E. M. Wiler
*International Research in Electronics and Allied Fields. Symposium I—The Role of the IRE and URSI*

#### High Power Tubes

*M-Type Backward Wave Oscillators*, J. Hull
*Considerations of Various Structures for High Average Powers in the UHF Region*, D. Preist
*Design Information on Large Signal Traveling-Wave Amplifiers*, J. E. Rowe
*A New Beam Power Tube for UHF Service*, W. B. Bennett
*An Ion Trapped High Voltage Pentode*, R. E. Hellers

#### Automatic Control

*Non-Linear Compensation of an Aircraft Instrument Servo-mechanism*, D. Lebell
*The Stabilization of Non-Linear Servomechanisms Encountered in Antenna Instrumentation*, J. Bacon
*Synthesis of a Non-Linear Control System*, I. Flugge-Lotz and C. F. Taylor
*Theory of Non-Linear Feedback Systems Having a Multiple Number of First-Order Operating Points*, J. A. Narud
*Noise in Non-Linear Servos*, G. O. Young and C. J. Savant

#### Telemetry and Remote Control

*Wow and Flutter Compensation in FM Telemetry*, W. H. Chester
*Aliasing Errors in Sampled Data Dystems*, A. J. Mallinckrodt
*Air-to-Ground Propagation over Desert Terrain at Telemetering Frequencies*, G. L. McCone

*Pulse Width Data Multiplexing of an FM/FM Subcarrier*, A. S. Westnest
*The Use of A C Excited Gauges in a PDM/PM Telemetering System*, W. F. Carmody

**Thursday Afternoon**

### Microwave Theory

*Periodic Structures for Traveling-Wave Tubes*, M. Chodorow
*Conversion of Maxwell's Equations into Generalized Telegraphist's Equations*, S. A. Schelkunoff
*On the Expansion of Fields in Lossless Microwave Junctions*, T. Teichmann
*Conformal Mapping of Rounded Polygons by a Wave-Filter Analogue*, H. A. Wheeler

### Broadcast Transmission Systems

*The Perfect Television System*, O. H. Schade
*The Subjective Sharpness of Simulated Color TV Pictures*, H. F. Huntsman
*The Conversion of a Standard TV Mobile Unit for Greater Flexibility and Operating Convenience*, H. F. Huntsman
*High Speed Duplication of Magnetic Tape Recordings*, J. M. Leslie
*Color TV Magnetic Tape Recording System*, H. F. Olson

### Computers I—Digital Computer Applications and Design Techniques

*A Punched Card Method of Evaluating Systems of Boolean Functions with Special Reference to Analysis of Relay Circuits*, W. R. Abbott
*The Elecom 50—A New Type of Computer*, Evelyn Berezin and Phyllis Hersh
*Logical Design of the Remington Rand High Speed Printer with Emphasis on the Checking and Editing Features*, M. Jacoby
*Theory, Principles and Applications of Statistical Computers*, H. Blasbalg and W. O'Hare
*A Glow Transfer Shifting Register Utilizing R-F Gas Discharge*, D. C. Engelbart
*Ferroelectric Hysteresis in Barium Titanate Single Crystals*, H. H. Wieder

### Engineering Management

*Small Engineering Company Organization—a Philosophy and Method*, T. W. Jarmie
*Is the Yardstick for Estimating Individual Engineering and Scientific Potential Reliable?* A. H. Schooley
*Management in Production Engineering*, C. Blahna
*Market Development—The Neglected Companion of Product Development*, A. D. Ehrenfried
*Cross Functional Engineering Managemetn*, C. M. Ryerson

### Aeronautical and Navigational Electronics

*An Improved Simultaneous Phase Comparison Guidance Radar*, H. H. Sommer
*Antenna Design Considerations for Helicopters*, J. B. Chown
*High Voltage Impulse Generation for Measurement of Receiver Susceptibility to Interference Encountered in Aircraft*, A. Newan and J. R. Stahmann
*Experimental Results of Conductive Cooling Tests on Airborne Equipment*, R. L. Berner

**Thursday Evening**

*Medical Electronics Panel Discussion*

**Friday Morning**

### Computers II—Analogue Computer Components and Applications

*Automatic Data Accumulation System for Wind Tunnels*, John Wedel
*Data Recorder for Evaluation of a Fire Control System*, J. T. Ator and L. P. Retzinger, Jr.
*Transistors in Current Analog Computing*, B. P. Kerfoot
*The Use of Electronic Analog Computers in the Solution of Certain Radar Noise Problems*, J. A. Aseltine
*Precision Electronic Switching with Feedback Amplifiers*, C. M. Edwards

### Circuit Theory II—Synthesis Problems

*New Methods of Transformerless Driving-Point Impedance Synthesis*, Stanley Hurst
*General Synthesis of Quarter-Wave Impedance Transformers with Given Insertion Loss Function*, Henry J. Riblet
*The Approximation Problem in the Synthesis of R-C Networks*, K. L. Su and B. J. Dasher
*A Precise Method of Designing High- and Low-Pass R-C Filters with Active Elements*, M. McWhorter
*Signal Flow Graphs for Random Signals*, W. H. Huggins

### Medical Electronics

*Recent Developments in Color-Translating Ultra-Violet Microscopy*, R. B. Holt
*Some Theoretical and Practical Aspects of Microscanning*, W. E. Tollers, et al.
*The Electrocardiophone—A New Surgical Tool*, A. J. Morris and J. P. Swanson
*Instrumentation for Spectral Phonocardiography*, George N. Webb

### Electron Tubes

*A UHF Traveling-Wave Amplifier Tube Employing an Electrostatically Focused Hollow Beam*, C. B. Crumly
*Design of Solenoids for Traveling-Wave Tubes*, J. E. Etter, A. W. Friend and W. Watson
*Light Weight Solenoids of Aluminum Foil*, W. G. Worcester and A. L. Weitzmann
*The Serrodyne—A Single Sideband Synchrodyne*, R. C. Cumming
*Recent Dark Trace Tube Developments*, S. Nozick
*Recent Developments in the Use of Dispenser Cathodes in Low and Medium Power Magnetrons*, R. S. Briggs

### Microwave Techniques

*Waveguides for Long Distance Communications*, A. C. Beck
*Recent Advances in Microwave Filter Techniques*, Seymour Cohn
*Geometrical Methods for the Analysis of Two-Part Networks*, G. A. Deschamps
*Some Applications and Characteristics of Ferrite at Wavelengths of 0.87 and 1.9 cms*, Clyde Stewart
*Measurement and Control of Microwave Frequencies by Lower Radio Frequencies*, R. C. Mackey et al.

**Friday Afternoon**

### Antennas II

*Radiation Characteristics with Power Gains for Slots on a Sphere*, Y. Mushiake and R. E. Webster
*Radiation Patterns of Asymmetrically Fed Prolate Spheroidal Antennas*, H. A. Myers
*Phase Properties of Antennas for the Dovap Millile Tracking System*, T. Morita and C. W. Steele
*Rotationally Symmetric Dialectric Microwave Lenses with Two-Dimensional Wide Angle Scanning Characteristics*, A. Mayer and E. Wantuch

### Radio Relay Systems Design

*Design of FM Radio Relay Equipment for Multi-Channel Operation*, J. W. Halina
*Factors Affecting the Spacing of Radio Terminal in an UHF Link*, I. H. Gerks
*Radio Communication with Secondary Power*, H. E. Hollmann
*Single Sideband Multiplexing as it Applies to Microwave Relays*, T. L. Leming

### International Research in Electronics and Allied Fields

### II. The International Geophysical Year Program

*The International Geophysical Year, 1957–1958*, R. J. Slutz
*Absorption Measurements During the International Geophysical Year*, Gordon Little
*Vertical Incidence Ionosphere Sounding Measurements during I. G. Y.*, J. M. Watts
*Back-Scattering Measurements During I. G. Y.*, A. M. Peterson

# Professional Groups

**Aeronautical & Navigational Electronics**—*Chairman*, Edgar A. Post, Navigational Aides, United Air Lines, Operations Base, Stapleton Field, Denver 7, Colo.

**Antennas & Propagation**—*Chairman*, Delmer C. Ports, Jansky & Bailey, 1339 Wisconsin Ave., N.W., Washington 7, D. C.

**Audio**—*Chairman*, W. E. Kock, Bell Tel. Labs., Murray Hill, N J.

**Automatic Control**—*Chairman*, Robert B. Wilcox, Raytheon Manufacturing Co., 148 California St., Newton 58, Mass.

**Broadcast & Television Receivers**—*Chairman*, W. P. Boothroyd, Philco Corp., Philadelphia 34, Pa.

**Broadcast Transmission Systems**—*Chairman*, O. W. B. Reed, Jr., Jansky & Bailey, 1735 DeSales St., N.W., Washington, D. C.

**Circuit Theory**—*Chairman*, J. Carlin, Microwave Res. Inst., Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y.

**Communications Systems**—*Chairman*, A. C. Peterson, Jr., Bell Labs., 463 West St., New York 14, N. Y.

**Component Parts**—*Chairman*, Floyd A. Paul, Reliability Bendix Development Lab., 116 W. Olive Avenue, Burbank, Calif.

**Electron Devices**—*Chairman*, George Espersen, Microwave Tube Section, Philips Labs., Irvington-on-Hudson, N. Y.

**Electronic Computers**—*Chairman*, J. H. Felker, Bell Labs., Whippany, N. J.

**Engineering Management**—*Chairman*, C. J. Breitwieser, Lear, Inc., 3171 S. Bundy Drive, Los Angeles 34, Calif.

**Industrial Electronics**—*Chairman*, George P. Bosomworth, Engrg. Lab., Firestone Tire & Rubber Co., Akron 17, Ohio

**Information Theory**—*Chairman*, Louis A. DeRosa, Federal Telecommunications Lab., Inc., 500 Washington Avenue, Nutley, N. J.

**Instrumentation**—*Chairman*, Robert L. Sinck, Consolidated Engrg. Corp., 300 N. Sierra Madre Villa, Pasadena, Calif.

**Medical Electronics**—*Chairman*, Dr. Julia F. Herrick, Inst. of Experimental Medicine, Mayo Found., Rochester, Minn.

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**Ultrasonics Engineering**—*Chairman*, M. D. Fagen, Bell Labs., Whippany, N. J.

**Electron Devices**—*Chairman*, J. S. Saby, Electronics Laboratory, G. E. Co., Syracuse, N. Y.

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**Central Florida (3)**—Hans Scharla-Nielsen, Radiation Inc., P.O. Drawer 'Q', Melbourne, Fla.; G. F. Anderson, Radiation Inc., P.O. Box 'Q', Melbourne, Fla.

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**Detroit (4)**—N. D. Saigeon, 1544 Grant, Lincoln Park 25, Mich.; A. L. Coates, 1022 E. Sixth St., Royal Oak, Mich.

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**Emporium (4)**—E. H. Boden, R.D. 1, Emporium, Pa.; H. S. Hench, Jr., R.D. 2, Emporium, Pa.

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**Israel**—Franz Ollendorf, Box 910, Hebrew Inst. of Tech., Haifa, Israel; J. H. Halberstein, P.O.B. 1, Kiriath Mozkin, Israel

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*(Cont'd on next page)*

* Numerals in parentheses following Section designate Region number. First name designates Chairman, second name, Secretary.

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**San Antonio (6)**—C. M. Crain, Engineering Bldg. 149, University of Texas, Austin 12, Texas; W. H. Hartwig, Dept. of Elec. Engr., University of Texas, Austin 12, Texas

**San Diego (7)**—F. X. Brynes, 1759 Beryl St., San Diego 9, Calif.; R. T. Silberman, 4274 Middlesex Dr., San Diego, Calif.

**San Francisco (7)**—A. J. Morris, 1827 Cordilleras Rd., Redwood City, Calif.; J. S. McCullough, 1781 Willow St., San Jose 25, Calif.

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**Williamsport (4)**—J. E. Snook, 1629 Warren Ave., Williamsport, Pa.; F. T. Henry, 1345 Pennsylvania Ave., Williamsport, Pa.

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**Berkshire County (1)**—S. C. Leonard, Cheshire, Mass.; W. E. Neubert, 21 Highlawn Dr., Pittsfield, Mass.

**Buenaventura (7)**—E. C. Sternke, 320 Vista Del Mar, Camarillo, Calif.; Oliver La Plant, 325 North 'J' St., Oxnard, Calif.

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**Charleston (3)**—W. L. Schachte, 152 Grove St., Charleston, S. C.; Arthur Jones, 21 Madden Dr., Charleston Heights, S. C.

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**Fort Huachuca (2)**—Officers to be elected.

**Lancaster (3)**—G. W. Scott, Armstrong Cork Co., Lancaster, Pa.; G. E. Mandell, 522 E. King St., Lancaster, Pa.

**Mid-Hudson (2)**—E. J. Breiding, 54 S. Grand Ave., Poughkeepsie, N. Y.; E. S. Wilson, I.B.M., Poughkeepsie, N. Y.

**Monmouth (2)**—G. F. Senn, 81 Garden Rd., Little Silver, N. J.; C. A. Borgeson, 82 Garden Rd., Little Silver, N. J.

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**Palo Alto (7)**—John V. N. Granger, 320 Encinal Ave., Menlo Park, Calif.; W. B. Wholey, 342 Verano Dr., Los Altos, Calif.

**Pasadena (7)**—Officers to be elected.

**Richland (7)**—R. G. Clark, 1732 Howell, Richland, Washington; R. E. Connally, 515 Cottonwood Dr., Richland, Wis.

**Tucson (7)**—William V. Record, 4511 E. Ninth St., Tucson, Ariz.; A. J. Bersbach, 5326 E. Seventh St., Tucson, Ariz.

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**Wichita (6)**—H. G. Swift, 1642 South Main, Wichita, Kan.; Vernon N. Johnson, 1652 S. Edgemoor, Wichita 17, Kan.

# Abstracts of Transactions of the IRE

The following issues of "Transactions" have recently been published, and are now available from the Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N. Y. at the following prices. The contents of each issue and, where available, abstracts of technical papers are given below.

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## AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

### Vol. ANE-2, No. 1, March, 1955

**University of Dayton Honored**

Chairman's Report

**Trajectory Precision Requirements for Automatic Landing—J. L. Ryerson**

Requisite to landing an aircraft automatically is the perception of its altitude. This reports attempts to resolve the theoretical accuracy of an aircraft's altitude by standard statistical equations for the propagation of error.

Available aircraft trajectory data have been employed to determine the bandwidth of aircraft trajectories during approach and "landing." Such data have also been incorporated to establish the bandwidth of the noise superimposed by apparent radar target "wander."

A method of removing the noise by optimum filtering techniques, is discussed. A means of retrieving the high-frequency components of the aircraft trajectory information by inertial equipment is proposed. It is further proposed that similar smoothing techniques be utilized to remove low-frequency drift terms which appear as noise in the inertial equipment. The two sources of information are subsequently recombined to obtain broadband trajectory coverage having a greatly improved signal-to-noise ratio.

**Airborne UHF Communication Equipment—G. H. Scheer**

After more than two years' experience, it has been determined that airborne UHF communications is satisfactory. Many unforeseen types of interference have been found, some of them serious. Antenna patterns on aircraft are not ideal, but are usable. The newest subminiature airborne transceiver has undergone unique operational engineering tests. Results show that its design is highly desirable from the aspects of performance, installation and maintenance. Until the ultimate design of an equipment requiring no maintenance, industry should improve quality of components to increase reliability and reduce repairs.

**Flight Director Design Trends—G. Iddings and E. Martino**

The flexibility of the flight director to include practically any all-weather mission is accomplished by the coupling of the basic navigation data to the central flight director computer with special signal-shaping circuits designed on a modular basis. This paper describes the nature of these mission-coupling devices and an extension of their use to improve the basic flight panel information.

**Cooling Requirement Charts for Electronic Equipment—L. J. Lyons**

Efficient installation of airborne electronic equipment requires adequate data to design a heat removal system for the airplane. For many years aircraft generators have been supplied with rating charts to define their cooling requirements: electronic equipment now requires similar treatment. The form of the chart depends on the nature of the equipment, but consists, essentially, of a nomograph showing required cooling airflow and pressure drop vs air temperature. When pertinent, it should also include such factors as life, output, surrounding wall temperature, etc. Such charts make it convenient to design and compare performance of ram and fan driven cooling systems, and to compare different equipment designs in terms of their flow and pressure drop requirements. Examples of several different types of charts, their preparation and their use, are presented.

**The Accuracy of the VHF Omni-Range System of Aircraft Navigation: A Statistical Study—W. G. Anderson**

This paper describes a statistical treatment of the errors encountered in the VHF Visual Omni-Range Navigation System. The first part of the paper consists of a rudimentary discussion of statistical theory which serves to acquaint the reader with the methods used in the second portion of the report.

Each of ten different errors are described in terms of a normal distribution function with the parameters $\bar{x}$ and $s$, the mean and standard deviation respectively. The ten errors are summed up and are compared to the error distribution which was obtained from error data gathered by the pilots of various scheduled airlines. The close agreement justifies the method used.

Of the ten errors, only five are significant, and the VOR ground station error is shown to be greater than the sum of the other nine errors.

**Effect of Internal Fluctuations and Scanning on Clutter Attenuation in MTI Radar—G. S. Grisetti, M. M. Santa and G. M. Kirkpatrick**

The approximate effect of internal fluctuations and scanning on clutter attenuation in MTI radar is derived and the results presented on two charts. It is proposed that derivative antenna patterns be used to increase the clutter attenuation when scanning. The theoretical improvement from the use of derivatives is appreciable, as shown on the second chart.

## AUDIO

### Vol. AU-3, No. 3, May–June, 1955

**Complexity and Unreliability in Electronic Equipments—G. H. Scheer**

**Connecting Piezoelectric Pickups to Magnetic-Pickup Amplifiers—B. B. Bauer**

**IRE-PGA Election and Convention Summary.**

**Bereskin New Editor for IRE TRANSACTIONS on AUDIO**

**Formation of Syracuse IRE-PGA Chapter—W. W. Dean**

**Dayton IRE Section Organizes Audio Chapter—A. B. Henderson**

**PGA Chapter NEWS**

**Perceptibility of Flutter in Speech and Music—F. A. Comerci**

The perceptibility of flutter at various rates, in recordings of speech and music, was investigated in relation to the development of a flutter meter which will give a direct indication of the effect upon programs as judged by listeners.

**Sound Measurements at Very High Levels—Arnold Peterson**

The behavior of a number of microphones at high sound levels is described. Some of the problems encountered in making measurements at high sound levels are discussed.

**Electronic Organ Tone Radiation—D. W. Martin**

The principles of design for electronic organ tone chambers are outlined. The differences between the design goals for loudspeaker enclosures for organs and for other purposes are explained in fundamental terms. The construction of new organ tone cabinets for indirect radiation is described in detail. A few organ installation examples are given.

**PGA Institutional Listings**

## COMPONENT PARTS

### PGCP-3, March, 1955

**Report from the Chairman**

**The Effective Leakage Resistance of Several Types of Capacitors—R. W. Tucker and S. D. Breskend**

A rate-of-charge method for measuring the effective leakage resistance of good quality, high valued capacitors is described. This method yields results rapidly and directly. The effective leakage resistance of various types of capacitors as a function of time of applied voltage at different temperatures was measured. A method for calculating the change in capacitance with time of applied voltage is given. A polytetrafluoroethylene capacitor had the best direct-current properties of any of the types tested.

**Transformer Design Chart—Reuben Lee and N. E. Mullinix**

This paper describes a transformer design chart by which the design of two winding, sixty cycle, low voltage transformers can be made without most of the time-consuming design procedures. Its use is intended for transformer design engineers; therefore, it provides only the winding information that varies for each design. A specific series of cores and operating conditions are assumed. The turns, wire size and approximate winding resistance of both the primary and secondary can be determined from voltage requirements and secondary volt ampere rating. Equations upon which the chart is based are given together with an example problem. A set of rules for applying the chart to other transformer designs is also given.

**Problems Encountered and Procedures for Obtaining Short-Term Life Ratings on Resistors—W. T. Sackett, Jr.**

The purpose of this paper is to give an abbreviated review of Battelle Institute's activities in the electronic-component field, and to give a more detailed discussion of the particular phase of those activities having to do with the development of procedures for obtaining short-term life ratings on components.

**Subminiature Transformers and Their Application to Junction-Transistor Circuits—E. F. Dunkin and D. L. Johnson**

The technical limitations of subminiature audio or control frequency transformers or inductors are discussed and some features of their design described. Statistical methods are used to control quality and specification limits as the physical sizes are too small to allow for individual adjustment. A miniature toroidal-shell construction has given results comparing favorably with laminated assemblies below a certain size and in this form, optically-finished lapped joints are employed in the magnetic circuit. The use of subminiature core assemblies as transductors has also been investigated.

## ELECTRONIC COMPUTERS

### Vol. EC-4, No. 2, June, 1955

**PGEC Student Activities and Education in Computers—H. H. Goode**

**A Survey of Electronic Analog Computer Installations—L. B. Wade and A. W. Wortham**

A survey has been made of real-time electronic analog computer (differential analyzer) installations. This survey was conducted so that a directory of the installations could be compiled and so that various data regarding the installations could be made available for analysis. The survey was conducted by a mail questionnaire. Information was obtained regarding size of installation, size of staff, weekly usage of the equipment, age of installation, and availability to outside organizations from 96 installations having a total of 8,320 computer amplifiers. The results of the survey have been analyzed and are presented in this paper, together with the directory.

**A Digital Computer for Use in an Operational Flight Trainer—W. H. Dunn, C. Eldert, and P. V. Levonian**

The requirements for a digital computer for use in an operational flight trainer are presented with emphasis being placed on the real-time aspects of the problem. The general purpose digital computer is shown to be inadequate for this purpose and a special purpose digital computer is described which meets the requirements.

**A Diode Multiplexer for Analog Voltages—H. J. Gray, Jr., M. Rubinoff and J. Tompkins**

A diode multiplexer switch is described for time-sharing 64 analog voltages in a digital computer application. Apart from its relative simplicity and economy, the multiplexer characteristics of microsecond switching speeds, maximum settling time of 133 microseconds for a 10-volt operating range, and accuracies of better than 1 per cent full scale are confirmed both by theoretical equations and by experimental results.

**Some Notes on Logical Binary Counters—R. M. Brown**

The properties of binary counters which utilize non-transient storage elements for the count information are presented. The four possible sets of logical connections between the two storage elements necessary for each stage are described. The binary numbers represented in the storage elements are shown to be the actual count in one set of elements and in the other set a Gray code representation of twice the actual count. Examples of bi-directional counters are given.

**A Variable Binary Scaler—D. B. Murray**

The binary elements of a counter or "scaler" may be interconnected in many ways. This paper discusses a class of interconnections in which some elements are "forward-counting" and some are "reverse-counting." By changing the interconnections any arbitrary integral scaling ratio (up to the counter capacity) may be obtained.

**Time-Delay Circuits—W. E. Thomson**

**Contributors**

**News: S. B. Disson**

**Reviews: Reviews of Current Literature—H. D. Husky, ed.**

## ENGINEERING MANAGEMENT

### PGEM-3, March, 1955

**More Engineering Per Dollar—Burgess Dempster**

**Some Factors Related to Management of an Applied Research Project—Harley Iams**

Among the factors important to the management of an applied research project are the arrangement of working areas, the provision of labor-saving supporting equipment, and the supplying of adequate shops for making experimental parts. But even more important is the management of the people, including their organization into effective teams, seeing that engineers and scientists are retained for their best contribution to national defense, and upgrading the abilities of the staff.

**A Practical Approach Toward Integration of Project and Group Theories in Establishing an Engineering Organization—C. F. Horne**

**Engineering Management and the Changing World—Maurice Nelles**

**Are Engineers People?—A. M. Zarem**

**Biographical Notes on the Authors**

## INDUSTRIAL ELECTRONICS

### PGIE-2, March, 1955

**Mutual Problems in Industrial Electronics and Communications—E. W. Allen**

**Numerical Control of Machine Tools—Leroy U. C. Kelling**

Industry needs more flexible methods of programming machine cycles to achieve automatic operation of machine tools in limited-quantity production. Part of this need is met by numerical control systems which command the machine in accordance with prepared numerical instructions read from a storage medium. Such numerical instruction can be stored in binary or decimal numerical form on punched telegraph tape, punched tabulating cards, magnetic tape, and many other data storage mediums. Numerical control systems are well adapted to control of machine tools such as lathes, turret punch presses and boring, drilling and milling machines. The numerical control system for a Wiedemann turret punch press is a typical example illustrating the problems of joining controls and machines into a smoothly working combination.

**Electromechanically Stabilized DC Amplifier for Use in Transducing and Telemetering of Milli-Voltage and Micro-Ampere Signals—Hubert A. Riester, Jr.**

**Industrial Applications of X-Ray Techniques—T. H. Rogers**

**The Application of Radioactivity to Measurement and Control—Norman E. Walters**

**A Magnetic Thickness Gage for Rubber and Plastic Applications—Albert M. Dexter**

**Advantages of Electronic Process Control—C. E. Mathewson**

Contrary to measurement requirements, the modern concept of closed loop, or feedback, control has created an urgent requirement for high speed response. The pure dead times and variable exponential lags characteristic of pneumatic communication systems have limited control performance to an appreciable extent. Electrical communication is the obvious solution and equipment designed for its utilization will be described herewith.

**Electronically Produced and Controlled Illumination—Harold E. Edgerton**

After a brief summary of the theory of electronically produced pulses of light, energy storage systems, and energy converting devices, a review is made of commonly used triggering and controlling elements. A series of practical devices utilizing the previously described methods are discussed, showing typical stroboscopes, high-speed single-flash photographic lights, high-speed motion-picture lights, and flash sequence equipment.

**Automatic Detection of "Green-Rot" in Shell Eggs—K. H. Norris**

**Power Oscillators for Dielectric Heating—T. L. Wilson**

**Survey of Today's Use of Power Rectifiers in Industry—L. W. Morton**

**Electronic Considerations in the Theory and Design of Electric Spark Machine Tools—E. M. Williams and J. B. Woodford, Jr.**

## INFORMATION THEORY

### Vol. IT-1, No. 1, March, 1955

**An Analysis of the Detection of Repeated Signals in Noise by Binary Integration—J. V. Harrington**

An analysis of the detection of repetitive signals in noise by binary integration techniques is made. An expression for the effective signal-to-noise ratio of the quantized video is obtained and is shown to apply to any half-wave second detector. A comparison of analog and digital integration is made, and it is further shown that digital integration is, at most, 1.9 db poorer due to the quantization loss. However, the loss due to nonideal analog integration can make the two types equivalent. The optimum settings for quantizer and counter thresholds are derived, and expressions for the final-detection and false-alarm probabilities are determined. Lastly, the results are modified to include the effect of nonuniform amplitudes in the set of signals being quantized and integrated.

**An Expansion for Some Second-Order Probability Distributions and its Applications to Noise Problems—J. F. Barrett and D. G. Lampard**

In this paper it is shown that, in general, second-order probability distributions may be expanded in a certain double series involving orthogonal polynomials associated with the corresponding first-order probability distributions. Attention is restricted to those second-order probability distributions which lead to a "diagonal" form for this expansion.

When such distributions are joint probability distributions for samples taken from a pair of time series, some interesting results can be demonstrated. For example, it is shown that if one of the time series undergoes an amplitude distortion in a time-varying "instantaneous" nonlinear device, the covariance function after distortion is simply proportional to that before distortion.

Some simple results concerning conditional expectations are given and an extension of a theorem, due to Doob, on stationary Markov process is presented.

The relation between the "diagonal" expansion used in this paper and the Mercer expansion of the kernel of a certain linear homogeneous integral equation, is pointed out and in conclusion explicit expansions are given for three specific examples.

**Predictive Coding—Peter Elias**

Predictive coding is a procedure for transmitting messages which are sequences of magnitudes. In this coding method, the transmitter and the receiver store past message terms, and from them estimate the value of the next message term. The transmitter transmits, not the message term, but the difference between it and its predicted value. At the receiver this error term is added to the receiver prediction to reproduce the message term. This procedure is defined and messages, prediction, entropy and ideal coding are discussed to provide a basis for Part II, which will give the mathematical criterion for the best predictor for use in the predictive coding of particular messages, will give examples of such messages, and will show that the error term which is transmitted in predictive coding may always be coded efficiently.

**The Linear, Input-Controlled, Variable-Pass Network—B. E. Keiser**

This paper describes the study and development of a linear, variable-pass network system which is controlled by the Fano short-time autocorrelation function of the input. Given an input function, the message, whose short-time power spectrum varies in an unpredictable manner with time, and to which there has been added a different function, the disturbance, whose short-time power spectrum is either time-invariant or varies in a completely known manner, a linear, input-controlled, variable-pass network can be specified which minimizes the mean-square error between the message input and the total output, taking network delay into account. Methods for mathematical computation of the mean-square error have been devised.

The linear, input-controlled, variable-pass network has been found to have a lower mean-square error than that attainable with an optimum-mean-square, linear, fixed, selective network, for certain types of input messages.

**Spectral Density Functions in Pulse Time Modulation—H. Kaufman and E. H. King**

Spectral power density functions corresponding to various types of pulse shapes, probability distribution functions arising in the study of pulse time modulation problems are computed. The results are presented in tabular form. The following cases are considered: PAM and PPM, for arbitrary pulse shape, PDM, for rectangular, Gaussian, and error-function pulse shapes, and SEM, for rectangular pulse shape.

**A Note on the Sampling Theorem—L. J. Fogel**

The human operator often perceives rate as well as amplitude information in sampling various displayed continuous parameters. It is therefore necessary to extend the Sampling Theorem to allow the analysis of certain man-machine relations. The result is stated and the required mathematics included in the appendix. Certain distinct problem areas where this extension can be fruitfully employed are indicated.

**Statistical Calculation of Word Entropies for Four Western Languages—G. A. Barnard**

Using a modified version of Shannon's method, comparative figures for the word-letter entropies of printed English, French, German, and Spanish are obtained and the method described.

**Papers presented at WESCON in Los Angeles, August 25–27, 1954**

**On the Modulation Levels in a Frequency Multiplexed Communication System by Statistical Methods—R. L. Brock and R. C. McCarty.**

This paper presents a mathematical analysis with experimental verification of the distribution of the instantaneous voltage of a complex signal resulting from the combination at random of a small number $n$ of sinusoidal oscillations. The resulting calculated distributions are plotted in the form of a set of probability curves for comparison with curves obtained by experiment. Further laboratory measurements in which the individual sinusoidal oscillators are frequency-modulated in a manner suitable for communicating information in a binary form, yield substantially no change in the amplitude distribution as determined for the unmodulated oscillators. Consequently, the results of the mathematical analysis may be applied in the determination of $M$, the degree to which each subcarrier may amplitude modulate a final carrier in an fm-am frequency multiplexed system. $M$ may be determined for any desired degree of overmodulation in excess of one per cent and for as many subcarriers as are required in the system. Modulation levels determined according to approximate methods and the method described here are tabulated and compared.

**On the Response of a Certain Class of Systems to Random Inputs—Jack Heilfron**

This paper deals with the connection between vector Markoff processes and the response of a lumped constant parameter linear system composed of a finite number of elements. It was known that if a Gaussian process which is one component of a vector Markoff process passes through such a system, the result is also Gaussian and may be considered as one component of a higher dimensional vector Markoff process. We show that the term Gaussian may be excluded in the above statement. The practical importance of this result is that if one can determine the initial and transition probabilities of this vector Markoff process, one can also determine the complete statistical properties of the output of the system. This further implies that the determination of the properties of the output for the class of not necessarily Gaussian inputs mentioned above is not as difficult as might be expected from the results for just the first probability distribution for non-Gaussian inputs.

**Noise in Driven Systems—J. M. Richardson**

It is known that a direct relation exists between the noise in a system in equilibrium and transient drift toward equilibrium. It seems that a similar relation should exist for a system in a nonequilibrium stationary state. It is now necessary to distinguish between two types of transients; those produced by selecting those systems satisfying certain initial and those produced by actual physical perturbation. It is is shown that a simple relation exists between noise and the transients produced by selection and that no relation exists in the case of transients produced by perturbation. In the equilibrium case it is shown that the two types of transients, though still logically and operationally distinct, can be described by the same impedance operator.

**Design and Performance of Phase-Lock Circuits Capable of Near-Optimum Performance Over a Wide Range of Input Signal and Noise Levels—R. Jaffe and E. Rechtin**

## MICROWAVE THEORY AND TECHNIQUES

Vol. MTT-3, No. 3, April, 1955

**Editorial Comment by Saad**

**Frontispiece of W. W. Mumford**

**Editorial—W. W. Mumford**

**Advances in Microwave Theory and Techniques—D. D. King**

**Planar Transmission Lines—David Park**

This paper derives formulas for the transmission properties—characteristic impedance and attenuation—in the principal mode of a transmission line consisting of one or two long strips of metal foil embedded in a dielectric material between two long metal strips considerably wider than the central ones. The width and spacing of the central strips is arbitrary, and it is also necessary to take account of their thickness in computing the attenuation. A graphical method is given for evaluating the characteristic impedance in general, and analytic approximations are given for a number of special cases. Finally the question of the leakage of power from between the outer strips is considered briefly.

**Measurement of Time-Quadrature Components of Microwave Signals—J. H. Richmond**

A phase-sensitive coherent detector used for microwave laboratory measurements is described. The receiver measures the real ($|E| \cos \alpha$) and imaginary ($|E| \sin \alpha$) components of a signal $E$ with equipment which is less elaborate than that required for measuring the amplitude $|E|$ and phase $\alpha$. Furthermore, many calculations are more convenient if $E$ is presented in rectangular rather than polar form.

Measurements made with the receiver on known fields in waveguides are included to demonstrate its accuracy. The receiver has a sensitivity of $-125$ dbw at 9,375 mc.

**Optimum Design of Stepped Transmission-Line Transformers—S. D. Cohn**

This paper describes the optimum stepped-transmission-line transformer structure for matching two unequal characteristic impedances. For any specified bandwidth, the steps are designed to yield a Tchebycheff-type (or equal-ripple) reflection-coefficient response. Over this band, the maximum vswr is less than that obtainable with any other stepped-transformer having the same number of steps. The design method and the technique for eliminating discontinuity-capacitance effects are given. The measured results for a coaxial and a waveguide model are presented and found to verify the method.

**The Use of Scattering Matrixes in Microwave Circuits—E. W. Matthews**

Difficulties arising from the use of the impedance concept in microwave circuitry have led to the introduction of the scattering representation for work at these frequencies. This paper presents a development of the scattering approach in terms of fundamental transmission-line phenomena. The physical meaning of the quantities involved is brought out wherever possible and the relationships among the various elements of the scattering matrix are given. Several examples of the application of scattering techniques to analysis of the properties of microwave junctions are presented, and methods for measuring scattering parameters of such junctions are outlined.

**Some Applications and Characteristics of Ferrite at Wavelengths of 0.87 Cm and 1.9 Cms—Clyde Stewart**

This paper describes the use of ferrites in waveguide to produce Faraday rotations at 0.87 cms and 1.9 cms wavelengths. The Dicke radiometric receiver is briefly reviewed and its improvement by the use of ferrite waveguide components is described. Experimental equipment for securing data on the behavior of ferrites is discussed. Details are given for the construction of a unidirectional waveguide transmission line for 0.87 cms wavelength.

**Probes for Microwave Near-Field Measurements—J. H. Richmond and T. E. Tice**

To be satisfactory for microwave near-field measurements, a probe must have desirable polarization characteristics, must have an aperture small enough to indicate the field at a point, must deliver sufficient signal voltage to permit accurate measurement, and yet must not seriously distort the fields. The design of a probe may be simplified if the fields to be measured are known to be almost linearly polarized or to consist only of a traveling wave. Comparison of measurements made with various probes has led to the development of a small open-ended waveguide probe which is simple to construct and has given excellent results.

**Measurement Techniques for Multimode Waveguides—A. C. Beck**

This paper surveys some of the techniques that have been worked out for multimode waveguide measurements. Equipment has been developed for measuring one mode at a time by taking advantage of the differences between the modes. Illustrations of its use are given.

**Reports of International Organizations—L. G. Cumming and W. W. Mumford**

**Addenda to "Bibliography on Directional Couplers"—R. G. Medhurst and R. F. Schwartz**

## MICROWAVE THEORY AND TECHNIQUES

### Vol. MTT-3, No. 2, March, 1955

**Microwave Printed Circuit—An Historical Survey—R. M. Barrett**

The microwave printed circuit, as described in this paper, is an extension of the well-known technique which is of such importance in the lower frequency regions, where lumped element circuits are practical. This new circuit possesses all of the virtues of other printed circuits, such as light weight, cheapness, ease of manufacture, miniaturization, etc., along with the ability to be used at frequencies as high as 10,000 mc. The basis of the new technique is the planar or "flat strip" coaxial transmission system which was developed during World War II but which has remained unpublished and relatively unknown in the postwar period; and for which an adequate theoretical analysis had not been available.

**Microwave Strip Circuit Research at Tufts College—A. D. Frost and C. R. Mingins**

The research work on microwave strip circuits which has been in progress since November, 1952 is described. Experimental investigations have included measurement of the characteristic impedance of various lines, the design of transitions from coaxial connectors to stripline, the preparation and adjustment of matched resistive terminations, and most recently the evaluation of the effects of simple line discontinuities such as bends or steps. Theoretical studies on the calculation of characteristic impedance and line loss have also been carried out.

**Characteristics and Some Applications of Stripline Components—W. E. Fromm**

Basic characteristics of Stripline in various frequency bands from 1000 to 16,000 mc are summarized. Various components such as transitions to coaxial line, attenuators, hybrid rings, directional couplers, and filters are shown. Some applications of these components in practical high performance microwave circuits and equipment in the frequency range of 2500–10,000 mc are also described.

**Photoetched Microwave Transmission Lines—Norman R. Wild**

Microwave transmission line and components of unusual light weight and compact construction can be made employing photoetching techniques to produce strip type transmission line. This report will be a general description of work done at Sanders Associates, Inc., to develop techniques for the design and manufacture of photoetched microwave transmission lines. Discussion will include measurements of attenuation and radiation leakage on parallel plate strip lines, as well as shielded type Tri-plate lines, the problem of mode purity and its relation to electrical parameters, various schemes of making transitions from standard waveguide to photoetched strip line. The basic design and performance of various components, as well as items of test equipment, such as slotted lines, matched loads, fixed attenuators, variable attenuators, directional couplers, crystal holders, phase shifters, hybrid rings, coax to Tri-plate transistors, etc., will also be treated. In addition, data will be presented showing impedance and susceptance values of simple discontinuities and impedance matching transformers. A simple technique for constructing gyrators and resonators will be presented, and the design and fabrication of an S-band signal generator employing photoetched microwave Tri-plate line will be shown, illustrating that practical microwave systems can be constructed far more economically than would be possible utilizing conventional waveguide techniques.

**Characteristics and Applications of Microstrip for Microwave Wiring—M. Arditi**

The experimental results of the transmission properties of Microstrip are compared with the values to be expected from a first order theory based on the assumption of a T.E.M. mode of propagation. The characteristics of various Microstrip components are given. These components include: waveguide or coaxial transducers, hybrids, directional couplers, crystal mixers, attenuators, filters, ferrite modulators, gas discharge modulator tubes and a wide-band noise source. The design considerations stress the wide-band properties of Microstrip similar to those found in coaxial lines.

The methods of measurement used in Microstrip are outlined and they show the simplicity of the experimental set-up required for the application of Deschamps' method for determining the principal characteristics.

The applications of Microstrip to the design of complete systems such as microwave receivers are discussed and examples in "S" band, "C" band and "X" band are given.

**Miniature Strip Transmission Line for Microwave Applications—E. N. Torgow and J. W. E. Griesmann**

The construction of a strip line whose physical size is kept as small as possible consistent with reasonable electrical performance is presented. This line is fabricated by relatively simple techniques and can be shaped to fit line components into relatively confined spaces. The line has good power handling capacity and moderately low attentuation. Various components have been developed in this line, including a broadband $\frac{3}{8}$-inch coaxial line to strip line adapter, a broadband matched load, attenuators, and high and low pass filters.

**Strip Type Components for 2000 Mc Receiver Head End—K. E. Zubulin**

Recent experimental work has evolved some components using air-spaced strip type transmission line that have been used successfully in connection with a variable attenuator, cavity and crystal mixer. Bandwidth, VSWR, and NF measurements are comparable with a commercial receiver head-end presently in use. The asymmetric air-spaced strip-above-ground transmission line used results in a simple configuration for coupling the line to the cavity. It also facilitates the application of a variable attenuator using a ferrite slab of high attenuation per unit length with good VSWR properties.

**Properties of Dielectric Image Lines—D. D. King**

The properties of a dielectric rod on an image surface are reviewed, and experimental results on straight sections, various bends, and a twist are presented. Techniques for measuring insertion parameters and field distributions are described.

**Practical Dielectric-Filled Waveguide—T. N. Anderson**

This paper describes the development of a laminated teflon-filled dielectric waveguide using techniques similar to what has been done in the flexible coaxial line. This paper describes the development of dielectric waveguide giving the theoretical design of the teflon-filled dielectric waveguide from both a mechanical and electrical point of view.

The emphasis on this dielectric waveguide development has been to arrive at a practical waveguide construction which would be suitable for a radar systems application. This dielectric waveguide is intended to provide a miniaturized waveguide circuit which will have essentially the same peak power handling capabilities as standard waveguide which would be suitable for use up to 200° C.

The fabrication technique is described along with a description of the measurement procedure for determining the characteristics of this dielectric waveguide including match, attenuation and high power breakdown.

The design of special transitions from air-filled waveguide to dielectric-filled waveguide are described also.

This work was performed under contract number AF33(600) 26763 for Wright Air Development Center and is intended to eventually yield a series of dielectric-filled waveguides, both rigid and flexible.

**Measurement of Attenuation and Phase Velocity of Various Laminate Materials at L-Band—M. E. Rigenbach and H. W. Cooper**

Measured data are plotted for the characteristic impedance, velocity of propagation, and attenuation of dielectric sheet supported strip transmission lines for four dielectric materials: Teflon bonded glass cloth, epoxy bonded glass cloth, polyester bonded glass mat, and XXXP paper base phenolic. At 1000 megacycles, the teflon material is excellent and the epoxy and polyester materials satisfactory for low Q applications, such as microwave transmission lines.

The equivalent physical length of a dielectric sheet supported strip transmission line right angle is reported.

**The Input and Mutual Impedance of Dipole Strips Between Parallel Planes—W. H. Hayt, Jr.**

A center-fed filamentary dipole is parallel to and between two parallel, infinite, perfectly conducting planes and carries a sinusoidal current. The longitudinal electric field intensity corresponding to such a current distribution is then obtained by an application of the image principle to the field of a single center-fed filamentary dipole in free space.

This longitudinal electric field is then used directly to obtain the input impedance and mutual impedance between filamentary dipoles of resonant lengths by the induced emf method. The impedances appear as an infinite series of integrals which are approximated by simple expressions having errors of less than one-quarter of one per cent. Curves are obtained giving the input impedance of dipoles having various resonant lengths and locations between the guard planes, and for several separations of the guard planes, the latter value being maintained less than one-half wavelength to avoid any propagating modes. Mutual impedance is shown as a function of dipole separations as well.

The results are then extended to dipoles having a cross-section which is a circle or a zero-thickness strip. Curves are obtained for the input impedance of strips between parallel planes for several plane separations and several dipole widths, as a function of dipole length.

**Problems in Strip Transmission Lines—S. B. Cohn**

A review is given of characteristic-impedance formulas for shielded-strip transmission lines. From these formulas, a set of approximate relationships for the attenuation and Q of a dielectric-filled shielded-strip transmission line is derived. The method makes the standard assumption that the current distribution is that of a lossless line and the surface resistivity that of an infinite-plane conductor. Although this method applies accurately to most other types of lines, in this case, an error of the order of 10% is believed to occur due to the failure of the assumptions at the corners of the strip. However, the error is in a direction that makes the computed values conservative, and the accuracy should be sufficient for most practical purposes. The derivation of a correction term is now being attempted.

In addition to the discussion of attenuation, attention is given in this paper to the design considerations involved in a shielded-strip-line impedance meter, and to some preliminary data obtained with this device. Also, the future topics for investigation under this research and development program are mentioned.

**Equivalent Circuits for Discontinuities in Balanced Strip Transmission Line—A. A. Oliner**

Theoretical formulas are derived for the equivalent circuit parameters of a variety of discontinuities in balanced strip transmission line. These formulas are simple in form and are obtained by employing a small aperture procedure or a Babinet equivalence procedure in conjunction with an approximate model of the line. The results for a number of discontinuities are presented and comparison is made with the available measured data.

**A Universal Approximate Formula for Characteristic Impedance of Strip Transmission Lines with Rectangular Inner Conductors—R. L. Pease and C. R. Mingins**

An explicit expression is developed for the characteristic impedance of a microwave strip transmission line with rectangular inner conductor of arbitrary dimensions. The expression is exact for zero thickness and arbitrary width, exact for zero width and arbitrary thickness, and quite accurate (within 3% for the extreme case of a square inner conductor of dimensions about 0.01 of plate separation, but in most cases of practical interest, within 0.1%) throughout the entire range of thickness and width.

**Stripline Radiators—E. G. Fubini**

Progress on the use of strip conductors as microwave antennas indicates that the technique is flexible and economical. Broadside curtains can be fabricated with sufficient accuracy. Several types of balanced Stripline feed have been considered, and twists have been successfully built. A variety of baluns have been evaluated and used to feed colinear Franklin arrays through binary splits.

**Slot Array Employing Photoetched Tri-Plate Transmission Lines—D. J. Sommers**

Microwave printed circuit techniques are readily adapted to the construction of compact antennas ideal for flush mounting on high speed aircraft. This paper describes the development of a two-dimensional X-band array consisting of 16 slots fed by photoetched Tri-plate transmission line. The design of a unity coupled series slot and the resulting mode purity problems are discussed. Several power divider configurations are illustrated and data on the performance of some of these devices is presented. The construction of a 4 slot E-plane, a 4 slot H-plane and the combination 4×4 E-, H-plane array utilizing these power dividers is shown. Radiation patterns of each of these arrays were measured and a comparison of the individual and combination array patterns is made.

**Bandpass Filters Using Stripline Techniques—D. R. White and E. H. Bradley**

Strip lines provide a convenient transmission medium for the realization of microwave filters. Since bandpass filters designed in waveguide and coaxial lines would be large at ultra-high frequencies, strip lines afford a practical means of realizing filters which are simply fabricated, are readily reproduced, and, in most cases, represent an appreciable savings in size and weight. Of the different types of strip transmission lines currently in use, the so-called "sandwich" structure has been employed at Melpar for two reasons: (1) very broad-band coax-to-strip-line transitions are easily realized; and (2) the electromagnetic field is essentially confined between the two ground planes, thus reducing problems in packaging.

Using design techniques developed for direct-coupled cavity-type waveguide and coaxial filters, experimental strip-line filters having ten per cent bandwidths in the u-h-f spectrum have been developed. These units have less than 1 db mid-band insertion loss and provide a rejection of greater than 40 db at frequencies twelve per cent from the center frequency. The design techniques discussed in this paper are general and, therefore, are not restricted to the realization of the above filter characteristics. Some limitations pertaining to the realizability of the cavity parameters in different dielectric media and the existance of spurious responses are discussed.

**Resonator and Preselector in Stripline—J. F. Moore and Max Michelson**

One of Raytheon's commercial applications involves microwave circuitry in balanced strip line, with a $\frac{3}{8}$-inch spacing between ground planes. Though the less critical parts of the unit are etched in copper-clad Teflon-Fibreglas, two of the components are of higher Q than can reliably be obtained in the presence of plastic. They are: (1) A resonator, for use as a frequency stabilizing reference element in an AFC circuit; and (2) A four-stage maximally flat preselector filter. These units are made of separate strips of metal, and do not depend on the plastic sheet for their support.

The unit is intended to operate over a $4\frac{1}{2}$% band centered on 6725 mc, and was designed to avoid expensive parts and assemblies.

The design requirements were for a resonator with an unloaded Q of 1980, and with no more than ±0.45 mc frequency variation over operational extremes of temperature and humidity; and for a preselector with less than 3.5 db. loss and a 30 mc pass band. Both units satisfy the overall electrical and mechanical design without requiring special high-cost structures. In fact, the microwave head is, in balanced strip line, about half as expensive as in conventional plombing. In addition, the present cost will be further reduced as larger quantities are considered.

**Broad-Band Microstrip Crystal Mixer with Integral DC Return—Eric Carlson**

A light and compact microwave mixer using microstrip has been designed for use in airborne equipment. The mixer features a low-input voltage standing wave ratio over a frequency range of one octave. The local oscillator is coupled to the input transmission line by a microstrip directional coupler having an integral d-c return. Minimum coupling to the d-c return is obtained by placing it in a region of low field intensity.

## RELIABILITY AND QUALITY CONTROL

### PGRQC-5, April, 1955

**Statistical Design—A Means to Better Products of Lower Cost—R. C. Miles**

Present evidence suggests that electronic reliability problems are being attacked with too little emphasis on the basic problem, which is one of *equipment* rather than merely component reliability. Among the reasons for this situation are: lack of enforcement of equipment reliability requirements, resulting in part from difficulty of enforcement; vague or unrealistic statement of equipment reliability requirements; lack of an adequate quantitative basis for predicting the reliability of a proposed equipment design.

The popular concept of "guaranteed" reliability is basically a fallacy, since reliability cannot be positively guaranteed in any useful sense. If the guarantee concept were valid, an equipment using only "guaranteed" components should have a life at least equal to the shortest guaranteed component life; that such is not the case in practice proves the fallacy of reliability "guarantees."

A more practical concept of reliability involves the formulation of a statistical definition. For example, reliability may be defined as "the probability that a component or equipment will operate satisfactorily under given circumstances," time of operation being included as one of the "circumstances."

Although such a definition makes it possible to relate equipment reliability to the reliabilities of the individual components, presently available component data is not suitable for the purpose. In particular, most component reliability data is deficient as regards the variation of reliability with operating time. Even such component data as does exist applies to operating times at which the component reliability has become intolerably poor in terms of the requirements of equipment of even moderate complexity.

It appears further that the majority of existing component improvement programs will do little to improve the situation. Such programs seem to have been conceived on an unrealistic basis, concentrating on improving reliability at operating times near the end of the component useful life, rather than maintaining very high values of component reliability for as long as possible in the interests of improved equipment reliability.

Substantial progress toward more reliable electronic equipment requires a combination of:

1. A realistic quantitative basis for equipment reliability requirements.
2. Education of equipment and component designers and users as to the basic nature of the reliability problem.
3. Better data on the reliability of existing components, in addition to development of improved components.
4. Closer cooperation between component and equipment engineers, in order that each may acquire a better understanding of the other's needs and problems.

**Contributing Factors to Component Parts Reliability and Extended Service—J. A. Goetz**

Reliable electronic equipment performance depends fundamentally upon a sound application of engineering data and service information. Essential elements of a program tailored to this need by the equipment manufacturer include:

1. Maintenance of a realistic field evaluation program on component parts and assemblies thereof;
2. A coordinated engineering liaison program between consumer and vendor of component parts;
3. A sound source qualification and parts improvement program;
4. Development of adequate specifications and application data covering extended life applications of component parts.

These elements are discussed as they apply to the current manufacture of electronic accounting and data processing machines by IBM.

**Acceptance Sampling of Reliable Tubes—B. P. Goldsmith**

The traditional method of checking acceptability of a lot of tubes for a particular electrical characteristic has been to test a large sample and count the number of tubes beyond the minimum or maximum limits—inspection by attributes. This reduces the chances of accepting a lot with a high percentage of defectives, but gives no assurance that the lot is centered close enough to bogey or that the spread of the distribution is tight enough.

The increasing complexity of circuitry and the high standards of performance required in many types of equipment have increased the need for such assurance. It is gained most simply from inspection by variables.

Simply adding variables inspection criteria to the minimum and maximum limits already on the TSS will not do an efficient job of separating good lots from bad. By proper coordination of the two types of inspection, attributes and variables, a high degree of discrimination can be achieved with a modest amount of testing.

Examples are based on recent TSS for type JAN 5744WA.

**Cathode Interface Impedance Desimplified—H. B. Frost**

Cathode interface impedance has usually been treated as if it could be represented at any given time by a parallel R-C combination. In actuality, however, the impedance can be represented accurately only by an R-C network containing four elements. Moreover, the interface undergoes a reversible change of state with a relaxation time near one second as the cathode current is changed. These characteristics merit important consideration when specifications concerning cathode interface impedance are written.

Modern methods of preventative maintenance frequently allow replacement of those tubes with cathode interface impedance before they can cause the failure of large electronic systems such as digital computers. However, serious cases of cathode interface impedance may cause the tube population of such a system to have a short average life, perhaps less than 10,000 hours. Under such conditions, catastrophic failures—primarily an affliction of young tubes—very likely will be more prevalent than would be the case if the average life were greater. Any increased level of catastrophic failures will cause a reduction system reliability which may be attributed indirectly to cathode interface impedance.

**The Definition of Terms of Interest in the Study of Reliability—C. R. Knight, E. R. Jervis, and G. R Herd**

The aim of this paper is to propose certain concepts and definitions as aids in studies of the reliability of various products. "Reliability" and other terms commonly used in such studies are so defined that they can be measured and expressed quantitatively; and the theoretical relationship of components to the system is discussed. Reliability is studied in terms of discrete variables and continuous variables and their combined effects, with consideration of the interdependence of components. The concept of dependence is developed to facilitate measurement of the effectiveness with which components are incorporated into a system. The paper advances an alternate definition of "satisfactory performance" to the generally accepted one based on current specification practices. The new definition takes into account user acceptance or rejection of the product. Weighting functions are proposed to give mathematical expression to user opinion versus equipment performance characteristics.

## ULTRASONICS ENGINEERING

### PGUE-2, May, 1955

**Composite Piezoelectric Resonator—W. G. Cady**

Various types of composite resonators and their uses are summarized. The general equations are given for the transducer for plane ultrasonic waves, consisting of a crystal assembly with back and front plates. Applications are made to several simple cases, and expressions are tabulated for the amplitude of vibration for various combinations of half- and quarter-wave components. Considerat on is given to the effect of the acoustic load on the frequency for maximal amplitude. Theoretical formulas are compared with experimental results for rods of aluminum and fused quartz excited in vibration by piezoelectric crystals.

**Ultrasonic Cleaning of Miniature Devices—Q. C. KcKenna**

Ultrasonic cleaning gives industry a new method of obtaining cleaning results previously unattained. By irradiating liquid cleaners with appropriately arranged transducers, large volumes of intricate parts can be cleaned. Barium titanate ceramic transducers offer many advantages as sound generating elements. They can be operated at low voltages compared with quartz and can be cast in shapes which give high ultrasonic intensities. Through focusing, the ultrasonic cleaning process usually results in a more economical method, saving time, labor, and space.

**Power Measurements in Ultrasonics—O. E. Mattiat**

**A Temperature Invariant Solid Ultrasonic Delay Line—Edwin Voznak and R. W. Mebs**

A study was made of various metals and alloys in an effort to obtain a solid ultrasonic delay line that would be thermally stable with respect to time delay. Experimental data showing the effect of temperature on the propagation of ultrasonic waves are presented. An isoelastic alloy possessing a temperature coefficient of delay time of not more than 8 ppm/°C over a temperature range of $-50$ to 200°C is described. This characteristic is superior to that of quartz or mercury by an order of magnitude.

**Some Applications of the Linear Piezoelectric Equations of State—Rudolf Bechmann**

**Notes on the Uses of Ultrasonics for the Finishing of Cathode Ray Tube Guns and Gun Components—W. F. Niklas**

**Biographical Notes on the Authors**

**Letter to the Editor**

**Recent Books on Ultrasonics**

**Cross Index IRE-PGUE Transactions 1-3**

# Books

### Transistor Audio Amplifiers by Richard F. Shea

Published (1955) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 207 pages+xiii pages+5 page index+5 page bibliography. Illustrated. 9¼×6. $6.50.

The stated object of this book is to provide the practical fundamentals of transistor applications and to show how these fundamentals may be used in the construction of audio amplifiers. In the reviewer's opinion the author has fulfilled his purpose competently. The book is a useful, lucid compilation of junction transistor circuit fundamentals, typical data on commercial units, design formulas, and practical audio circuitry.

Intended primarily for the experimenter and the designer of practical circuitry, the book does not treat transistor physics or technology and uses a minimum of mathematics throughout. The emphasis is definitely on presentation of facts and figures rather than detailed whys and wherefores. One-third of the text is devoted to transistor characteristics and parameters, including the relationships among the various equivalent parameters in current usage, and citing a good many commercial transistor specifications. One of the eight chapters gives a full treatment of the three basic amplifier configurations, complete with expressions for input and output resistances, and various gain figures under matched as well as unmatched conditions. Numerous graphs are included, which show at a glance how gains and input and output resistances vary with source and load resistances, supply voltages, currents, and temperature. The remaining chapters are devoted to coupled stages, with a detailed comparison of the various possible combinations of the three basic structures; preamplifiers, including, for example, consideration of transistor noise figure; Class A and B power amplifiers, with considerable number of transfer-characteristic graphs comparing departure from linearity under various conditions; and, finally, a few examples of hearing-aid and phonograph amplifier circuits, including the design steps leading to these circuits.

*Transistor Audio Amplifiers* covers a fairly wide range of material in relatively few and small-size pages; the treatment is therefore necessarily brief, and a number of points are brushed over rather lightly. Aside from this, and aside from a few inconsequential errors, *Transistor Audio Amplifiers* can well be recommended to those interested in becoming familiar with this increasingly important subject.

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### Sonics by T. F. Hueter and R. H. Bolt

Published (1955) by John Wiley and Sons, Inc., 440 Fourth Ave., New York 16, N. Y. 440 pages+xi pages+15 page index. Illustrated. 9¼×6. $10.00.

The announced purpose of this book is to provide a treatment of sonics—defined as the technology of sound as applied to problems of measurement, control, and processing—to serve the needs of the physicist as well as the practical design engineer. In this purpose the book has succeeded admirably. An adequate discussion is given of the fundamental acoustical principles, the properties of transducers—principally piezoelectric and magnetrostrictive—and their applications in such practical applications as drilling, cleaning, sonic processing of metals, liquids and gases, and ultrasonic inspection of materials. The point is made that there is no frequency division into audible and inaudible sound for many of the processes. Thus, drilling of brittle material by an ultrasonic drill is similar in principle to the drilling of oil wells by much larger units working at subaudible frequencies.

Of interest to the physicist and chemist is a chapter on the principles of sonic testing and analysis. In this chapter there is a discussion of the various techniques for measuring the elastic properties of solids, i.e., both the dissipative and elastic moduli. A discussion of methods for measuring the visco-elastic properties of normal and polymer liquids is included. The appendix presents a short discussion of the significance of such measurements in the interpretation of the structure of liquids and solids.

The coverage of techniques is complete and includes foreign techniques as well as domestic. For engineers and physicists desiring to acquaint themselves with the various techniques, methods for constructing apparatus and what can be done with them, *Sonics* is highly recommended. The book does not completely cover the communication field since no mention is made of the use of wave transmission in delay lines or mechanical wave filters and their applications in the communication systems. In the interpretive field, only basic principles are covered. In the opinion of the reviewer, this is the most complete book on techniques and applications of sonic processes that has yet appeared, and it should be in the library of all engineers and physicists dealing with these processes.

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### Handbook of Microwave Measurements: Two Vols., Edited by Moe Wind and Harold Rapaport

Published (1954) by Polytechnic Inst. of Brooklyn, 55 Johnson St., Brooklyn 1, N. Y. Volume I, 20 Sections; Volume II, Illustrations; 4 Appendices. 8½×11. $12.00.

The material in these two volumes has been assembled for the guidance of technical personnel in the field of microwave measurements. The editors, recognizing that much of the information in this field is widely dispersed, have endeavored "to present a unified collated handbook of microwave measurement methods" in order that many sound methods may not be overlooked and remain unutilized. The two volumes, comprising upward of one thousand pages, are the work of twenty-five contributing authors. The material has been divided into twenty sections, each devoted to a particular characteristic quantity such as power, attenuation, impedance, etc. A unique feature is that all of the text is contained in Volume One and all of the illustrations are in Volume Two.

The material was originally prepared for the Signal Corps Engineering Laboratories, evidently as an instruction manual for students without previous experience in the field. For this reason highly detailed step-by-step procedural instructions are given. The amount of detail is perhaps greater than the average reader would wish, but it does serve to acquaint him with many aspects of the diverse measurement methods available. Although many of the methods are evaluated as regards accuracy, the distinction between fundamental methods and those relying on secondary standards is not emphasized. In particular, it was noted that the method of measuring conversion loss by directly measuring the input and output power is not given.

Each section begins with a theoretical analysis, quite detailed and complete, of the subject under discussion. Particularly valuable are the sections on propagation constant, impedance and dielectric constant. A great amount of detailed information is contained in these two volumes and those working in microwave measurements should find much which is useful to them.

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### Television Interference, Third Edition, Edited by Philip S. Rand

Published (1953) by Remington Rand, Inc., 315 Fourth Ave., New York, N. Y. 104 pages. $.25.

This book, like the two preceding editions, consists of reprints of technical articles on the subject of interference suffered by television receivers. The present volume contains 31 such articles and a list of recommended reading.

In view of the recent action of the FCC in proposing rules for the control of spurious radiations, this subject has become a matter of urgent interest to the designers of all electronic equipment. Most of this book is directed to the constructors and operators of amateur transmitters, but the subject of controlling spurious radiations in television receivers is also covered.

In a compilation of this type it cannot be expected that all of the material will be on a uniformly high technical level. However, taken with the preceding two editions, this book brings together a worthwhile collection of previously published material. Mr. Rand and the Remington Rand Corporation deserve the thanks of the industry for making this material available at a most modest price.

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### The Oscilloscope at Work by A. Haas and R. W. Hallows

Published (1954) by Iliffe and Sons, Ltd., Stamford St., London S.E., England. 167 pages+4 page index. 319 figures. 8½×5½. 15 s. 0d.

The book deals primarily with measurements of electrical circuit characteristics by the use of the oscillograph as the indicating instrument. The electrical circuits for which the oscilloscope's use is described include basic electrical circuits, audio frequency amplifiers, rf amplifiers, oscillators, rectifiers, modulators, phase shifting and wave shaping circuits, and certain limited television receiver measurements. In describing the oscilloscope's use, many waveforms are employed and can be a valuable aid to the reader as a general guide in the types and limitations of measurements that can be made in these fields. One complete chapter is devoted to oscilloscope operating troubles. Although somewhat limited, it is an excellent guide to the more important defects an oscillograph might have and their effect on application of an oscillograph as an instrument tool.

It is the reviewer's opinion that the title of the book *The Oscilloscope at Work* is somewhat misleading. The actual fields in which the oscilloscope works are many, and yet this book covers only one small sector of them. The circuits in the first chapter are extremely simple compared to the modern cathode-ray oscillograph.

In summary, let me say that the electrical engineer who has not used an oscillograph may find this a valuable aid in learning some of the fundamentals of the oscillograph in his measurement work.

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### Operations Research for Management by Joseph F. McCloskey and Florence N. Trefethen

Published (1954) by The Johns Hopkins Press, Baltimore 18, Md. 350 pages+5 page index+xxiv pages. 9¼×6¼. $7.50.

This is a comprehensive collection of articles prepared by individuals who are experts in their respective areas to this new science. Careful reading should do much to explain *Operations Research* and how it may be used as a tool of management. The majority of the articles were presented at a seminar held by Johns Hopkins University in the spring of 1952. The volume is divided into three parts preceded by a well-written introduction by Dr. Ellis Johnson. This introduction paves the way for what is to follow, making it more understandable.

Part I covers the history of operations research and the concepts of it as a profession and a science. This section starts with World War II and follows through its evolution to the present time. It points out the similarity of systems evaluations, operations evaluation, operational analysis, and operations research.

The second part, dealing with methodology, describes some of the mathematical and statistical techniques employed as well as some of the basic philosophy underlying the use of these techniques. This part of the text is particularly difficult reading for the layman. Many of the words used are not normally encountered nor will they be understood by the average member of management. The authors, however, have done their utmost to define their unusual or complex vocabulary and symbolic logic in terms that are generally understood.

Part III contains a carefully selected number of case histories. The variety of cases is such that most members of management will find general ideas indicating how operations research might well be applied in their own organization. The article by Dr. Horace C. Levinson, "Experiences in Commercial Operations Research," is worth reading.

One should read the introduction and Part I first. Part III should be read next. Part II should be read last. While the great majority of managers may not thoroughly understand the complexity of the methods described in Part II, it should give them a very definite idea of the value of these methods. It is unfortunate that a glossary of the technical terms used in this volume was not included. The bibliography is complete and well prepared.

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**Advances in Electronics and Electron Physics: Volume Six Edited by L. Marton**

Published (1955) by Academic Press Incorporated, 123 East 23 St., N. Y., N. Y. 518 pages+xi pages+19 page index. Illus. 9¼×6. $11.80.

This book is a collection of eight comprehensive reviews prepared by outstanding authorities from the United States, England and The Netherlands. Under the able editorship of L. Marton of the National Bureau of Standards, Volume Six carries on the fine tradition of this series. Reader interest will be particularly strong among physicists, chemical physicists, and radio engineers whose curiosity extends beyond the mundane problems of the radio-TV spectrum into areas where advance work is laying the foundation for knowledge from which many of the electronic inventions of tomorrow will spring.

The contributing authors are Elihu Abrahams, Rudolf G. E. Hutter, Henry F. Ivey, and W. M. Webster of the USA; M. E. Haine and A. B. Pippard of England; and J. Smit, J. Van Den Handel and H. P. J. Wijn of The Netherlands. The book is made up of chapters on: *Metallic Conduction of High Frequencies and Low Temperatures, Relaxation Processes in Ferromagnetism, Physical Properties of Ferrites, Space Charge Limited Currents, A Comparison of Analogous Semiconductors and Gaseous Electronics Devices, The Electron Microscope, Traveling-Wave Tubes, and Paramagnetism.*

Each chapter presents the reader with an erudite cross-section of contemporary research in the subject. A comprehensive list of references follows each writing. The authors offer more to interest the mature scientist-engineer than the beginner. But whether the volume is used for instruction or for reference material for research underway, the reader will gain from those parts which pertain to his field of interest or specialization.

The Editorial Board assisting Dr. Marton in bringing this excellent compilation to press consists of Allibone, Casimir, DeVore, Dow, Nier, Nottingham, Piore, Ponte, Rose and Smith.

This is a fine book and would be an excellent addition to the library of anyone interested in advanced work in the areas reported upon.

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**Laplace Transforms for Electrical Engineers by B. J. Starkey**

Published (1954) by Iliffe and Sons Ltd., Dorset House, Samford St., London, S.E. 1, England. 276 pages+3 page index. Illustrated. 8¾×5¾. 30s.

Perhaps the most efficient way of evaluating this book is to compare it to *Transients in Linear Systems* by Gardner and Barnes, since the latter is well-known to virtually every electrical engineer. The word similar refers only to the expressed intent of the books, each being an introduction to the use of Laplace transform methods in solution of practical problems. Of the two, Starkey begins with more elementary considerations but quickly, in 81 pages, covers most of the material to be found in Gardner and Barnes work, although with fewer examples; Starkey's examples are all concerned with electric circuits while the other work also treats mechanical and acoustical problems.

The bulk of Starkey's book is concerned with topics which do not appear at all in *Transients in Linear Systems:* i.e., complex variable theory and the evaluation of inverse Laplace transforms by contour integration. These topics are developed from the start with sufficiently clear and detailed explanations so that a person, not previously familiar with Cauchy's theorem, will find everything he needs in order to understand mathematical methods commonly used to find inverse transforms.

This book can be recommended very highly to the serious student who wishes to obtain more than just a smattering of the Laplace transform method, who wants to understand it from a mathematical standpoint sufficiently well so that he is freed from dependence on tables of transforms, a desirable objective that becomes more and more necessary in advanced work.

On the negative side, however, one word of warning is necessary. Starkey depends strongly on intuition and his mathematics has more vigor than rigor. Occasional liberties are taken with regard to convergence of integrals, interchange of order of integration, etc., which a mathematician would find quite hair-raising. For example, the integral in equation (9.6) does not exist unless one makes a qualification that does not appear in the text until three paragraphs later. Of course, to a practical man these mathematical questions will be regarded as mere nuisances; nevertheless they exist, and a prospective student should be cautioned that even in the most practical problems it will sometimes be necessary to use higher standards of mathematical rigor, not just for artistic reasons but in order to get the right answer.

This criticism is to be regarded as a very mild one; it is undoubtedly good pedagogy to defer considerations of rigor until after the student has a preliminary view of the field.

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**Electromagnetics by John D. Kraus**

Published (1953) by the McGraw-Hill Book Company, Inc., 330 West 42nd St., New York 36, N. Y. 555 pages+10 page index+7 page bibliography+30 page appendix+xiii pages. 379 figures. 6×9¼. $9.00.

This excellent text on electromagnetic theory is distinguished by its clarity and logicality. Mathematical material is fully developed, few steps being omitted, or, where they are omitted their justification being clearly described in a concise manner. All notation is completely defined. The order of presentation of theoretical developments is generally that which is most satisfactory for purposes of learning; in most cases the dependent variable is first formulated in an initial equation in terms of independent variables of broad significance, which, in turn are then determined in more specific terms. The book is copiously provided with illustrations and with worked examples, which, in themselves, form part of the text, and serve not only to apply the relationships obtained but also to develop them further. As the author says in the preface, simple special cases are usually considered first, and then with these as a background, the corresponding general cases are evolved. Vector concepts and operations are demonstrated as a part of the text material, wherever they are needed.

These qualities of lucidity, though obviously indispensable in any textbook, are actually found in so few that Dr. Kraus's book may be fairly said to be one of the best available.

Emphasis is on electromagnetic field theory, but, though this point of view is stressed, circuit theory is not neglected, and its relation to field theory is pointed out.

The first seven chapters of the text are intended for use in an introductory one-semester field-theory course at about the third or fourth-year college level, while the last seven chapters are written for a somewhat more advanced course of the same length at the senior or first year graduate level.

Subjects treated include the static electric field, the steady electric current, the static magnetic field, charged particles in electric and magnetic fields, time-changing electric and magnetic fields, Maxwell's equations, plane waves in dielectric media, plane waves in conducting media, transmission lines, wave guides, antennas, and boundary value problems.

This book can be highly recommended, not only as a teaching text-book, but as a reference book for engineers and physicists.

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**Introductory Circuit Theory by Ernst A. Guillemin**

Published (1953) by John Wiley & Sons, Inc., 440 Fourth Ave., New York, N. Y. 545 pages+4 page index+xxv pages. 199 figures. 6×9. $8.50.
Ernst A. Guillemin is Professor of Electrical Communication at the Massachusetts Institute of Technology, Cambridge, Mass.

We have here a text destined to have substantial influence on electrical engineering education and practice. Its importance is large, and it marks one of those milestones which give the steps—as contrasted with the inchlines—by which major progress is measured. We propose to discuss the work of Dr. Guillemin under four categories: its value in pointing the way toward a radical course (and ultimately curriculum) change; emphasis on recently recognized techniques which contribute, at a cost, much toward simplification and condensation; new material for which the text would be valuable irrespective of its other contributions, and finally, some criticisms—not all favorable—of details of the book.

University curricula in engineering and a few other fields receive severe and continuing pressures to introduce new material, move "advanced" material to lower class levels, and otherwise to include in a fixed span of time more and more. At the same time, raised living standards inculcate ideas of extensive leisure time, less strenuous work, and other concepts which permeate oncoming generations and lead to expectations of corresponding limitations on scholastic work weeks. By radical course and curriculum changes, electrical engineers who have kept in touch with engineering education have seen more and more intellectual material included in fewer and fewer class hours per week without great detriment to the engineering education of the students. Large parts of the "radical curriculum changes" have resulted from changes in emphasis, increases in efficiency, and condensation and elimination.

Dr. Guillemin's book qualifies under the first two categories, as a forerunner of another sharp change. Some material heretofore confined to advanced courses is presented for sophomore or junior use. Some emphases, such as that on sine-wave driving forces, are reduced and postponed in order to introduce new and preceding emphases on transient, pulse, and impulse analysis. Some new material virtually untaught previously is introduced. The total is an increase in efficiency of major proportion and a change in emphasis long overdue to bring first circuit theory courses more in accord with today's electrical engineering.

We consider a few of the items which support this statement. Just as the elements of matrix theory can be taught without difficulty to sophomores "whose mental attitude is not preconditioned" whereas seniors who have met matrices by backdoor methods and rumors have a psychological barrier which must be overcome, so, the teaching of the response of simple circuits to pulses and impulses can be used as a starting point of circuit analysis. And likewise the impedance concept, far more general than that of the impedance associated with the steady alternating state response of linear systems having impressed sine-wave driving forces, can be introduced early in technical life provided a firm base of general transient analysis precedes. Guillemin does these things, thus recognizing first that although the sine-wave driving force and response still remain paramount, *relatively* they are considerably less important than formerly and this should be reflected in beginning circuit theory courses; and second, that the impedance concept and the general pole-and-zero approach have reached a maturity which calls for introduction in the same beginning circuit theory courses.

The book under review emphasizes certain techniques—not original with the author—which in some respects have had far too little attention. To cite but one relatively simple example, a considerable emphasis is placed on Thevenin's theorem so that the student will learn how it joins with duality in enabling many circuit problems to be solved by way of one solution. The insistence made that circuit theory is an abstraction which may or may not have a one-to-one correspondence with a specific physical circuit is highly desirable in a beginning circuit theory course. And the insistence on the use of simple numbers and problems and examples (e.g., $R=1$, $L=2$, $C=3$, although somewhat of a far cry from some of Dr. Guillemin's previous works, recognizes the current trend to minimize distractions from basic ideas and mechanical effort required of the student to demonstrate his prowess in a highly theoretical field.

New material in the text stands out from the first chapter. Network topology for sophomores or juniors may seem somewhat startling at first glance, but it is actually a topic which should long since have been introduced in beginning circuit courses. Dr. Guillemin's book is probably the first of its kind in this respect, and the job is well done for the level at which the text aims, which is not to say that there are not weaknesses in it. But the general idea that, to take one example, the usual fuzzy introduction of loop currents in a circuit theory course should be superseded by going back an order of magnitude in theory to enable the student to have an intelligent grasp of the problem behind the introduction of loop currents is simply another of those long overdue reforms needed in elementary circuit theory. There is much incidental other new material in the book, on which we will not touch.

In the introductory paragraph of this review we reserved our final comments for "some criticisms—not all favorable—of details of the book." For a text of so much potential value, the criticisms are of an order of magnitude less in importance than the favorable comments which have preceded. With this understanding, we mention our first and outstanding complaint—verbosity. The book could have been written in about three-quarters of the space it takes, without loss to the reader. We will not illustrate this, and it can be taken as simply the opinion of the reviewer, but practice condensations have been supplemented by informal comments of others, so that the reviewer is inclined to feel that the defect stands out. Furthermore, occasional poor English and misuses of technical words are jarring. Guillemin's use of "potential" for example is often dubious to say the least, and the definition of "passive" has to be caught on the run. Could it be more than coincidence that neither "potential" nor "passive" appear in the index of the book?

Words like "resistor" and "resistance" both appear, but there is no consistency in their use. And words like "clearly" and "surely," those common introductions to a poor argument, appear in good-sized groups (five in one paragraph on page nine). But more important, the author uses "s" instead of "p" for complex frequency, and the reviewer has reason to believe this has been done simply because of ignorance of IRE standards. To use a well-worn cliché, it is regrettable that a text so modern in essence should perpetuate outmoded nomenclature rather than join in the attempt to clear con fusion in the field.

There are a few statements—the definition of the " 'value' of an element" on page three—which seem peculiar; and even the last paragraph of the Introduction, telling of the simplest broad class of networks (to which the book is devoted) omits an essential qualification.

The book is so arranged that chapters four through eight constitute a unit which the author points out can be used for a reduced course. In using this text, it is desirable that at least some instructors experiment, and one suggestion is to have chapters one through three follow the four through eight sequence.

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## Recent Books

*Abstracts of the Literature on Semiconducting and Luminescent Materials and Their Applications.* Compiled by Battelle Memorial Institute. John Wiley and Sons, Inc., 440 Fifth Ave., New York 15, N. Y. $5.00.

Booth, Andrew D., *Numerical Methods.* Academic Press, Inc., 125 East 23 St., New York 10, N. Y. $6.00.

Jacobson, Arvid W., ed., *Proceedings of the First Conference on Training Personnel for the Computing Machine Field.* Wayne University Press, Detroit, Michigan. $5.00.

Marcus, William, and Levy, Alex, *Elements of Radio Servicing.* McGraw-Hill Book Company, Inc., 330 West 42 St., New York 36, N. Y. $6.00.

Petrovsky, I. G., *Lectures on Partial Differential Equations*, trans. by A. Shenitzer. Interscience Publishers, Inc., 250 Fifth Ave., New York 1, N. Y. $5.75.

*Rider's Specialized Tape Recorder Manual, Volume One.* John F. Rider Publisher, Inc., 480 Canal St., New York 13, N. Y. $4.50.

Shedd, Paul C., *Fundamentals of Electromagnetic Waves.* Prentice-Hall, Inc., 15 East 26 St., New York 10, N. Y. $6.00.

# 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 *Wireless Engineer*, London, England**

**NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.**



The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

### U.D.C. CHANGES

In anticipation of a new edition of the Universal Decimal Classification Abridged English Edition (BS 1000 A), certain changes in U.D.C. numbers will be made in this and subsequent issues. The new numbers used will be:

- Radio astronomy: 523.16
- Ultrasonics: 534 subdivisions with the special analytical subdivision -8 attached.
- Sound recording and reproducing: 534.85
- Electroacoustic problems, transduction, etc.: 534.86.

**The Index to the Abstracts and References published in the PROC. IRE from February, 1954 through January, 1955 is published by the PROC. IRE, April, 1955, Part II. It is also published by *Wireless Engineer* and included in the March, 1955 issue of that journal. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.**

## ACOUSTICS AND AUDIO FREQUENCIES

**534.121.2** **1535**
**Symmetry of Vibrating Square Membrane**—M. D. Waller (*Proc. Phys. Soc.*, vol. 67, pp. 895–898; December 1, 1954.) Vibrations of free square plates always conform with the symmetry of the surface; this experimental conclusion conflicts with the accepted theoretical conclusion that two normal modes of vibration of the same frequency can combine regardless of symmetry considerations. This conflict is discussed. See also *Acustica*, vol. 4, no. 6, pp. 677–680; 1954.

**534.2:534.833** **1536**
**The Propagation of Sound in Granular Substances**—H. Schmidt. (*Acustica*, vol. 4, no. 6, pp. 639–652; 1954. In German.) Theory is developed and measurements are reported of the loss factor and propagation velocity of granular substances (a) for substances packed in a bar-shaped container, (b) for a layer of the substance on top of an Al bar. In case (a) sand, with varying water content, glass spherules 1 mm in diameter, finely broken brick, cinders and coarse iron filings were used, in case (b) only sand and brick. For application as a filling e.g. for sound insulation in buildings, cinders or broken brick are recommended, with wood shavings or rock wool added where necessary to reduce the resonance frequencies to very low values.

**534.21-8** **1537**
**The Propagation of Sound in Solutions of Rubber and Perspex**—A. W. Pryor. (*Acustica*, vol. 4, no. 6, pp. 658–661; 1954.) Absorption measurements at four frequencies in the range 4–17.3 mc were made on three rubber-in-benzene solutions and on one perspex-in-pyridine solution. "The absorption was less than in the pure solvents and there is evidence of relaxation. Measurements of the shear viscosity of a 10 per cent rubber solution showed that the relaxation of the high flow viscosity was complete even at 50 kc/s. The absorption in the solutions must therefore be ascribed to the 'bulk viscosity' of the solvents."

**534.22-14** **1538**
**Measurements by Optical Methods of the Sound Velocity in Aqueous Solutions of Electrolytes in dependence on Concentration and Temperature**—K. Tamm and H. G. Haddenhorst. (*Acustica*, vol. 4, no. 6, pp. 653–657; 1954.)

**534.6:621.373.4:621.374** **1539**
**Generation and Use of Single-Frequency Pulses in Electroacoustics and Musical Acoustics**—H. Lackner. (*Ost. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 8, pp. 141–152; November /December, 1954.) Analysis is presented showing how the spectrum of a short train of sinusoidal oscillations depends on its duration and on the initial and final phase. A circuit for a generator providing pulses of variable duration and phase is described. Advantages of using such signals for testing loudspeakers and musical instruments are indicated.

**534.846** **1540**
**Improvements in the Acoustics of the Budapest Civic Theatre**—T. Tarnóczy. (*Acustica*, vol. 4, no. 6, pp. 665–671; 1954. In German.) The upper walls and part of the ceiling were provided with a diffusive surface, made of reinforced plaster of Paris. The rest of the ceiling was plastered over. The lower walls and the new proscenium were lined with wood paneling, which acted as an acoustic resonator. A sound-amplification system is available if required. These measures sufficed to remove flutter echoes and almost all other audible echoes, to increase the reverberation time to 1.6 seconds, and to make the distribution of sound energy completely satisfactory in almost every part of the theater. High frequencies have been enhanced and tonal quality is good.

**534.86** **1541**
**Advances in the Reproduction of Music**—F. Winckel. (*Funk u. Ton*, vol. 8, pp. 604–607 and 649–652; November and December, 1954.) A series of brief reports on lectures delivered at three conferences held in 1954. The subjects covered include a combination loudspeaker unit, automatic volume control, a new compressor-expander system with variable time delay, and the quality of microphones and loudspeakers.

**534.861:534.76** **1542**
**Experiences in Stereophonic Broadcast Transmissions**—J. J. Geluk. (*Funk u. Ton*, vol. 8, pp. 631–634; December, 1954.) Report dealing mainly with Dutch experimental transmissions.

**534.862:534.76** **1543**
**Stereophonic Sound-Film Recording and Reproduction**—H. Friess. (*Funk u. Ton*, vol. 8, pp. 622–630; December, 1954.) A review of current problems and technique.

**621.395.616** **1544**
**Note on the Stabilization of the Response of a Capacitor Microphone**—C. Colin. (*Jour. Phys. Radium*, vol. 15, pp. 820–822; December, 1954.) It is possible to improve the response of a capacitor microphone by negative feedback if (a) an auxiliary electrode is included, and (b) the voltage output is in phase with diaphragm displacement. An analysis is made of the necessary coupling conditions, both mechanical and electrical.

**621.395.625.3** **1545**
**A Survey of Magnetic Recording**—S. J. Begun. [*Elec. Engng.* (*New York*), vol. 73, pp. 1115–1118; December, 1954.] Includes a description of operation and an indication of the development and applications.

**621.395.625.3** **1546**
**Making Magnetic Sound-Records Visible**—W. Guckenburg. (*Funk u. Ton*, vol. 8, pp. 600–604; November, 1954.) A review of various methods using suspensions of ferromagnetic dust. Photographs illustrating applications, such as detecting faulty magnetic heads, are briefly discussed.

## ANTENNAS AND TRANSMISSION LINES

**621.315.28:621.395.44** **1547**
**A Transatlantic Telephone Cable**—M. J. Kelly, G. Radley, G. W. Gilman and R. J. Halsey. (*Proc. IEE*, Part B, vol. 102, pp. 117–130; March, 1955. Discussion, pp. 130–138.) The inadequacy of radio circuits to cope with transatlantic communications is discussed and a general description is given of the cable system for linking the United Kingdom, Canada and the United States; this is planned for completion in 1956 and will provide 36 telephone circuits across the Atlantic and 60 between Newfoundland and Nova Scotia. The project has been made possible by the development of submerged repeaters containing long-life tubes and other components.

621.372 1548
**The Launching of a Plane Surface Wave**—G. J. Rich. (*Proc. IEE*, Part B, vol. 102, pp. 237–246; March, 1955.) Experiments are described on the propagation of surface waves at about 3 cm λ over plane brass plates with polystyrene coatings of thickness such that only the $TM_0$ mode is propagated. Various launching devices were tried; the best found was a "double-cheese" device with a launching efficiency of about 50 per cent. Experimentally found values for the efficiency of this device are in agreement with theory for apertures of dimensions smaller than λ. Deviations from theoretical values observed for larger apertures indicate that to obtain good launching efficiency the field distribution within the aperture must be to a close approximation exponential.

621.372.22 1549
**The Matrix Equation of Loss-Free Exponential Lines**—B. Beghin. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 168–170; January 10, 1955.] A solution of the differential equation for the exponential line is presented in the form of the product of two affine exponential expressions representing a ladder arrangement of two quadripoles, the one a perfect transformer and the other analogous to a uniform line. Reflection can be eliminated by terminating the line with a complex impedance.

621.372.8 1550
**The Propagation of an Electromagnetic Wave along an Infinite Corrugated Surface**—R. A. Hurd. (*Canad. Jour. Phys.*, vol. 32, no. 12, pp. 727–734; December, 1954.) The problem is analyzed by a method based on the calculus of residues; the slot walls constituting the corrugations are assumed to be vanishingly thin. Exact expressions are obtained for the mode amplitudes and phase velocities; these are valid for frequencies near cut-off, when the number of slots per wavelength is about five or more.

621.372.8 1551
**A Simple Waveguide Directional Coupler**—P. Andrews. (*Jour. Brit. IRE*, vol. 15, pp. 112–116; February, 1955.) A simply constructed, orthogonally crossed unit is described giving constant coupling and directivity over a wide frequency band. Coupling factor and directivity are calculated for two typical cases for 3 inch×$1\frac{1}{2}$ inch waveguide. An expression is derived for the frequency sensitivity of the device.

621.372.8.002.2 1552
**Surface Roughness and Attenuation of Precision-Drawn, Chemically Polished, Electropolished, Electroplated and Electroformed Waveguides**—J. Allison and F. A. Benson. (*Proc. IEE*, Part B, vol. 102 pp. 251–259; March, 1955.) "Detailed examinations of certain 3 cm waveguides have shown that the surface finish of precision-drawn tubing as manufactured at present is quite adequate for most applications. Such surfaces, however, may be improved, if desired, by careful chemical or electrolytic polishing or electroplating in bright baths under closely controlled conditions. Some information on the surface finish of copper guides electroformed on various types of mandrel is also presented. Formulae for calculating the attenuation of any H or E mode in a rectangular waveguide, so as to take account of surface roughness, have been developed from the original expressions derived by Kuhn. A method is given for determining the actual value of attenuation in a waveguide sample without having to make careful measurements with elaborate equipment on long specimens. A description is included of a new and simple technique, involving electropolishing, for examining the internal surface finish of waveguides; the method cannot, however, be used successfully on silver-plated sections."

621.372.8.029.65.002.2 1553
**The Electroforming of Components and Instruments for Millimetre Wavelengths**—A. F. Harvey. (*Proc. IEE*, Part B, vol. 102, pp. 223–230; March, 1955.) Electroforming processes considered for producing waveguide components include periodic-reverse-current plating. Either permanent or disposable formers may be used. The electroformed parts are designed to facilitate subsequent machining. Data are tabulated on a range of waveguides with internal cross sections from 0.28×0.14 inch to 0.034×0.017 inch which have been standardized for the Joint Services, and various units incorporating these elements are illustrated.

621.396.67:539.23 1554
**Possibilities of Radiation from Thin Metal Films**—M. Gourceaux. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 170–172; January 10, 1955.] Systems with rotational symmetry and circular current paths are considered. Expressions are derived for the values of film thickness and current frequency for which radiation can occur with cylindrical and spherical arrangements.

621.396.676.2 1555
**The Notch Aerial and some Applications to Aircraft Radio Installations**—W. A. Johnson. (*Proc. IEE*, Part B, vol. 102, pp. 211–218; March, 1955.) Analysis and experiments show that metallic sheets with notches cut perpendicular to an edge can make very efficient antennas. Wide-band characteristics can be obtained with a notch about λ/4 deep, but the detail of the polar diagram may vary with frequency, depending on the shape and size of the sheet. Notches short compared with λ—termed "nitches"—can be operated either as fixed-tuned narrow-band antennas or, in association with a tuning unit, over a wider band. For reception at frequencies around 100 mc a notch of length about 3 inches is adequate. Tests made with aircraft tail fins and wings are described; hf., vhf and uhf types are practical.

621.396.677:523.16 1556
**Aerial Smoothing in Radio Astronomy**—Bracewell and Roberts. (*See* 1638.)

621.396.677.029.62/.63:621.397.6 1557
**U.H.F. and V.H.F. Antenna**—R. F. Kolar. (*Radio-Electronics*, vol. 25, pp. 60–62; December, 1954.) A low-cost television receiver antenna unit is described; the design is based on stacked uhf V's and an uhf dipole and reflector. The power gain relative to a λ/2 dipole is a few decibels.

621.396.677.31 1558
**Optimum Element Spacing of Uniform Broadside Arrays**—R. F. H. Yang. (*Wireless Eng.*, vol. 32, pp. 115–116; April, 1955.) Collinear and curtain arrays are considered. A brief analysis indicates that the optimum element spacing for a collinear array is λ, irrespective of $n$, the number of elements, while the optimum spacing for a curtain array is given approximately by the formula $d/\lambda = 0.6 + \frac{1}{3}\log_{10} n$. This result is different from that obtained by Hammond (1909 of 1953).

621.396.677.8.029.65 1559
**Some Experiments on the Reflecting Properties of Metal-Tube Lens Medium**—E. M. Wells. (*Marconi Rev.*, vol. 17, no. 114, pp. 74–85; 3rd Quarter, 1954.) Experiments described show qualitatively the variation of the reflection coefficient with angle of incidence, plane of incidence, and plane of polarization. With the $E$-vector in the plane of incidence a very pronounced anomalous reflection was observed. This was further investigated. The wavelengths used were in the band 8–10 mm.

## AUTOMATIC COMPUTERS

681.142 1560
**Wide-Band Analog Function Multiplier**—J. A. Miller, A. S. Soltes and R. E. Scott. (*Electronics*, vol. 28, pp. 160–163; February, 1955.) Operation of the multiplier is based on the formula $xy = \frac{1}{4}[(x+y)^2 - (x-y)^2]$. The squaring is effected by means of beam-deflection tubes with parabolic characteristics [2940 of 1950 (Soltes)].

681.142 1561
**Analogue Computer with Stepping-Switch Drive for the Solution of Algebraic Equations of up to Sixth Degree**—H. Hörner and H. Zemanek. (*Ost. Z. Telegr. Teleph. Funk Fernsehtech.*, vol. 8, pp. 153–158; November/December, 1954.) The analogy used is that between a sinusoidal voltage and a complex quantity. The representation of the polynomials is expected using feedback amplifiers as computing elements. The zero points corresponding to the solutions are found automatically.

681.142 1562
**Some Devices permitting Study of the Variable Regime in the Transportation of Gas in Pipes**—A. Blanc. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 45–46; January 3, 1955.] A particular application is discussed of the analog-computer arrangements described previously (2317 of 1954).

681.142 1563
**Programming a Digital Computer for Cell Counting and Sizing**—W. Welkowitz. (*Rev. Sci. Instr.*, vol. 25, pp. 1202–1204; December, 1954.)

681.142:538.221 1564
**Magnetic Materials for Digital-Computer Components: Part 1—A Theory of Flux Reversal in Polycrystalline Ferromagnetics**—Menyuk and Goodenough. (*See* 1713.)

681.142:621.3.042 1565
**Logical and Control Functions performed with Magnetic Cores**—S. Guterman, R. D. Kodis and S. Ruhman. (PROC. IRE, vol. 43, pp. 291–298; March, 1955.) Use of magnetic cores with square hysteresis loops in digital-computer systems based on the "single-line" shift register is described.

681.142:621.375.23 1566
**A Technique for Nonlinear-Function Generation**—P. N. Nikiforuk. (*Electronic Eng.*, vol. 27, pp. 118–119; March, 1955.) A method is described for converting a circuit whose output is proportional to the $n$th root of the input to give the $n$th power, or vice versa. A practical circuit, based on a variable-gain feedback amplifier, for dealing with squares and square roots is illustrated.

681.142:621.376 1567
**Two New Electronic Analog Multipliers**—M. A. Meyer and H. W. Fuller. (*Rev. Sci. Instr.*, vol. 25, pp. 1166–1172; December, 1954.) Improved types of four-quadrant multipliers for deriving the product of two time functions are described. One uses a double-amplitude modulation scheme applying the work of Sternberg and Kaufman on the two-frequency-modulation-product problem (2212 of 1954). The other uses successive phase modulation and amplitude modulation of a carrier. The accuracy obtainable is discussed.

681.142 1568
**Automatic Digital Computation [Book Review]**—Publishers: H. M. Stationery Office, London, Eng. 296 pp., 21s. (*Instr. Practice*, vol. 8, p. 1094; December, 1954.) Contains discussions and original papers presented at the symposium held at the National Physical Laboratory in March, 1953.

## CIRCUITS AND CIRCUIT ELEMENTS

621.314.22.042 1569
**Calculation of Losses in Transformer Cores**—A. L. Morris. [*Engineer* (*London*), vol. 198, pp. 837–839 and 875–877; December 17 and 24,

1954.] Expressions for core losses are developed which take account of the nonuniformity in the flux distribution across the face of the core plates. The nonuniformity across the stack of plates is neglected.

621.318.4:538.221:621.396.822 1570
**The Noise generated in a Coil with a Ferromagnetic Core**—G. Builder and D. Haneman. (*Aust. Jour. Phys.*, vol. 7, pp. 654–658; December, 1954.) The discussion presented indicates that when there is no varying or alternating magnetization of the core, the noise can be calculated from the Nyquist formula, provided that the resistance of the coil is measured using an alternating voltage of such magnitude that the magnetization process remains linear.

621.318.435.3 1571
**Composite Cores for Instrument Transductors**—E. H. Frost-Smith and A. E. DeBarr. (*Proc. IEE*, Part II, vol. 101, pp. 663–667; December, 1954. Discussion, pp. 667–671.) The characteristics of self-excited transducers are markedly dependent on the properties of the core material. Cores with satisfactory properties can be produced at reasonable cost by combining different materials. Experimental results are reported for a core comprising a mumetal bridge and a U-shaped grain-oriented Si-Fe yoke.

621.318.435.3:621.375.327 1572
**Auto-Self-Excited Transductors and Push-Pull Circuit Theory**—A. G. Milnes and T. S. Law. (*Proc. IEE*, Part II, vol. 101, pp. 643–662; December, 1954. Discussion, pp. 667–671.) The modes of operation of transductors with parallel, bridge, and center-tap connections are explained. Expressions are derived for the sensitivity and time-constant under idealized conditions. An examination is made of the use of transductors in push-pull pairs to obtain magnetic amplifiers with outputs whose polarity depends on the signal sense; the transductors are assumed to be of the auto-self-excited type with resistance loads.

621.318.5 1573
**Component Design Trends—Special-Purpose Relays gain New Uses**—F. Rockett. (*Electronics*, vol. 28, pp. 150–156; February, 1955.)

621.318.5:621.318.134 1574
**Time Delay in High-Speed Ferrite Microwave Switches**—R. C. LeCraw and H. B. Bruns. (*Jour. Appl. Phys.*, vol. 26, p. 124; January, 1955.) Measurements indicate that for an *X*-band switch using a rod of a particular Mg-Mn ferrite and actuated by a current pulse of given magnitude and rise time 17 mμs, the time delay of the switch is about 3.2 mμs.

621.319.4:621.315.614.6 1575
**Study of the Dielectric in Paper Capacitors**—E. Briganti. (*Alta Frequenza*, vol. 23, pp. 139–156; June/August, 1954.) Measurements were made of the loss angle of wet and dry paper at −5 degrees C., and of the permittivity of the impregnant and of the impregnated paper. From these an equivalent circuit was derived for the capacitor, which can be used in estimating the quality of the paper and desirable conditions of manufacture. Breakdown-strength measurements were also made to determine the optimum applied voltage.

621.372.5 1576
**Arc Inductance and Dynatron Capacitance**—J. Groszkowski. (*Bull. Acad. Polon. Sci.*, Classe 4, vol. 2, no. 1, pp. 41–45; 1954. In English.) The stability of systems with nonlinear negative resistances of arc and dynatron types is discussed. Analysis indicates that though the arc can have some inductive properties and the dynatron capacitive ones, these are not essential to an explanation of the behavior of the systems. Transient conditions and steady oscillation states are considered.

621.372.5 1577
**Further Bounds existing on the Transient Responses of Various Types of Networks**—A. H. Zemanian. (PROC. IRE, vol. 43, pp. 322–326; March, 1955.) Five theorems further to those presented previously (2331 of 1954) are proved and illustrated.

621.372.5 1578
**An Approximate Treatment of Cascaded Four-Terminal Networks**—H. L. Armstrong. (*Electronic Eng.*, vol. 27, pp. 130–131; March, 1955.) "An approximate expression is derived for the *n*th power of a 2×2 matrix. The result is used in an approximate treatment of a ladder network used as a filter."

621.372.5 1579
**The Approximation Problem of Network Synthesis**—S. Winkler. (*Trans. IRE*, vol. CT-1, no. 3, pp. 5–20; September, 1954.) A review with 240 references.

621.372.5 1580
**Series Resonant Circuit Theory**—A. J. Lyon. (*Wireless Eng.*, vol. 32, pp. 107–108; April, 1955.) Expressions are derived for the fractional errors in tuning capacitance, resonance frequency, maximum current, and selectivity at current resonance due (a) in the case of frequency tuning, to the frequency dependence of the circuit resistance, and (b) in the case of capacitance tuning, to capacitor losses and coil self-capacitance.

621.372.5:512.83 1581
**The Mesh Counterpart of Shekel's Theorem**—S. Seshu. (PROC. IRE, vol. 43, p. 342; March, 1955.) It is shown that whereas the determinant of the admittance matrix of a network is independent of the choice of reference mode [2878 of 1954 (Shekel)], the corresponding statement for the mesh determinant is not true.

621.372.5:621.3.018.7 1582
**Distortion of Arbitrary Waveforms by Resonance Sections**—E. William. (*Funk u. Ton*, vol. 8, pp. 592–599; November, 1954.) The method developed earlier (2032 of 1954) for *RC* sections is extended to sections including an inductance.

621.372.5:621.318.134 1583
**Extension of Nonreciprocal Ferrite Devices to the 500–3000 Mc/s Frequency Range**—R. H. Fox. (*Jour. Appl. Phys.*, vol. 26, p. 128; January, 1955.) Calculations indicate that devices, such as circulators, of reasonable dimensions can be designed for operation at frequencies below 3 kmc by using static magnetic fields of intensity greater than required for ferromagnetic resonance.

621.372.512.2.029.65:621.372.8 1584
**A Short-Slot Hybrid for 9 mm**—E. M. Wells. (*Marconi Rev.*, vol. 17, pp. 86–87; 3rd Quarter, 1954.) The *X*-band junction described by Riblet (1833 of 1952) was redesigned for a wavelength of 9 mm. A brief illustrated note on this junction is given.

621.372.512.24 1585
**Resonance Conditions in a System of Two Circuits with Inductive Coupling**—U. Ruelle. (*Alta Frequenza*, vol. 23, pp. 157–177; June/August, 1954.) Assuming $M$, $R$, $L$ and $C$ constant, conditions are derived for the existence of one minimum and two maximum values of current in the secondary, in the general and in two particular cases. The results are applied to a band-pass filter, and presented graphically for two different parameter values. Low $Q$ values are assumed.

621.372.54 1586
**Microwave Filters**—E. Willwacher. (*Fernmeldetech. Z.*, vol. 7, pp. 694–704; December, 1954.) Theory of filters comprising coaxial lines and/or waveguides is developed by reference to equivalent lumped-constant circuits. Band-pass and band-stop filters of ladder, amplifier and bridge types are discussed.

621.372.54:621.396.67 1587
**Tunable Microwave Aerial Diplexer**—O. Laaff. (*Fernmeldetech. Z.*, vol. 7, pp. 688–693; December, 1954.) A diplexer for the band 2.1–2.3 kmc is based on the use of continuously tunable resonant circuits comprising coaxial-line sections short-circuited at the one end and having a third coaxial conductor insulated from and sliding within the inner conductor. In a particular design illustrated, for 100-mc separation between transmission and reception frequencies, the transmitted energy penetrating into the receiver is attenuated by 65 db.

621.372.543.2 1588
**Theoretical Investigation of Three-Stage Tchebycheff-Type Band-Pass Filters**—B. Betzenhammer and E. Henze. (*Arch. elekt. Übertragung*, vol. 8, pp. 545–552; December, 1954.)

621.373.4 1589
**Judging the Quality of Oscillator Circuits**—H. Haller. (*Funk u. Ton*, vol. 8, pp. 565–575; November, 1954.) Discussion of the frequency and amplitude stability of oscillators considered as composed of a slightly nonlinear frequency-independent amplifier and a linear frequency-dependent feedback quadripole. Stability is found to depend on the rate of change of the imaginary part of the voltage transfer characteristic of the feedback quadripole. The use of this criterion in the design of bridge-stabilized *RC* oscillators is illustrated. Measurements on and various faults of experimental oscillators are briefly discussed.

621.373.4:621.374:534.6 1590
**Generation and Use of Single-Frequency Pulses in Electroacoustics and Musical Acoustics**—Lackner. (*See* 1539.)

621.373.431.1+621.375.2.018.756 1591
**Experimental Investigations on Multivibrators and Amplifier Circuits with Secondary Electron Emission Valves as described by Kroebel**—K. E. Rumswinkel. (*Z. angew. Phys.*, vol. 6, pp. 551–556; December, 1954.) Multivibrators of the type described by Kroebel (383 of February) were investigated. Pulse flank slopes of the order of $10^{10}$ v/second, up to pulse amplitudes of 67 v, were obtained. The multivibrator can be modified to act as a pulse amplifier without feedback.

621.373.44:535.33 1592
**Equipment for Excitation of Spectra by High Frequency Pulses**—L. Minnhagen and L. Stigmark. (*Ark. Fys.*, vol. 8, pp. 471–479; December 14, 1954. In English.) The equipment comprises a pulse generator which controls a Clapp oscillator (2193 of 1948) followed by a frequency multiplier and amplifiers. Average hf power transferred to the discharge tube is about 500 w, with peak power of 3 kw; the frequency is about 9 mc. Ar spectra obtained are shown.

621.373.52 1593
**Transistor Frequency Standard**—J. H. Smith, Jr, and M. Campbell. (*Tele-Tech*, vol. 13, pp. 90–91, 135; December, 1954.) This unit, designed primarily for geophysical prospecting, has a printed-circuit base and uses transistors in all stages. Accuracy is within 1 part in $10^4$ over the range −40 degrees F to +140 degrees F. An 8-kc oscillator is followed by a pulse-forming amplifier and first divider stage, then by two dividers in parallel, whose outputs are mixed to obtain an output frequency of 100 cps.

621.375.1.024 1594
**The Transient Response of Direct-Current Amplifier Systems**—J. H. Sanders. (*Jour. Sci. Instr.*, vol. 31, pp. 453–455; December, 1954.) Dc amplifiers of the dc–ac conversion type

have an upper frequency response limited by the detector circuit, and when negative feedback is used transients are amplified considerably more than the steady signal. The form of the transient response and methods of reducing its magnitude are discussed.

621.375.2.049 1595
**A Long-Lived Packaged Amplifier for Aircraft**—J. G. Matthews. (*Bell. Lab. Rec.*, vol. 32, pp. 462–466; December, 1954.) The development and construction of units with a probable life of 2,000 hours are described. Selected tubes are pressed into cast Al wells lined with a silicone rubber paste which is hardened by a short curing process; this is more effective for heat transfer than Al dust or foil. Deposited carbon resistors and capacitors serviceable at 125 degrees C. are mounted on a phenolic board supported by the Al base and wired to a recessed plug. The assembled unit, after testing, is embedded in liquid plastic which is then solidified. Test figures for heat dissipation under different conditions are given.

621.375.221:621.372.512 1596
**Amplifier Stages with Transitionally Coupled Two-Stage Band-Pass Filters, particularly for Large Bandwidths**—W. Mansfeld. (*Funk u. Ton*, vol. 8, pp. 576–591; November, 1954.) The amplitude and group-delay characteristics of an amplifier stage consisting of two coupled circuits are analyzed for the case when the amplification is constant over a wide frequency band ("transitional" coupling). The case when the damping factors $d_1$ and $d_2$ of the two circuits are equal is considered first, and formulas are also given for the cases of either $d_1$ or $d_2$ tending to zero. Formulas are also given for transforming a filter with indirect inductive coupling into one with direct inductive coupling. Design curves are shown.

621.375.23:621.3.016.35 1597
**Nyquist's Criterion for Multiple-Loop Feedback Circuits**—L. Tasny-Tschiassny. (*Wireless Eng.*, vol. 32, pp. 114–115; April, 1955.) A method of deriving the stability criterion alternative to that of Cutteridge (3489 of 1954) and based on conformal transformation is presented. See also 72 of February (Cutteridge).

621.375.232 1598
**The Effect of Inverse Feedback on Input Impedance**—J. B. Earnshaw. [*Radio Elect. Rev.* (*Wellington, N. Z.*). vol. 9, pp. 37–40 and 34–35; December, 1954 and January, 1955.] Formulas are given for the input impedance, gain, and gain without feedback of twelve single-stage amplifiers and these, together with the circuits and their equivalents, are tabulated.

621.375.3 1599
**Alteration of Dynamic Response of Magnetic Amplifiers**—R. O. Decker. [*Elec. Engng.* (*New York*), vol. 73, p. 1088; December, 1954.] Digest of paper to be published in *Trans. Amer. IEE*, Part I, *Communication and Electronics*, 1954; pp. 658–664. Analysis is presented indicating how a magnetic amplifier of full-wave self-saturating type can be made to exhibit phase lead or lag by adjusting the parameters of the feedback networks.

621.375.4:621.314.7 1600
**Analysis of the Common-Base Transistor Circuit**—F. Oakes. (*Electronic Engng.*, vol. 27, pp. 120–126; March, 1955.) Simple equations are obtained by choosing a hybrid inverted-Π network as the equivalent circuit for investigating the operation of the grounded-base point-contact transistor amplifier.

621.375.4:621.314.7 1601
**D.C. Stability of Transistor Circuits**—F. Oakes. (*Wireless World*, vol. 61, pp. 164–167; April, 1955.) The design of amplifiers using junction transistors is discussed, with particular reference to the influence of the base-to-collector leakage current, which increases rapidly with rising temperature. For stable operation, the change of collector current produced by a change of leakage current should be low; circuit arrangements for achieving this are indicated.

621.376.3:621.3.018.78:621.372.5 1602
**The Distortion of F.M Signals in Passive Networks**—R. H. P. Collings and J. K. Skwirzynski. (*Marconi Rev.*, vol. 17, no. 115, pp. 113–136; 4th Quarter, 1954.) Expressions for the fundamental and first four harmonics of the output instantaneous frequency are given in a series expansion in terms of modulation frequency and modulation index, the coefficients in the expansion depending on the network parameters. Effects of detuning are considered, and detailed results for the Butterworth circuit are presented.

621.376.3:621.3.018.78:621.372.543.2 1603
**The Linear Distortion of F.M. Signals in Band-Pass Filters for Large Modulation Frequencies**—J. K. Skwirzynski. (*Marconi Rev.*, vol. 17, no. 115, pp. 101–112; 4th Quarter, 1954.) The linear distortion of a FM signal in a highly selective bandpass filter follows almost exactly the static response curve of the network, provided the following conditions are fulfilled:—(a) the modulation frequency is not less than two-thirds of the semi-bandwidth, (b) the modulation index does not exceed unity, (c) the network $Q$ is sufficiently large.

621.376.32:621.318.134 1604
**A Ferrite Frequency Modulator**—F. Slater. (*Marconi Instrumentation*, vol. 4, pp. 186–193, 200; December, 1954.) The modulator described comprises a ferroxcube-B4 ring carrying a rf winding and located in the gap of a Ni-Fe core carrying lf and polarizing windings. It is suitable for operation over the frequency range from 400 kc to uhf, with slight modifications. Its demodulated distortion is 5 per cent at an oscillator frequency of 400 kc and a 15-kc frequency deviation, and 2 per cent at 170 mc and 100 kc, respectively.

621.376.332 1605
**Discriminator Circuit Analysis**—F. L. Morris. (*Wireless Eng.*, vol. 32, pp. 93–98; April, 1955. Correction, *ibid.*, vol. 32, p. 142; May, 1955.) Theory and performance figures are given for a simple frequency-discriminator circuit using an asymmetrical arrangement. The conversion efficiency compares favorably with that of the equivalent Foster-Seeley discriminator. It is particularly recommended for purposes where high efficiency is of greater importance than accurate linearity over a wide frequency range.

621.396.049.75 1606
**A Universal Printed Circuit**—J. R. Goodykoontz. (*Tele-Tech.*, vol. 13, pp. 74–75; December, 1954.) The universal printed-circuit board has a standard pattern of parallel wires on one side and is useful at the design stage, when prototypes are required for testing, since circuit changes may be made with relative ease.

621.372.5 1607
**Amplitude-Frequency Characteristics of Ladder Networks [Book Review]**—E. Green. Publishers: Marconi's Wireless Telegraph Co., Chelmsford, 1954, 155 pp., 25s. (*Brit. Jour. Appl. Phys.*, vol. 5, p. 457; December, 1954.) Suitable mainly as a reference work for telecommunication engineers.

## GENERAL PHYSICS

53.081 1608
**Proposals for Units for Area, Electric Displacement and Magnetic Field Strength**—P. Cornelius. (*Philips Res. Rep.*, vol. 9, pp. 444–457; December, 1954.)

537.21+538.12 1609
**The Two-Dimensional Magnetic or Electric Field of a Single Isolated Pole-Piece**—N. H. Langton and N. Davy. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 431–435; December, 1954.) Theoretical investigation for a pole-piece consisting of a thick plate terminated by a concave semicircular cylinder. Conformal transformations and elliptic functions are used. The variation of the field strength along the edge of the pole-piece and along the external axis of symmetry is calculated and shown graphically.

537.224 1610
**Observation of the Costa Ribeiro Effect on the Dissolution of Naphthalene Crystals**—E. Rodrigues. (*Ann. Acad. Bras. Sci.*, vol. 26, pp. 381–383; December 31, 1954.) Report of experiments indicating that electric charges are developed on dissolving single crystals of naphthalene in a solvent of very low conductivity such as toluene; the crystals thus treated constitute natural electrets.

537.226:536.421.1:537.29 1611
**Electro-fusion: a New Phenomenon observed in the Phase Changes of Dielectrics under the Influence of an Electric Field**—J. Costa Ribeiro. (*Ann. Acad. Bras. Sci.*, vol. 26, pp. 349–355; June, 30 1954.) Experiments are described which indicate that application of a field between the electrodes of a capacitor with the dielectric partly in the solid and partly in the liquid state accelerates the phase change in the dielectric. From measurements of the current and calculation of the Joule energy dissipated, it is shown that this energy is several hundred times smaller than the heat necessary for the normal fusion of the corresponding mass of the dielectric.

537.226:536.421.1:537.29 1612
**Field-Induced Melting of Dielectrics**—B. Gross. (*Ann. Acad. Bras. Sci.*, vol. 26, pp. 289–291; June 30, 1954.) Theory is presented relevant to the phenomena described by Costa Ribeiro (1611 above).

537.311.1 1613
**Plasma Oscillations in a Periodic Potential: the One-Zone Theory**—J. Hubbard. (*Proc. Phys. Soc.*, vol. 67, pp. 1058–1068; December 1, 1954.) The plasma-oscillation theory of collective electron interactions developed by Bohm and Pines (1375 of 1954 and back references) can be applied to the problem of the conduction electrons in metals by modifying it to take account of the periodic potential present. This modification is carried out, neglecting the interactions of electrons in different zones and assuming that the effect of interzone transitions on the collective behavior is small enough to be treated by perturbation theory. The main effect of the potential is to alter the effective mass of the electrons.

537.311.31+536.212.2 1614
**The Electrical and Thermal Conductivities of Monovalent Metals**—J. M. Ziman. (*Proc. Roy. Soc.* A, vol. 226, pp. 436–454; December 7, 1954.) "Numerical calculations for the case of sodium, using the Bardeen (1937) formula for the scattering cross-section and Blackman's (1951) value for a 'longitudinal Debye temperature,' agree better with observation than do the simple Bloch expressions, but there still remain discrepancies." Further study of the scattering cross-section function might remove these.

537.311.31+537.312.62 1615
**High-Frequency Resistance of Tin and Indium in the Normal and Superconducting State**—C. J. Grebenkemper. (*Phys. Rev.*, vol. 96, pp. 1197–1198; December 1, 1954.) Further measurements at 24 kmc confirm results obtained previously [3060 of 1952 (Grebenkemper and Hagen)].

537.311.31 1616

**The Physical Nature of a Metal Surface in Conduction Theory**—H. A. Müser. (*Phil. Mag.*, vol. 45, pp. 1237–1246; December, 1954.) A discussion is presented in which the reflection of electrons striking the surface from inside is treated as a diffraction phenomenon.

537.5:061.3 1617

**Gaseous Electronics Conference in New York**—F. L. Jones. [*Nature* (*London*), vol. 175, pp. 154–155; January 22, 1955.] Papers presented at the conference held in October, 1954 are surveyed briefly. Physics of plasma and ionization processes leading to electrical breakdown were among the subjects discussed.

537.52 1618

**Study of Gaseous Discharges by Magnetic Resonance**—D. J. E. Ingram and J. C. Tapley. [*Research* (*London*), vol. 7, supplement, pp. S63–S64; December, 1954.] Measurements are reported which indicate that the technique of magnetic resonance may prove useful for studying the characteristics of low-pressure gas discharges.

537.523.5:621.396.822 1619

**Relaxation Oscillations and Noise from Low-Current Arc Discharges**—M. I. Skolnik and H. R. Puckett, Jr. (*Jour. Appl. Phys.*, vol. 26, pp. 74–79; January, 1955.) Noise measurements were made on arcs, using various electrode materials and gases, over the frequency range 0.1 mc–4.5 kmc. The results for Al electrodes in air are plotted in comparison with the computed spectrum. Pulses generated in the circuit formed by the series limiting resistor and the capacitance across the discharge are considered responsible for most of the noise.

537.525 1620

**Formative Time of the Cathodic Space Charge**—D. Brini, O. Rimondi and P. Veronesi. (*Nuovo Cim.*, vol. 12, pp. 915–922; December 1, 1954. In English.) A criterion for evaluating the formative time is based on observations of rise time and overshoot in intermittent discharges. Different kinds of intermittent discharges are reviewed.

537.525 1621

**On Intermittent Discharges in Air at Low Pressure**—D. Brini, O. Rimondi and P. Veronesi. (*Nuovo Cim.*, vol. 12, pp. 948–949; December 1, 1954. In English.) An experimental investigation is reported of the dependence of the discharge frequency on the gas pressure and electrode separation.

537.533 1622

**Forces between Parallel Electron Streams**—J. Webb. [*Elec. Rev.* (*London*), vol. 155, pp. 1037–1038; December 31, 1954.) A simplified fundamental analysis is presented, taking account of the forces of acceleration, es repulsion, em attraction, gravitation and any local fields. The results indicate that for all electron velocities up to the immediate threshold of the velocity of light the resultant force between parallel electron streams is repulsive.

537.56 1623

**Spectroscopic Studies of Highly Ionized Argon produced by Shock Waves**—H. E. Petschek, P. H. Rose, H. S. Glick, A. Kane and A. Kantrowitz. (*Jour. Appl. Phys.*, vol. 26, pp. 83–95; January, 1955.)

537.56 1624

**Kinetic Theory of Weakly Ionized Homogeneous Plasmas: Part 1**—M. Bayet, J. L. Delcroix and J. F. Denisse. (*Jour. Phys. Radium*, vol. 15, pp. 795–803; December, 1954.) A rigorous method is presented for integrating Boltzmann's equation by successive approximations for a weakly ionized gas subjected to an alternating electric field and a static magnetic field. The electron-velocity distribution function is developed with respect to the electric field, and the individual functions obtained are developed as spherical functions. The method is applied to investigation of a Lorentz-type gas. See also 2910 of 1954.

537.56:537.311.37 1625

**Electrical Conductivity of Highly Ionized Argon produced by Shock Waves**—Shao-Chi Lin, E. L. Resler and A. Kantrowitz. (*Jour. Appl. Phys.*, vol. 26, pp. 95–109; January, 1955.)

538.22 1626

**Antiferromagnetism and Ferrimagnetism of Non-Stoichiometric Compounds**—E. W. Elcock. (*Proc. Roy. Soc. A*, vol. 227, pp. 102–114; December 21, 1954.) A simple quantitative treatment is given of the magnetic properties of a substance containing vacancies, enabling the most important magnetic properties of many nonstoichiometric compounds to be interpreted.

538.3 1627

**Representation of Electromagnetic Fields of Any Frequency using the Energy-Quantum Model**—H. Zuhrt. (*Arch. Elekt. Übertragung*, vol. 8, pp. 565–577; December, 1954.) Development of theory presented previously (1004 of May). Wave propagation phenomena including reflection, interference and diffraction are explained in terms of the energy-quantum model; static fields are also considered. Voltage, current and characteristic impedance are related to mechanical quantities, and an appropriate system of dimensions and units is presented.

538.561:537.533 1628

**Čerenkov Radiation from Extended Electron Beams**—M. Danos. (*Jour. Appl. Phys.*, vol. 26, pp. 2–7; January, 1955.) Calculations are made of the radiation emitted by bunched beams passing close to dielectric surfaces; both plane and cylindrical geometries are considered. See also *Trans. IRE*, vol. MTT-2, no. 3, pp. 21–22; September, 1954.

538.566:535.42 1629

**Calculated Diffraction Patterns of Dielectric Rods at Centimetric Wavelengths**—C. Froese and J. R. Wait. (*Canad. Jour. Phys.*, vol. 32, pp. 775–781; December, 1954.) Calculations of the field behind the cylinder are made for the case of a normally incident wave with the electric vector (a) parallel and (b) perpendicular to the cylinder axis. The dielectric materials considered include polystyrene, lucite and tenite. The diffraction pattern is only slightly dependent on the dielectric constant of the rod. Small discrepancies between the calculated values and values obtained experimentally [3538 of 1954 (Wiles and McLay)] are probably due to resistive loss in the dielectric.

538.569.4 1630

**Absorption of Microwaves by Oxygen in the Millimeter Wavelength Region**—J. O. Artman and J. P. Gordon. (*Phys. Rev.*, vol. 96, pp. 1237–1245; December 1, 1954.) An account is given of an experimental investigation of the absorption at high and at low pressure. Theories advanced in explanation of the line-broadening effects are discussed.

538.569.4 1631

**Spectral Investigations in the Wavelength Range around 1 mm**—L. Genzel and W. Eckhardt. (*Z. Phys.*, vol. 139, pp. 578–598; December 20, 1954.) A description is given of the construction and method of operation of an infrared-type spectrometer which has been used with thermal radiator and receiver to obtain absorption spectra at wavelengths over 1 mm. The spectrum obtained for water vapor over the range 0.14–1.4 mm is shown as an example. Results obtained with HCN and $H_2S$ are also reported.

538.569.4:535.34:621.372.413 1632



**Variable-Frequency Microwave Cavity Spectrometer**—R. J. Collier. (*Rev. Sci. Instr.*, vol. 25, pp. 1205–1207; December, 1954.) Molecular-gas spectra at wavelengths around 10 cm are investigated using a system in which a single coaxial cavity serves as Stark or Zeeman modulation cell and as frequency reference unit.

548.0:53 1633

**Simplified Impurity Calculation**—G. F. Koster and J. C. Slater. (*Phys. Rev.*, vol. 96, pp. 1208–1223; December 1, 1954.) The methods developed previously (699 of March) are used to investigate the case of a local perturbation in a simple cubic lattice.

## GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 1634

**Abnormal Galaxies as Radio Sources**—B. Y. Mills. (*Observatory*, vol. 74, pp. 248–249; December, 1954.) Results of radio-telescope observations at Sydney indicate that galaxy NGC 1316 is probably a radio source, whereas NGC 1947 is probably not.

523.16 1635

**Observations of Galactic Radiation on a Wavelength of 33 cm**—J. F. Denisse, É. Leroux and J. L. Steinberg. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 278–280; January 17, 1955.] Report of measurements made using the Wurzburg mirror at Marcoussis. Results are presented in the form of an isophot map of the galaxy with intensities expressed as apparent temperature in arbitrary units. The 33-cm radiation is attributed principally to ionized hydrogen. A table giving the positions of localized sources includes five thought to have been observed for the first time.

523.16:523.72 1636

**Study of Solar R.F. Radiation on 9350 Mc/s around Sunset and Sunrise**—I. Kazes and J. L. Steinberg [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 493–495; January 31, 1955.] Report of measurements made at Paris, using a parabolic mirror of diameter 1.5 m. Graphs are presented showing the variation of refraction with solar angle and quasi-periodic intensity variations observed when the angle of elevation is less than 15 degrees. Attenuation due to absorption by oxygen is observed at angles less than about 20 degrees.

523.16:523.72 1637

**Fine Structure of Solar Radio Transients**—G. Reber. [*Nature* (*London*), vol. 175, p. 132; January 15, 1955.] Brief report of observations made during 1948–1950. Solar bursts observed at 480, 160 and 51 mc were found to be composed of numerous discrete pips with median duration approximately proportional to wavelength. The spectral width of a pip is a few per cent of its mean frequency. The frequency of occurrence of the pips was greatest at 160 mc.

523.16:621.396.677 1638

**Aerial Smoothing in Radio Astronomy**—R. N. Bracewell and J. A. Roberts. (*Aust. Jour. Phys.*, vol. 7, pp. 615–640; December, 1954.) Theoretical considerations show that the antenna does not register those spatial Fourier components of the true distribution of radio brightness having frequencies beyond a cut-off determined by the antenna aperture. Components of lower frequency are registered but their relative strengths are altered. The consequences are that (a) there are invisible distributions which produce no response when scanned by the antenna, and (b) in conducting a survey the measuring points must be closer together than half the period of the Fourier component at cut-off.

523.74/.75 1639

**Prominence Activity (1905–1952)**—R. Ananthakrishnan. (*Proc. Indian Acad. Sci.*, Section A, vol. 40, pp. 72–90; August, 1954.)

Charts prepared from observations made at Kodaikanal show the salient features of solar prominence activity during the last four sunspot cycles. A preliminary examination suggests that during the maximum phase of the cycle geomagnetic activity shows a better correlation with prominences than with sunspots.

523.752 1640
**The Emission of Radiation from Model Hydrogen Chromospheres: Part 2**—J. T. Jefferies and R. G. Giovanelli. (*Aust. Jour. Phys.*, vol. 7, pp. 574–585; December, 1954.) An improved method is presented for calculating the characteristics of the radiation field of $H\alpha$, $L\alpha$, $L\beta$, and the Lyman continuum emitted by model hydrogen atmospheres at kinetic temperatures of $10^4-2.5\times10^5$ degrees K. A useful application of the results would be in interpreting observations of prominences and flares.

550.385 1641
**Qualitative Explanation of the Commencement of Some Polar Magnetic Disturbances based on the Theory of Chapman and Ferraro**—G. Grenet. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 448–450; January 24, 1955.]

550.385:550.375 1642
**Geomagnetic and Geoelectric Variations**—J. G. Scholte and J. Veldkamp. (*Jour. Atmos. terr. Phys.*, vol. 6, pp. 33–45; January, 1955.) Analysis of records of geomagnetic pulsations, especially those with periods of $10-10^2$ seconds, indicates that they may be associated with ionospheric vibrations caused by a disturbance of the ionization equilibrium. Information regarding the distribution of ground conductivity is obtained from the relation between the magnetic pulsations and the associated transverse-electric field pulsations.

550.385.25:551.510.535 1643
**The Abnormal Variations of the Horizontal Magnetic Intensity at Huancayo (Peru)**—Z. Ibrahim. (*Proc. Math. Phys. Soc. Egypt*, vol. 5, pp. 21–24; 1953.) The magnetic field of the ionospheric current system arising from lunar atmospheric tidal motion is calculated for various points on the earth's surface. The daily variation of the horizontal component at Huancayo is found to be nearly twice as great as at Batavia, though the latitudes of the two points are nearly the same. Observations at a point between the geographic and magnetic equators in the northern hemisphere should help to explain this anomaly.

550.385.4:551.510.535 1644
**On the World-Wide Disturbance in $F_2$ Region**—T. Obayashi. (*Jour. Geomag. Geoelect.*, vol. 6, pp. 57–67; June, 1954.) The average disturbance in the value of $f_0F_2$ during magnetic storms is separated into the storm-time part $D_{st}(f_0F_2)$ and the local-time part $D_s(f_0F_2)$. These were calculated in the cases of 10 typical storms from data obtained from 40 stations in the northern hemisphere. Both their range and phase are correlated with the magnetic activity. The phase of $D_s(f_0F^2)$ during the active stage of the magnetic storm is almost entirely dependent on the local time but after the cessation of activity the disturbance moves with the rotating earth. See also *Jour. Radio Res. Labs., Japan*, vol. 1, pp. 41–50; June, 1954.

550.386"52" 1645
**An Analysis on the Diurnal Variation of the Terrestrial Magnetism, especially on the Day-Time-Variation of Geomagnetically Quiet Days**—M. Ota. (*Jour. Geomag. Geoelect.*, vol. 6, pp. 83–98; June, 1954.)

551.510.534 1646
**Vertical Distribution of Atmospheric Ozone at Longyearbyen, Spitzbergen (78 degrees N)**—S. H. H. Larsen. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 46–49; January 1955.)

551.510.535 1647
**The Constitution of the Upper Atmosphere and the Ionosphere Research Station of the Institute of Geophysics at Genoa**—M. Bossolasco and A. Masotto. (*Geofis. Met.*, vol. 2, pp. 80–86; September/December, 1954.) An outline of knowledge on the structure and electrical properties of the ionosphere is followed by a description of ionosphere sounding equipment recently put into operation and comprising pulse transmitter for the range 2–15 mc, double-superheterodyne receiver and cro recorder.

551.510.535 1648
**Calculation of the Collision Frequency in the Ionosphere**—L. Caprioli. (*R.C. Accad. naz. Lincei*, vol. 17, pp. 365–370; December, 1954.) It is shown that, if $z_0$ is the true reflection height corresponding to $\omega_0$, the lowest frequency of the sounding sweep, and $z_E$ is the true reflection height corresponding to the first critical frequency $\omega_E$ greater than $\omega_0$, then provided that the distribution of ion concentration and collision frequency are known for values of $z$ up to $z_0$ and the absorption is observed over the range $\omega_0-\omega_E$ the function $\nu(z)$ expressing the height distribution of collision frequency can be determined over the range $z_0-z_E$ by solving an integral equation due to Abel.

551.510.535 1649
**Turbulence in the Upper Ionosphere**—A. Maxwell. (*Phil. Mag.*, vol. 45, pp. 1247–1254; December, 1954.) "From the experimental data at present available it is shown that the Reynolds number in the upper $F$ region (300–400 km level) is of the order of 300. The region may therefore be turbulent. It is suggested that the high level diffracting screens which give rise to spread $F$ echoes and to radio star fading are caused by non-laminar flow, and that their non-appearance during the daylight hours may be due to the inhibition of turbulence by large temperature gradients, by lower drift velocities, or by an increase in the kinematic viscosity."

551.510.535 1650
**Motion of Clouds of Abnormal Ionization in the Auroral and Polar Regions**—E. L. Hagg and G. H. Hanson. (*Canad. Jour. Phys.*, vol. 32, pp. 790–798; December, 1954.) A study has been made of unusual types of echo exhibited by film records of rapid-succession sweep observations at several stations in Northern Canada. Three distinct types of echo are identified, probably corresponding to (a) horizontally moving $E_s$ clouds, (b) clouds descending vertically from the $F$ to the $E$ layer, and (c) clouds moving at extremely high velocities in the $E$ layer. Type (c) echoes were observed only at stations very close to the auroral zone maximum and may be due to sweeping of the auroral ionizing agent. Type (b) echoes also appear to be related to an auroral ionizing agent.

551.510.535 1651
**Information obtained from Ionization Charts**—R. Eyfrig, E. Harnischmacher and K. Rawer. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 446–448; January 24, 1955.] Ionization charts drawn up monthly for two zones covering (a) America and (b) Europe, Africa and Asia are discussed. The distributions observed generally confirm the theory of geomagnetic control of the $F_2$ layer [2898 of 1946 (Appleton)], but some exceptions are observed; there appears to be a lower limiting value for the height of the sun below which the geomagnetic control does not operate. The control is effective at latitudes as high as 70 degrees. Measurements at other times besides midday are required to study the effect. Asymmetry as between northern and southern hemispheres is observed. The north polar zone was studied in detail; the auroral $E$ zone appears to be centered on the geographic pole.

551.510.535 1652
**Velocity of Movement of Sporadic-E Clouds**—M. R. Kundu. (*Sci. and Cult.*, vol. 20, p. 303; December, 1954.) Curves of the diurnal variation of the $E_s$-layer critical frequency for four Japanese stations indicate a progressive retardation of the time of occurrence of the maximum on passing from the highest-latitude to the lower-latitude stations. A value of about 70 m/s is deduced for the horizontal component of the velocity of the $E_s$ clouds; this is of the same order as the velocity of $E$-layer winds found by various methods.

551.510.535 1653
**Interpretation of Measurements on the Ionosphere $F_1$ layer**—K. Rawer. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 331–333; January 17, 1955.] The maximum-ionization level of the $F_1$ layer is often located within the lower part of the $F_2$ layer, so that echo sounding yields a continuous curve for the $F_1$ and $F_2$ layers, with a more or less marked maximum or merely a point of inflection indicating the virtual height of the $F_1$ layer. In consequence, the critical frequency of this layer is not clearly defined. A correction is required, the sign of which depends on the theory accepted for the origin of the $F_2$ layer. Mohler's theory (1014 of 1941) gives results in good agreement with observations.

551.510.535 1654
**Lunar Tides in the Ionosphere $F_2$ Layer at Dakar**—F. Delobeau. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 222–224; January 10, 1955.] Hourly records of $f_0F_2$ and $h'F_2$ for the years 1950–1953 have been analyzed; maximum variations and times of occurrence of maxima are tabulated for the different seasons. The semidiurnal component is predominant except in summer; during winter the tidal variations of $f_0F_2$ and $h'F_2$ have amplitudes of $\pm0.2$ mc and $\pm5$ km respectively. The phases and amplitudes are considerably different from those observed at other tropical stations; anomalies may be due to the circumstance that the $F_2$ layer at Dakar is very thick in summer; the ionization maximum occurs at the equinoxes and not in winter.

551.510.535 1655
**On the Meridional Distribution of the Minimum Virtual Height of the $F_2$ Layer**—T. Shimazaki. (*Jour. Radio. Res. Labs. Japan*, vol. 1, pp. 15–25; March, 1954.) Results of an analysis of daily variations in $h'F_2$ from 35 observatories indicate the presence of two periodic terms: (a) the diurnal variation, which is dependent on the geographic latitude, and (b) the semidiurnal variation, which depends on the geomagnetic latitude. The former may be considered as the thermal effect, the latter as the tidal effect in the $F_2$ region.

551.510.535 1656
**The Effect of the Solar Tides and the Temperature Change on the Daily Variation in Electron Density and Height of the $F_2$-Layer**—T. Shimazaki. (*Jour. Geomag. Geolect.*, vol. 6, pp. 68–82; June, 1954.) The departure $\Delta n_m$ of the maximum electron density from the norm of a static Chapman region is inversely proportional to the temperature variation, except in the morning when the electron density is small. The semidiurnal variation in $\Delta n_m$ appears even in the case when account is taken of the diurnal temperature variation only. As regards $\Delta n_m$, the departure from the norm of the height of $n_m$, the result is similar to that obtained by Weiss (1348 of 1953) except that it is more strongly temperature dependent. There is some interaction between the temperature and the tidal effects.

551.510.535:523.3 1657
**Influence of the Moon on the Maximum Ionization of the Ionosphere E Layer**—E. Harnischmacher. [*Compt. Rend. Acad. Sci.*

(*Paris*), vol. 240, pp. 553–555; January 31, 1955.] An analysis of $f_0E$ variations over a period of four years, observed at six stations, indicates that after taking account of seasonal, solar-cycle and equation-of-time effects, there is evidence of a lunar variation of amplitude 1 per cent centered on the full moon.

551.510.535:523.72 1658
**Soft X Radiation from the Quiet Solar Corona**—G. Elwert. (*Z. Naturf.*, vol. 9a, pp. 637–653; July/August, 1954.) The mechanisms involved in the production of the soft *X*-rays are discussed, and the significance of this radiation for the formation of the normal ionosphere *E* layer is indicated. See also 726 of 1954.

551.510.535:621.317.3 1659
**Equipment for the Measurement of Changes of the Phase Path of Ionospheric Echoes**—H. Yuhara, T. Koseki and Y. Aono. (*Jour. Radio Res. Labs, Japan*, vol. 1, pp. 11–14; March, 1954.) The equipment described is built with units conventional in pulse applications and is developed from that described by Findlay (397 and 404 of 1952). Oscillograms of ground pulse beats and echo patterns of the *E* and *F* regions are shown.

551.510.535:621.396.11 1660
**Reflection Conditions for Vertical Propagation in the Ionosphere in the Presence of Collisions and of the Earth's Magnetic Field. Case of the E Layer**—Lepechinsky and Durand. (*See* 1767.)

551.510.535:621.396.11.029.51 1661
**The Development of an E-Region Model consistent with Long-Wave Phase-Path Measurements**—R. E. Jones. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 1–17; January, 1955.) An electron-density model is evolved by modification of the Chapman theory to include the effects of variable scale height, dissociation of $O_2$, and variable recombination. The model is checked against phase-height data for 150 kc and 2.4 mc and against $f_0E$ values.

551.510.535:621.396.812 1662
**Measurement of Attenuation in the Ionosphere**—Ochs. (*See* 1764.)

551.510.535:621.396.812.3 1663
**The Autocorrelation of Randomly Fading Waves**—Banerji. (*See* 1772.)

551.510.535(98):621.396.11 1664
**Statistical Studies of Polar Radio Blackouts**—J. W. Cox and K. Davies. (*Canad. Jour. Phys.*, vol. 32, pp. 743–756; December, 1954.) "A statistical study of high frequency radio blackouts in Canada is made from records taken at several ionosphere sounding stations. Both vertical incidence and communication data are examined to determine the geographical, seasonal, and diurnal distributions of the frequency of occurrence of blackout. It is found that blackouts are most abundant in the morning and that the time of maximum occurrence increases with increasing latitude."

551.578:621.396.11:621.396.96 1665
**The Microwave Properties of Precipitation Particles**—K. L. S. Gunn and T. W. R. East. (*Quart. Jour. Roy. Met. Soc.*, vol. 80, pp. 522–545; October, 1954.) "The theory of scattering and attenuation by rain, snow and cloud is reviewed and theoretical results are presented in the form of equations, tables and graphs, so that the radar response to meteorological particles can be calculated at six wavelengths (10, 5.7, 3.2, 1.8, 1.24 and 0.9 cm) and various temperatures. Particular emphasis is placed on developments since Ryde's comprehensive paper in 1946. Published experimental results are compared with the theory. All results computed from the theory are contained in Tables 4 and 5. The attenuation by water vapour and oxygen is given in an Appendix."

551.594.6 1666
**An Attempt to observe Whistling Atmospherics near the Magnetic Equator**—J. R. Koster and L. R. O. Storey. [*Nature* (*London*), vol. 175, pp. 36–37; January 1, 1955.] According to the theory of Barkhausen and Eckersley, the mode of propagation of whistling atmospherics is such that they should not occur near the magnetic equator. A report is given of observations made at Achimota, over the period from December, 1951 to March, 1954; no whistlers were detected, though other types of atmospheric were frequent.

## LOCATION AND AIDS TO NAVIGATION

621.396.93:551.594.6 1667
**Low-Frequency Direction Finder**—C. Clarke and V. A. W. Harrison. (*Wireless Eng.*, vol. 32, pp. 109–114; April, 1955.) A more detailed account of the instrument described previously [3226 of 1954 (Horner)].

621.396.933 1668
**Radio Installations of the Danish Airways System**—K. Svenningsen. [*Teleteknik* (*Copenhagen*), vol. 5, pp. 391–400; December, 1954.] The navigation aids and communication systems used are described.

621.396.96:551.578:621.396.11 1669
**The Microwave Properties of Precipitation Particles**—Gunn and East. (*See* 1665.)

621.396.962.33 1670
**Radar Receiver with Elimination of Fixed-Target Echoes**—H. Tanter. (*Elec. Commun.*, vol. 31, pp. 235–248; December, 1954.) English version of paper abstracted in 2407 of 1954.

## MATERIALS AND SUBSIDIARY TECHNIQUES

535.5 1671
**Design and Operation of Evapor-ion Pumps**—R. H. Davis and A. S. Divatia. (*Rev. Sci. Instr.*, vol. 25, pp. 1193–1197; December, 1954.) Operation of the pump described depends on the gettering action of continuously evaporated Ti in conjunction with ion pumping. The lowest pressure attainable is about $2.10^{-7}$ mm Hg. The dependence of pumping speed on the temperature of the gettering surface, the pressure, and the rate of evaporation of Ti is investigated.

535.215:537.311.33:546.817.23 1672
**Response Time of Photoconductivity of Lead Selenide**—L. Sosnowski and M. Chmielewski. (*Bull. Acad. Polon. Sci.*, Classe 3, vol. 1, nos., 3/4, pp. 119–121; 1953. In English.) An oscillographic method for investigating response times of less than 1 $\mu$s is described. An exponential timebase is used and the specimen is illuminated in synchronism by light pulses at repetition rates up to 50,000 per second. The response times of three different PbSe cells were 0.25, 0.35 and 0.9 $\mu$s within $\pm 0.1$ $\mu$s, their respective resistances and sensitivities being 23, 51, and 100 $k\Omega$, and 12, 15, and 40 arbitrary units.

535.215:537.311.33:546.817.23 1673
**Photoconductive and Photovoltaic Layers of Lead Selenide**—H. Checińska. (*Bull. Acad. Polon. Sci.*, Classe 3, vol. 1, nos. 3/4, pp. 123–135; 1953. In English.) The method of preparing PbSe layers exhibiting these effects is described and some results of determinations of limiting sensitivity and spectral sensitivity in the range 0.5–3.6 $\mu$ are given. At room temperature the sensitivity to radiation from an ordinary incandescent lamp is less than one tenth that of PbS.

535.37 1674
**The Shape of the Emission Bands of Luminescent Solids**—C. C. Vlam. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 443–446: December, 1954.)

535.37 1675
**Luminescence in High Polymers**—H. Hinrichs. (*Z. Naturf.*, vol. 9a, pp. 617–630; July/August, 1954.) An investigation of organic phosphors embedded in polystyrol.

535.37:537.311.33 1676
**Temperature Dependence of the Energy-Gap in ZnS**—C. Z. van Doorn. (*Physica*, vol. 20, pp. 1155–1156; December, 1954.) Measurements on a single crystal showed that the temperature variation of the energy gap varied between $4.6\times10^{-4}$ev degrees K at 77 degrees K and $8.5\times10^{-4}$ev degrees K at 800 degrees K.

535.376 1677
**The Edge Emission of ZnS, CdS and ZnO and its Relation to the Lattice Vibrations of these Solids**—F. A. Kröger and H. J. C. Meyer. (*Physica*, vol. 20, pp. 1149–1156; December, 1954.)

535.376:538.615 1678
**The Effect of Intense Magnetic Fields on Electroluminescent Powder Phosphors**—A. N. Ince. (*Proc. Phys. Soc.*, vol. 67, pp. 870–874; December 1, 1954.) No quenching of electroluminescence was observed in phosphors in magnetic fields of up to $1.3\times10^5$ oersted. This is contrary to the predictions of Destriau (110 of 1949). The significance of this result is discussed.

537.226 1679
**High-Frequency Polarization of a Spherical Body and of an Assemblage of Particles of a Perfect Dielectric**—A. Colombani. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 503–505; January 31, 1955.] Analysis is given for a single particle, using Maxwell's equations in spherical—coordinate form. If the permittivity is real and very high the electric field and es energy within the particle may become very large. An expression is derived for the apparent susceptibility of an assemblage of the particles.

537.227:546.431.824–31 1680
**Observation of Paramagnetic Resonances in Single Crystals of Barium Titanate**—A. W. Hornig, E. T. Jaynes and H. E. Weaver. (*Phys. Rev.*, vol. 96, p. 1703; December 15, 1954.)

537.311.33 1681
**A Possible Mechanism of the Scattering of Current Carriers in Semiconductors**—T. A. Kontorova. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2217–2220; December, 1954.) The unusual form of the temperature dependence of the mobility of current carriers for certain semiconductors has not yet been satisfactorily explained. By taking account of the interaction of the current carriers not only with acoustic vibrations of the lattice but also with optical vibrations which are possible when the adjacent particles vibrate in opposite phase, the theoretical results are brought into agreement with experiment.

537.311.33 1682
**Use of *p-n* Junctions for Solar Energy Conversion**—E. S. Rittner. (*Phys. Rev.*, vol. 96, pp. 1708–1709; December 15, 1954.) Calculations of the efficiency to be expected from a Si cell lead to results in agreement with those of Cummerow (145 of January), except that for maximum power conversion efficiency the energy gap of the semiconductor should be about 1.5–1.6 ev; thus AlSb is preferable to Si in this respect.

537.311.33 1683
**Determination of the Optical Constants of Type-$A^{III}B^{V}$ Semiconductors at Infrared Wavelengths**—F. Oswald and R. Schade. (*Z. Naturf.*, vol. 9a, pp. 611–617; July/August, 1954.) Reflection and transmission measurements were made on a series of compounds of elements of Groups III and V of the periodic system, at wavelengths from 0.8 to 15.2 $\mu$. Values found for absorption constant, refractive index and

width of energy gap are tabulated; energy-gap values determined by other workers, using electrical methods, are given for comparison.

537.311.33 1684

**Theory of Conduction in Isotropic Semiconductors**—O. Madelung. (*Z. Naturf.*, vol. 9a, pp. 667–674; July/August, 1954.) The classical theory is developed and the equations for electrical and thermal current density are established. From these equations the coefficients of all the galvanomagnetic, thermoelectric and thermomagnetic effects are derived. Expressions for these coefficients are tabulated for small values of magnetic field strength, for degenerate and nondegenerate semiconductors with given scattering mechanism.

537.311.33:537.323 1685

**Theory of the Thermoelectric Power of Semiconductors**—C. Herring. (*Phys. Rev.*, vol. 96, pp. 1163–1187; December 1, 1954.) The marked rise of thermoelectric power which has been observed in some semiconductors at temperatures below room temperature [1093 of 1954 (Frederikse)] has been attributed to the effect of thermal lattice vibrations, the total thermoelectric power being composed of an electron term and a phonon term. This theory is developed; the results are supported by experimental data on Ge.

537.311.33:546.23 1686

**Crystallization of Selenium under Pressure**—D. N. Nasledov and P. T. Kozyrev. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2124–2135; December, 1954.) The effect of pressure up to 4,000 atm was investigated. Pressure retards crystallization at low temperatures (below 110 degrees C.) and accelerates it at high temperatures (above 170 degrees C.). A curve shows the relation between temperature and pressure at the point of fusion. Large single crystals (up to 5 mm) were formed from liquid Se under pressure. The electrical properties of the crystallized material were investigated. Possible mechanisms explaining the observed phenomena are discussed.

537.311.33:546.24 1687

**Tellurium Single Crystals prepared by the Czochralski Process**—J. Weidel. (*Z. Naturf.*, vol. 9a, p. 697; July/August, 1954.)

537.311.33:[546.28+546.289 1688

**Theory of Donor and Acceptor States in Silicon and Germanium**—C. Kittel and A. H. Mitchell. (*Phys. Rev.*, vol. 96, pp. 1488–1493; December 15, 1954.) The applicability of the Wannier equation to the donor states in Si and Ge is examined with particular reference to the multiple energy minima in the conduction band. The theory is extended to include degenerate bands, and it is shown that the Wannier equation is to be replaced by a set of coupled wave equations. The theory is also applied to acceptor states. Theoretical and experimental results are in fair agreement for both donors and acceptors.

537.311.33:546.28 1689

**Single Crystals and *p-n* Stratified Crystals of Silicon**—H. Kleinknecht and K. Seiler. (*Z. Phys.*, vol. 139, pp. 599–618; December 20, 1954.) Measurements are reported on specimens produced as described previously [422 of 1953 (Kleinknecht)]. The voltage dependence of the capacitance of the *p-n* specimens indicates a linear variation of impurity-center concentration. The diffusion voltages and the slope of the forward characteristic are not consistent with a carrier concentration conforming to the Boltzmann law at the boundary layer. The backward currents do not fit Shockley's theory; the discrepancy is explained by assuming the presence of traps acting as recombination centers. The concentration and effective cross section of these is evaluated from the increase of capacitance at low frequencies.

537.311.33:546.28:621.314.7 1690

**Electronic Behaviour of Certain Grain Boundaries in Perfect Crystals**—H. F. Mataré. (*Z. Naturf.*, vol. 9a, p. 698; July/August, 1954.) Some experiments are reported on the effect of incorporating structural inhomogeneities in semiconductor crystals; grain boundaries with highly nonlinear resistance were studied. A current gain of 50 was obtained with a three-electrode arrangement using a 5-Ω cm Si specimen. A practical form of grain-boundary transistor is illustrated.

537.311.33:546.289 1691

**The Production of Germanium from Zinc Residues**—[*Metallurgia (Manchester)*, vol. 50, pp. 277–278; December, 1954.] Some details are given of methods used in industry and in laboratory assay work for extraction of Ge and for the preparation of the metal. A different recovery scheme is required for each different complex residue. Poison hazards are mentioned.

537.311.33:546.289 1692

**An Observation of Circular Patterns in the Vicinity of Small-Area Alloyed Germanium *p-n* Junctions**—N. Holonyak, Jr. (*Jour. Appl. Phys.*, vol. 26, pp. 121–123; January, 1955.) Rings observed on the crystal surface in the neighborhood of small alloyed junctions are explained in terms of the action of bubbles accompanying the etching process.

537.311.33:546.289 1693

**Optical Studies of Injected Carriers: Part 3—Infrared Absorption in Germanium at Low Temperatures**—R. Newman. (*Phys. Rev.*, vol. 96, pp. 1188–1190; December 1, 1954.) Measurements are reported which confirm that carriers produced by injection and by impurity doping produce the same absorption effect. Anomalies observed in samples containing Fe may be due to trapping effects. Part 2: 1084 of 1954.

537.311.33:546.289 1694

**A Photoelectric Method for the Simultaneous Determination of Lifetime and Mobility of Injected Current Carriers in Semiconductors**—G. Adam. (*Z. Naturf.*, vol. 9a, pp. 607–611; July/August, 1954.) Additional carriers are formed by illumination with an elongated spot of light which is swept along the specimen. A probe in the vicinity of the illuminated region picks up a voltage proportional to the additional carrier concentration; this is recorded oscillographically. Methods of evaluating the oscillograms are described and illustrated by examples. Carrier lifetime and diffusion constant can be determined from a single oscillogram. Results are presented for Ge specimens containing various impurities.

537.311.33:546.289 1695

**Determination of the Relation between Mobility and Diffusion Coefficient for Photoholes in *n*-Type Germanium**—S. M. Ryvkin. (*Zh. tekh. Fiz.* vol. 24. pp. 2136–2149; December, 1954.) Detailed report of an experimental investigation of the diffusion and drift of photoelectrically produced holes. The results confirm the theoretical prediction that Einstein's formula (1) relating mobility to diffusion coefficient applies in this case. The stationary distribution of minority carriers in a partially illuminated semiconductor is discussed in an appendix.

537.311.33:546.289 1696

**Resistivity and Hall Effect of Germanium at Low Temperatures**—C. S. Hung and J. R. Gliessman. (*Phys. Rev.*, vol. 96, pp. 1226–1236; December 1, 1954.) Report of an extensive experimental investigation at temperatures from room temperature to that of liquid He. Anomalies in the Hall-constant curves at low temperature are explained on the assumption of small but finite mobility of carriers in the impurity states; the contribution of these carriers to the total conduction becomes important at low temperature because the concentration of carriers in the conduction band is then very low.

537.311.33:546.289 1697

**Transverse Hall and Magnetoresistance Effects in *p*-Type Germanium**—R. K. Willardson, T. C. Harman and A. C. Beer. (*Phys. Rev.*, vol. 96, pp. 1512–1518; December 15, 1954.) Calculations based on modification of the two-band model to take account of a small number of high-mobility holes give values for the magnitude, temperature dependence and magnetic-field dependence of Hall and magnetoresistance effects in good agreement with experimental results. The importance of making measurements both at large and small magnetic-field strengths is indicated.

537.311.33:546.431–31:535.215 1698

**Ultraviolet Absorption in Barium Oxide Films**—K. Okumura. (*Phys. Rev.*, vol. 96, pp. 1704–1705; December 15, 1954.) Measurements are reported briefly; results are compared with those of Tyler and Sproull (148 of 1952).

537.311.33:546.48.241.1 1699

**Semiconducting Cadmium Telluride**—D. A. Jenny and R. H. Bube. (*Phys. Rev.*, vol. 96, pp. 1190–1191; December 1, 1954.) In general, *n*-type specimens are obtained by adding Group-III or Group-VII impurities and *p*-type by adding Group-I or Group-V, the activation energies of the *p*-type impurities being much larger than those of the *n*-type impurities. The intrinsic energy gap is about 1.45 ev. Electron and hole mobilities are at least 30 cm per v/cm.

537.311.33:546.482.21:535.215 1700

**Determination of Trap Distribution from Interrupted-Illumination Measurements on Photoconductive Cadmium Sulphide Single Crystals**—E. A. Niekisch. (*Z. Naturf.*, vol. 9a, pp. 700–701; July/August, 1954.)

537.311.33:546.682–31 1701

**Investigations of Electrical and Photoelectric Conductivity of Thin Films of Indium Oxide**—G. Rupprecht. (*Z. Phys.*, vol. 139, pp. 504–517; December 20, 1954.) The conductivity of thin films of the *n*-type semiconductor $In_2O_3$ is markedly dependent on the surrounding atmosphere. Specimens prepared by evaporating In on to a quartz plate and heating in air at 700 degrees–1,000 degrees C. had thicknesses between 50 and 250 mμ and conductivities between 10 and $10^{-6}$ (Ω. cm)$^{-1}$. Measurements are reported of the temperature variation of conductivity and of the effect of an oxygen atmosphere. Above 500 degrees C. a balance is reached between the concentration of impurity centers and the external oxygen concentration. The photoelectric conductivity exhibits an irreversible increase in vacuum.

537.311.33:546 682.231 1702

**Electrical and Optical Properties of Indium Selenide**—R. W. Damon and R. W. Redington. (*Phys. Rev.*, vol. 96, pp. 1498–1500; December 15, 1954.) Measurements mainly on single crystals are reported. The optical absorption edge was not sufficiently well defined for the energy gap to be estimated unambiguously. The photoconductive response was mostly in the visible region, the sensitivity being comparable with that of grey Se. Attempts to determine the carrier type gave conflicting results; the material may not be a single-carrier-type semiconductor, at least within the surface region.

537.311.33:546.86:539.234 1703

**Fermi Level in Amorphous Antimony Films**—E. Taft and L. Apker. (*Phys. Rev.*, vol. 96, pp. 1496–1497; December 15, 1954.) Photoelectric experiments are reported which confirm that the amorphous form of Sb common in thin evaporated films is a semiconductor with Fermi level about 0.1 ev above the occupied band.

538.221 1704

**Investigations of Irreversible Magnetization and After Effect**—J. Kranz. (*Z. Phys.*, vol. 139, pp. 619–637; December 20, 1954.) An experimental arrangement is described for measuring the Barkhausen jumps, produced on reversing the field applied to a ferromagnetic specimen, by amplitude-analyzing and counting the pulses induced in a solenoid. Results for various materials are presented and discussed.

538.221 1705

**Large Magnetic Kerr Rotation in BiMn Alloy**—B. W. Roberts and C. P. Bean. (*Phys. Rev.*, vol. 96, pp. 1494–1496; December 15, 1954.) Brief illustrated note on observations of ferromagnetic domain patterns in large grains of BiMn.

538.221 1706

**Kinetics of Magnetization in Some Square Loop Magnetic Tapes**—C. P. Bean and D. S. Rodbell. (*Jour. Appl. Phys.*, vol. 26, pp. 124–125; January, 1955.) Curves showing flux reversal characteristics for permalloy tapes are discussed in relation to the domain-wall processes.

538.221 1707

**The Effect of Particle Shape Variations on the Coercivity of Iron-Oxide Powders**—W. P. Osmond. (*Proc. Phys. Soc.*, vol. 67, pp. 875–882; December 1, 1954.) Calculations show that a Gaussian distribution of particle shape factors about an observed mean value can satisfactorily explain the difference between measured values of coercivity of dispersed magnetic powders and the theoretical values for assemblages of identical particles.

538.221:621.318.134:537.311.33 1708

**The Nature of the Insulating Layers in Ferromagnetic Semiconductors**—R. Parker. (*Physica*, vol. 20, pp. 1314–1315; December, 1954.) Recent experimental results [3604 of 1954 (Volger)] can be explained on the following assumptions:—(a) that the appearance of spontaneous magnetization is the cause of the deviation from the normal relation between resistivity and temperature, and (b) that the insulating layer may be identified with the region in the material that is not spontaneously magnetized.

538.221:621.318.134 1709

**Saturation Magnetization and Crystal Chemistry of Ferrimagnetic Oxides**—E. W. Gorter. (*Philips Res. Rep.*, vol. 9, pp. 295–320, 321–365 and 403–443; August–December, 1954.) A thesis in which measurements of saturation magnetization, $\sigma$, against temperature, $T$, are reported for various mixed crystal oxides with spinel structure. Results are in agreement with Neel's theory; some of the anomalous $\sigma/T$ curves predicted have been found. Single ferrites investigated of the type $Me^{II}Fe_2^{III}O_4$ belong to a group with complete parallelism of the ionic moments inside each sublattice; mixed crystals of the type $Me_{1-a}Zn_aFe_2O_4$ with $a>0.4$ belong to a group with angles between the ionic moments inside one of the sublattices; for $Ca_{0.35}Zn_{0.65}Fe_2O_4$ the magnetic moment is higher than that of any MgZn ferrite; this is discussed with reference to Anderson's theory. Other materials investigated are ferrimagnetic spinels containing Ti and Al, and ferrimagnetic oxides containing Cr.

538.221:621.318.134 1710

**Low-Temperature Acoustic Relaxation in Ni-Fe Ferrites**—M. E. Fine and N. T. Kenney. (*Phys. Rev.*, vol. 96, pp. 1487–1488; December 15, 1954.) "An acoustic relaxation effect occurs near 40 degrees K in $Ni_{0.75}Fe_{2.25}O_4$ and is attributed to a stress-induced change in distribution of $Fe^{++}$ and $Fe^{+++}$ similar to that occurring in magnetite. The process involves electron diffusion. The activation energy is between 0.026 and 0.055 eV per electron jump."

538.221:621.318.134 1711

**Magnetic and Crystalline Behavior of Certain Oxide Systems with Spinel and Perovskite Structures**—L. R. Maxwell and S. J. Pickart. (*Phys. Rev.*, vol. 96, pp. 1501–1505; December 15, 1954.) Experiments are reported in which nonmagnetic trivalent ions were substituted for $Fe^{3+}$ in Ni ferrites.

538.221:621.318.134 1712

**Conference on Ferrites, Leningrad, 1st–5th February 1954**—(*Bull. Acad. Sci. URSS, sér. phys.*, vol. 18, pp. 307–416 and 419–520; May/June and July/August, 1954.) The text is presented of more than 20 papers covering theoretical and experimental investigation subjects discussed include Faraday effect at centimeter wavelengths and temperature dependence of electrical properties of ferrites.

538.221:681.142 1713

**Magnetic Materials for Digital-Computer Components: Part 1—A Theory of Flux Reversal in Polycrystalline Ferromagnetics**—N. Menyuk and J. B. Goodenough. (*Jour. Appl. Phys.*, vol. 26, pp. 8–18; January, 1955.) Output-voltage waveforms of computer storage elements are consistent with the assumption that the flux reversal is attributable to the creation and growth of 180 degree Bloch walls. A switching coefficient is defined having one component dependent on eddy current and another depending on spin relaxation; for ferrites and thin metal tapes the first of these components is much smaller than the latter. Consideration of various parameters involved indicates that it is better to produce hysteresis-loop squareness by grain orientation or magnetic anneal than by application of external stress or variation of chemical composition.

538.23 1714

**A Relation between Hysteresis Coefficient and Permeability: Part 2—Further Experimental Results**—M. Kornetzki. (*Z. Angew. Phys.*, vol. 6, pp. 547–550; December, 1954.) The investigation on ferrites reported earlier (756 of 1953) was extended to cover permanent magnets and various Fe, Fe-Si, Fe-Si-Al and Ni-Fe alloys with initial permeability, $\mu$, between 1.25 and 120,000 and hysteresis coefficient, $h$, between 0.65 and 9,000,000 cm/ka. A double-logarithmic plot of $h/(\mu-1)$ against $(\mu-1)$ shows that the points for most of the materials lie between a pair of parallel lines of slope 1.15 and separation, measured on the $h/(\mu-1)$ scale, corresponding to a ratio of 40. All groups, except the Fe and Fe-Si alloys, include low-hysteresis materials for which $h/(\mu-1)\approx 3$ cm/ka. Values of $\mu$ and $h$ are tabulated for over 40 materials, and graphs are plotted relating $(\mu-1)$ to $h$, $h/(\mu-1)$, and $h/(\mu-1)^2$.

621.315.613.1:537.529 1715

**Phenomena preceding Dielectric Breakdown in Mica**—B. Fallou (*Rev. Gén. Élect.*, vol. 63, pp. 643–653; November, 1954.) Report of an experimental investigation. Oscillograms are reproduced and discussed in relation to charge conditions at the surfaces of separation in the mica.

621.315.616:537.226 1716

**Dielectric Breakdown of Thermosetting Laminates**—N. A. Skow. (*Mod. Plast.*, vol. 32, pp. 152, 240; December, 1954.) A report is presented on short-time and endurance tests on laminates bonded with phenolic resin. The grades tested included three based on paper and one each on asbestos, cotton fabric, glass and nylon. The variation of the dielectric strength with temperature, direction of applied field, thickness of laminae and conditioning of the specimens is tabulated and some results are also presented graphically. The most suitable grades for use under various conditions (e.g. high humidity) are indicated.

621.315.614.4 1717

**Forest Products Research Special Report No. 8. The Dielectric Properties of Wood [Book Review]**—R. F. S. Hearmon and J. N. Burcham. Publishers: H. M. Stationery Office, London, 1954, 19 pp., 1s. 6d. (*Elec. Times*, vol. 126, p. 848; December 9, 1954.) An investigation of the influence of grain direction, density and moisture content on the permittivity and loss tangent of 12 species of wood over the frequency range 2 kc-60 mc is reported.

## MATHEMATICS

517.5 1718

**The Approximation to a Characteristic Function by its Fourier Series**—D. Dugué. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 151–152; January 10, 1955.]

517.512.2 1719

**Moving-Strip Fourier Analyzer**—H. J. Grenville-Wells. (*Rev. Sci. Instr.*, vol. 25, pp. 1156–1161; December, 1954.) A modified form of the device described by Robertson (*Phil. Mag.*, vol. 21, pp. 176–187; January, 1936.) Two-dimensional and three-dimensional summations can be performed.

517.564.3 1720

**The Asymptotic Expansion of Bessel Functions of Large Order**—F. W. J. Olver. (*Phil. Trans. A.*, vol. 247, pp. 328–368; December 28, 1954.)

517.9 1721

**The Asymptotic Solution of Linear Differential Equations of the Second Order for Large Values of a Parameter**—F. W. J. Olver. (*Phil. Trans. A*, vol. 247, pp. 307–327; December 28, 1954.)

519.272.119 1722

**Qualitative Evaluation of Correlation Coefficients from Scatter Diagrams**—T. M. Burford. (*Jour. Appl. Phys.*, vol. 26, pp. 56–57; January, 1955.) Analysis shows that the result previously obtained by Sugar (2457 of 1954) for the case of a Gaussian distribution can be applied to any distribution.

## MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74)+621.396.91 1723

**Standard Frequency Transmissions**—L. Essen. (*Proc. IEE*, Part B, vol. 102, pp. 173–178; March, 1955.) Discussion on 3289 of 1954.

621.3.018.41(083.74) 1724

**Standard Frequency Transmission Equipment at Rugby Radio Station**—H. B. Law. (*Proc. IEE*, Part B, vol. 102, pp. 166–173; March, 1955. Discussion, pp. 173–178.) "The transmissions are controlled by one of three highly stable oscillator-clock chains, which are checked daily in terms of a frequency standard, and are continuously intercompared in order that instability may be quickly detected. Automatic shut-down features are incorporated to reduce the risk of broadcasting incorrect frequencies under fault conditions. The vital parts of the equipment are protected against mains failure. The transmitted frequencies are normally kept within 1 part in $10^8$ of the currently-assessed nominal frequency, based on predicted clock performance, so as to keep within a tolerance of $\pm 2$ parts in $10^8$ in terms of finally-corrected time determinations. Day-to-day frequency variations are usually less than $\pm 2$ parts in $10^9$."

621.3.081.41(083.74):621.317.761 1725

**The Standard Frequency Monitor at the National Physical Laboratory**—J. McA. Steele. (*Proc. IEE*, Part B, vol. 102, pp. 155–165; March, 1955. Discussion, pp. 173–178.) Description of equipment for the automatic measurement and recording of the MSF standard frequency transmissions on 60 kc, 2.5 mc, 5 mc and 10 mc and of the Droitwich 200-kc transmission. The measurements on Droitwich and on MSF 60 kc are made by an extension of the established method for intercomparison of the N.P.L. frequency standards; the standard deviations for measurements lasting a few seconds are 3–4 parts in $10^9$ and 2–3 parts in $10^9$ respectively. "Measurements of the phase of the 1 c/s pulse modulation on MSF 60 kc/s can be made with very high precision. In daily comparisons, the scatter of a group of 60 readings does not usually exceed 30 $\mu$s. By a process involving frequency changing by continuous phase shifting, the received frequencies of MSF 2.5 and 5 Mc/s are recorded on a frequency meter of range $\pm 1$ c/s, the discrimination being 2 parts and 1 part in $10^8$ at these frequencies, respectively. The records obtained show well-defined diurnal variations on 2.5 Mc/s, particularly at sunrise, where the deviations may amount to several parts in $10^7$. The received frequency of 5 Mc/s is subject to continuous variations over a range of about 1 part in $10^7$ during daylight hours; in darkness larger changes are recorded."

**621.3.018.41(083.74):621.373.4.029.5 1726**

**Quartz Resonator Servo: a New Frequency Standard**—N. Lea. (*Marconi Rev.*, vol. 17, pp. 65–73; 3rd Quarter, 1954.) The response to an applied oscillator signal of a bridge detector circuit including a quartz crystal resonator is used to control a servodrive-operated variable capacitor in the oscillator tuned circuit. In an experimental instrument, using a 5-mc crystal, frequency variations were reduced to less than 1 part in $10^{10}$ for variations of $+20$ per cent in the applied hv, a lv change from 5 to 6 v, or an oscillator capacitance change which would, if uncorrected, cause a change of 200 parts in $10^{10}$. A rate of frequency correction of 1 part in $10^9$ per second was easily achieved. The circuit diagram given is briefly commented on. Experimental results are presented graphically.

**621.3.018.41(083.74):621.396.82 1727**

**Effect of Interference by Other Standard-Frequency Transmissions upon the Accuracy of Frequency Calibration by a Standard-Frequency Transmission**—K. Matsumoto, T. Nagatake and Y. Suguri. (*Jour. Radio Res. Labs, Japan*, vol. 1, pp. 41–48; March, 1954.) Experimental results indicate that interference with the JJY standard-frequency transmissions by station WWVH has a negligible effect on the accuracy of frequency calibrations when the reference beat method is used.

**621.3.018.41(083.74):621.396.91 1728**

**An Experiment on the Types of Time Signals superposed on the Standard-Frequency Transmission**—K. Matsumoto and T. Nagatake. (*Jour. Radio Res Labs, Japan*, vol. 1, pp. 49–56; March, 1954.) Coding of a JJY standard-frequency transmission by a 20-ms interruption of the 1-kc tone each second and a 200-ms interruption each minute, was compared with coding by the CCIR method of substituting a 1.4-kc tone for 5 ms each second. Using simple receivers at 1,000 km from the transmitter, the former method was found to be the more suitable one.

**621.316.84(083.74) 1729**

**A 1-100-Ω Build-Up Resistor for the Calibration of Standard Resistors**—B. V. Hamon. (*Jour. Sci. Instr.*, vol. 31, pp. 450–453; December, 1954.) A build-up resistor circuit developed at the Australian National Standards Laboratory comprises 10-Ω manganin resistors in series-parallel. The estimated accuracy when used as a ratio device is of the order of 1 part in $10^8$. Test results are tabulated and the construction is illustrated.

**621.317.3:621.315.61 1730**

**Measurements of Electrical Polarization in Thin Dielectric Materials**—R. W. Tyler, J. H. Webb and W. C. York. (*Jour. Appl. Phys.*, vol. 26, pp. 61–68; January, 1955.) A method suitable for measuring electrical effects such as are produced in film moving over a roller system consists in arranging the dielectric material in contact with a grounded metal backing plate and placing a field meter at a short distance in front of the dielectric. Tests made with a short strip of cellulose acetate film without the emulsion coating are described.

**621.317.321:538.632 1731**

**Apparatus for Measurement of Hall Effect and Magnetic Change of Resistance with Alternating Current**—K. A. Muller and J. Wieland. (*Helv. Phys. Acta*, vol. 27, pp. 690–696; December, 31, 1954. In English.) The null method described is operated at 73 cps and is suitable for detecting voltage changes down to $2\times10^{-8}$v in specimens of resistance between $10^{-6}$ and $10^{-2}$Ω. The pd across the specimen or the Hall emf is determined by compensating it by an equal and opposite voltage derived from a fixed resistor in series with the specimen. The null detector comprises an amplifier, filter, phase discriminator and and an indicator instrument which is either a cro or an aperiodically damped galvanometer.

**621.317.331:621.385.2 1732**

**A Method for recording Logarithmic Variations of Resistance**—H. A. Vodden. (*Jour. Sci. Instr.*, vol. 31, pp. 475–476; December, 1954.) A simple circuit of an ohm-meter recording resistance logarithmically is based on the fact that the logarithm of the anode current of a diode is proportional to the anode voltage over a range of negative voltage values. The useful range of the instrument described is between about $2\times10^2$ and $10^9$Ω.

**621.317.336:621.317.755 1733**

**Visual Impedance-Matching Equipment**—R. Dalziel and A. Challands. (*Wireless Eng.*, vol. 32, pp. 99–107; April, 1955.) A cro method is described for indicating the degree of match between a load and a cable, as e.g. in antenna feeding. The test oscillator is mechanically swept over the whole frequency range of 80–250 mc. The oscilloscope face is calibrated in terms of swr. Impedance measurements made with the equipment have yielded results in good agreement with those obtained by other methods. See also 2852 of 1948 (Libby).

**621.317.361:621.385.029.65 1734**

**Cold Measurements of 8 mm Magnetron Frequency and Pulling Figure**—Barrington. (*See* 1839.)

**621.317.382:538.632:537.311.33 1735**

**The Application of the Hall Effect in a Semiconductor to the Measurement of Power in an Electromagnetic Field**—H. E. M. Barlow. (*Proc. IEE*, Part B, vol. 102, pp. 179–185; March, 1955. Discussion, pp. 199–203.) Analysis shows that the mean value in time of the Hall emf is a direct measure of the power traversing the semiconductor in steady or varying fields. Residual rectifier effects are eliminated by operating with a strong magnetic field. Various types of wattmeter embodying the principle are described. Experiments with an *n*-type Ge crystal mounted between the inner and outer conductors of a coaxial line indicate that the Hall effect is approximately the same at 50 cps and at 300 mc, so that instruments for use at high frequencies can be calibrated at low frequency.

**621.317.443 1736**

**An Improved Precision Permeameter**—C. D. Mee and R. Street. (*Proc. IEE*, Part II, vol. 101, pp. 639–642; December, 1954.) A modified form of the dc permeameter described by Armour et al. (3499 of 1952) uses a saturable-inductor type of field-measuring device to give automatic and continuous indication of the required compensating-coil current at all points on the $B/H$ curve. Values of $H$ from $10^{-3}$ oersted upward can be measured.

**621.317.715:621.383.2 1737**

**Photodianode and Galvanometer Feedback**—L. Deloffre, É. Pierre and J. Roig. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 59–61; January 3, 1955.] Analysis is presented relevant to galvanometer measurements using the twin-anode photocell device previously described (2543 of 1954 and back reference) in a feedback arrangement. The galvanometer sensitivity can be multiplied by a factor as great as 10 in this way.

**621.317.72:621.3.018.3 1738**

**Two-Frequency Waveforms: Effects on Rectifier Instruments**—J. E. Parton and W. D. Sutherland. (*Trans. Soc. Instrum. Technol.*, vol. 6, pp. 147–161; December, 1954.) Asymmetrical waveforms resulting from the presence of even harmonics are included in this study. For a waveform with two frequency components, a voltmeter with full-wave metal rectifier is found to read within 2 per cent of the value calculated on the assumption of perfect rectification. The mean value differs from that given by a rms instrument in being dependent on the relative phase of the two components.

**621.317.733:621.317.4 1739**

**Mutual Inductance Bridge and Cryostat for Low-Temperature Magnetic Measurements**—R. A. Erickson, L. D. Roberts and J. W. T. Dabbs. (*Rev. Sci. Instr.*, vol. 25, pp. 1178–1182; December, 1954.)

**621.317.755:621.314.7 1740**

**An Alpha Plotter for Point-Contact Transistors**—T. P. Sylvan. [*Elec. Engng.*, (*New York*), vol. 73, pp. 1094–1098; December, 1954.] Description, with detailed parts list, of a cro test set.

**621.317.78.029.5/.64 1741**

**Broadband R.F. Power Meters**—I. Strauss. *Radio-Electronic Engng.* vol. 23, pp. 10–11, 36; December, 1954.) Equipment for measuring average rf power from 5 $\mu$w to 5 w in the frequency range 20 mc to 10 kmc is briefly described; three frequency sub-ranges are covered by separate instruments. The energy dissipated in a bolometer element in a Wheatstone bridge is kept constant by (a) varying the dc current through it, and (b) attenuating the rf energy.

**621.317.78.029.65 1742**

**A Calorimeter for Power Measurements at Millimeter Wavelengths**—W. M. Sharpless. (*Trans. IRE*, vol. MTT-2, pp. 45–47; September, 1954.) Description of an instrument suitable for measuring power of the order of 1 mw, in which equal temperature rises are produced in two waveguide-section power absorbers, one of which is heated by dc and the other by the rf power.

**621.317.784 1743**

**Audio-Frequency Power Measurements by Dynamometer Wattmeters**—A. H. M. Arnold. (*Proc. IEE*, Part B, vol. 102, pp. 192–199; March, 1955. Discussion, pp. 199–203. The screening necessary to obtain accuracy in the upper af range comparable to that at power frequencies is discussed. An account is given of methods used at the NPL to calibrate wattmeters. The useful upper frequency limit is taken as the frequency at which a significant deflection is obtained with voltage only or current only applied to the terminals.

**621.317.784:538.632:537.311.33 1744**

**The Design of Semiconductor Wattmeters for Power-Frequency and Audio-Frequency**

**Applications**—H. E. M. Barlow. (*Proc. IEE*, Part B, vol. 102, pp. 186–191; March, 1955. Discussion, pp. 199–203.) Design details and performance characteristics are given for two wattmeters based on the Hall effect in semiconductors (1735 above), for use at frequencies up to 150 cps and 20 kc respectively. The power-frequency instrument incorporates an iron-cored magnetizing coil; the af instrument uses air-cored coils and includes screening arrangements. These instruments offer advantages over other types for measurements of high power.

621.37.029.6.049.001.4 1745

**A Surface-Texture Comparator for Microwave Structures**—A. F. Harvey. (*Proc. IEE*, Part B, vol. 102, pp. 219–222; March, 1955.) The dependence of the attenuation coefficient of microwave components on the relation between surface roughness and skin depth is discussed and a description is given of a simple comparator scale covering the various classes of finish in normal use. Measurements are stated in terms of the center-line-average figure, in microinches, obtained on traversing a small stylus over a sample of the surface. Comparisons are made by sight and by touch.

621.373.42.001.4:621.317.361 1746

**Testing Precision Oscillators**—M. P. Johnson. (*Wireless World*, vol. 61, pp. 179–182; April, 1955.) The frequency stability of 124-kc oscillators used as masters for carrier-frequency telephone systems is determined by comparison with the 2.5-mc standard-frequency transmissions from Rugby, the measurement being made at a point 10 miles away. Counts of the difference frequency are made over regularly recurring sampling periods, and an output current proportional to the count is obtained. A linear recording meter is used; full-scale deflection is produced by a count of 58, corresponding to a frequency difference of 2 parts in $10^7$ for a sampling period of 116 seconds. The equipment is described and specimen records are shown.

621.373.52:621.314.7 1747

**Transistorized F.M. Signal Generator**—J. J. Hupert and T. Szubski. (*Electronics*, vol. 28, pp. 133–135; February, 1955.) An instrument covering the frequency range 20–100 mc has been designed giving an output of 10 mw across 10Ω. The vhf section comprises a FM oscillator operating at a half or a third of the output frequency, followed by a harmonic-selector stage. The frequency modulation is effected by a transistor acting as variable reactance. The relative merits of point-contact and junction-tetrode transistors for this circuit are discussed. The saving in bulk as compared with equipment using thermionic tubes is to some extent offset by the need for a constant-temperature enclosure for the oscillator.

621.375.2.024.083:681.142 1748

**Gain Measurements on Computing Amplifiers**—A. B. Johnson. (*Electronic Engng.*, vol. 27, pp. 127–129; March, 1955.) Special techniques are required for measurements on amplifiers for dc analog computers, which usually have very high gain. Methods are divided into two broad classes, (a) direct, in which the drift output is reduced without affecting the test signal, the amplifier operating effectively without feedback for the latter, and (b) indirect, in which the drift output is small because of negative feedback, and the gain is deduced from some other property.

621.397.5:535.623].001.4 1749

**Phase Measurement for Color TV and F.M.**—K. Schlesinger. (*Electronics*, vol. 28, pp. 142–146; February, 1955.) The principles of operation with the vectorscope are described.

621.317.3.029.6 1750

**Handbook of Microwave Measurements, Vols. 1 and 2 [Book Review]**—M. Wind and H. Rapaport, Eds. Publishers: Polytechnic Institute of Brooklyn, New York, 616 pp. and 320 pp. (*Wireless Eng.* vol. 32, p. 116; April, 1955.) A useful compendium not only for the technician but also, as a supplement to more theoretical treatments, for the engineer or physicist. The text is all in Volume 1, the diagrams are in Volume 2.

## OTHER APPLICATIONS OF RADIO AND ELECTRONICS

621.3:61 1751

**Electricity in Medicine**—S. N. Pocock. (*Proc. IEE*, Part II, vol. 101, pp. 629–638; December, 1954.) A survey of applications of electronic techniques in diagnosis and therapy.

621.317.39:531.71 1752

**A Direct-Reading Instrument for the Measurement of Small Displacements**—W. D. Corner and G. H. Hunt (*Jour. Sci. Instr.*, vol. 31, pp. 445–447; December, 1954.) Displacements down to $2\times10^{-4}$ cm were measured with an accuracy of within $10^{-7}$ cm by means of a bridge circuit using a differential capacitor, with a 100-v, 10-kc supply. The unbalance was measured by a direct-indicating tube voltmeter. The apparatus was designed for magnetostriction measurements.

621.384.612 1753

**Generating R.F. Energy for 6-kMeV Bevatron**—C. N. Winningstad. (*Electronics*, vol. 28, pp. 164–169; February, 1955.)

621.385.833 1754

**Extension of the Electron-Optical Theory of the Deflecting Electrostatic System to the Case of Relativistic Particles**—A. M. Strashkevich. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2264–2270; December, 1954.)

621.385.833 1755

**Aberrations of Relativistic Electron Beams**—A. M. Strashkevich and N. G. Gluzman. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2271–2284; December, 1954.) A mathematical discussion is presented of the operation of a system with a curved axis. Equations (9) and (10) are derived for a wide beam in an arbitrary electrostatic field for the relativistic case. Equations are also derived for the particular cases of a wide beam in an axially symmetric field (28) and in a plane field (30) as well as in the fields of a cylindrical lens (33) and a cylindrical condenser (35). The aberrations of axially symmetric lenses are calculated for the relativistic case (40)–(45), and also of cylindrical lenses (50)–(53).

621.385.833 1756

**Simple Presentation of the General Theory of Systems of Revolution in Electron Optics (covering Relativity and Aberrations)**—É. Durand. (*Rev. d'Optique*, vol. 33, pp. 617–629; December, 1954.) Calculations are simplified by introducing complex combinations into the Lagrangian.

621.387.424:537.52 1757

**A Cloud-Chamber Study of some Aspects of the Geiger Discharge**—P. J. Campion. (*Proc. Phys. Soc.*, vol. 67, pp. 1095–1102; December 1, 1954.)

621.396:623.451.8 1758

**Launching Control for Guided Missiles**—J. B. Schrock. (*Electronics*, vol. 28, pp. 122–127; February, 1955.) A description of the circuits which control the firing of the missile, actuate the guiding and telemetering equipment, and alert rocket-range control units.

## PROPAGATION OF WAVES

538.566 1759

**Coupled Wave Equations for Inhomogeneous Anisotropic Media**—K. Suchy. (*Z. Naturf.*, vol. 9a, pp. 630–636; July/August, 1954.) "A special system of coordinates has been introduced for the calculation of electromagnetic wave propagation in an inhomogeneous, anisotropic medium. One of the coordinate axes is parallel to the wave normal, the two others (perpendicular to it) are defined by the relation between the E and $\tilde{D}$ vector. In the coupled wave equations it is shown that the coupling terms can be neglected under certain conditions."

538.566 1760

**On the Possibility of Electromagnetic Surface Waves**—P. S. Epstein. [*Proc. Nat. Acad. Sci. (Washington)*, vol. 40, pp. 1158–1165; December, 1954.] An independent surface wave is defined as comprising two inhomogeneous waves which are independent of each other and run along the surface dividing two media, one wave in either medium. A discussion of the conditions for the existence of such a wave indicates that it could exist only at the boundary of two nonconducting media, one with a positive dielectric constant, the other with a negative one. Sommerfeld's solution for the field of an electric dipole at the surface of a plane earth is briefly commented on.

538.566 537.56 1761

**General Expression for the Absorption of Electromagnetic Waves in Lorentz-Type Plasmas (Ionosphere)**—M. Lozzi, R. Jancel and T. Kahan. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 240, pp. 162–164; January 10, 1955.] A general expression is derived from the theory presented previously [2364 of 1954 (Jancel and Kahan)].

538.566.2 1762

**Reflection and Refraction of Electromagnetic Waves in a Stratified Medium**—K. Försterling. (*Hochfrequenztech. u. Elektroakust.*, vol. 63, pp. 112–116; April, 1954.) Exact analysis is presented and the effect of certain approximations is discussed. Splitting due to the presence of magnetic fields is considered.

621.396.11 1763

**Theory of Radio Transmission by Tropospheric Scattering using Very Narrow Beams**—H. G. Booker and J. T. deBettencourt. (PROC. IRE, vol. 43, pp. 281–290; March, 1955.) When propagation is effected by scattering from turbulences in the troposphere [1757 of 1950 (Booker and Gordon)], the energy arriving at the receiver may be expected to extend over a substantial angle; it should be possible to demonstrate the effect by using narrow beams, of width less than about 1.5 degrees. Calculations are made for a communication path of length 300 km between paraboloidal antennas of diameter 100 λ; a beam width of 0.73 degree is assumed. A study is made of the effects to be expected on swinging the two beams in synchronism so that their axes always intersect; a lateral swing of 1 degree off the great-circle path would reduce the received power by about 7 db if propagation is controlled by scattering, whereas it would reduce the received power by about 40 db if propagation is controlled by refraction. The distortion of pulses to be expected as a result of beam swinging is evaluated. Choice of communication bandwidths is discussed.

621.396.11:551.510.535 1764

**Measurement of Attenuation in the Ionosphere**—A. Ochs. (*Arch. elekt. Übertragung*, vol. 8, pp. 535–544; December, 1954.) A fixed-frequency method based on a continuous photographic record of echo amplitude is discussed, and measurements made during the period October, 1952–January, 1953 are reported. Difficulties are introduced by ground reflection, ground irregularity, fluctuations of transmitted power, the effect of the extraordinary component, and nonuniformity of the ionosphere. The last-mentioned factor is especially important, and its effects are illustrated by field-strength records of a signal (a) once reflected and (b) twice reflected from the $F$ layer. The absence of correlation between these records can be explained by assuming that the reflected surface is inclined or curved, and/or that within the usual ionosphere strata there are local regions, or clouds, with higher concentrations of electrons. To eliminate errors due to the focusing effect of the curved surface, measurements must be averaged over a suitable period, must be made at suitably spaced points, and must be made simultaneously at different frequencies. If only a single frequency is used, this should be as low as possible with the available power. Details are given of the equipment used. Carrier frequencies of 2 mc and, later, 1.6 mc were used, with pulse duration 100 $\mu$s, pulse power 15–20 kw, and pulse repetition rate 50 or 1 per second. The results indicate considerable interdiurnal differences of the diurnal variation of attenuation.

**621.396.11:551.510.535 1765**

**Ionospheric Absorption Measurements at Prince Rupert**—K. Davies and E. L. Hagg. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 18–32; January, 1955.) Report and discussion of measurements made near the northern auroral zone between April, 1949 and March, 1950. Monthly median noon values of total absorption, $-\log \rho$, do not fit the inverse square law $-\log \rho \propto (f+f_L)^{-2}$; seasonal variation indicates little dependence on solar zenith angle $\chi$. Diurnal variation is very approximately represented by the relation $-\log \rho \propto (\cos \chi)^{0.5}$, maximum absorption generally occurring about 20 minutes after local noon. There is a pronounced correlation between night-time absorption at 2 mc and the 3-hour-range $K$ index for $K>4$. High night-time absorption is often associated with intense sporadic $E$.

**621.396.11:551.510.535 1766**

**Influence of the Inclination of the Earth's Magnetic Field on the Absorption of Radio Waves in the D Layer**—P. Lejay and D. Lepechinsky. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 136–138; January 10, 1955.] Analysis is presented and its application is illustrated by evaluating the absorption index for three particular directions of propagation. For the extraordinary ray the absorption increases considerably with increase of the angle between the direction of the earth's magnetic field and the direction of propagation; for the ordinary ray the variation of absorption is in the opposite sense. The results indicate that caution is necessary in applying absorption values obtained from vertical soundings to conditions along actual radio communication paths; the cosine law is not directly applicable.

**621.396.11:551.510.535 1767**

**Reflection Conditions for Vertical Propagation in the Ionosphere in the Presence of Collisions and of the Earth's Magnetic Field. Case of the E Layer**—D. Lepechinsky and J. Durand. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 333–336; January 17, 1955.] The type of propagation, i.e. quasi-transverse or quasi-longitudinal, occurring in a given region of the ionosphere is determined by $\theta$, the inclination of the earth's magnetic field, and the collision number. A graph shows the height $h$ at which the transition occurs, as a function of $\theta$; $h$ coincides with the $E$ layer (100 km) for a value of $\theta$ about 65 degrees, i.e. in France at about the latitude of Paris; propagation there is quasi-transverse or quasi-longitudinal according as the collision number increases or decreases. Consideration of the relation between refractive index and electron concentration for the two types of propagation indicates that neglect of collisions introduces error in the calculated values of $f_0E$ for values of $\theta>65$ degrees. These considerations are shown to explain the apparent flattening of the daily variation curve of $f_0E$ around midday.

**621.396.11.029.55 1768**

**The Prediction of Short-Wave Propagation**—B. Beckmann. (*Tech. Hausmitt. Nordw-Dtsch. Rdfunks*, vol. 6, nos. 9/10 and 11/12, pp. 211–219 and 247–259; 1954.) Concepts fundamental to ionospheric and radio weather (3665 of 1954) and the relations between them are discussed. The relevance of solar, geomagnetic and ionospheric factors in forecasting is indicated. An examination is made of the accuracy of forecasts and of the causes of error, e.g. scattering. Statistics are presented showing differences in reception of WWV transmissions at Norddeich and Munich.

**621.396.11.029.6 1769**

**An Experimental Study of the Propagation of 10 cm Radio Waves over a Short Non-optical Sea Path**—E. F. Stack-Forsyth. (*Proc. IEE*, Part B, vol. 102, pp. 231–236; March, 1955.) Tests over a path 1.14 times the optical-horizon distance were made off the coast of Natal during the winter months April–August, 1953. Vertical polarization was used. The effect on signal strength of variations in the structure of the refractive-index profile of the atmosphere in the first few hundred feet above sea level was studied. The results indicate that a duct about 120 feet high was present for a considerable part of the winter. The signal strength at a height of 47 feet in the duct was 6–10 db above the free-space value. The absolute value of the signal strength and its variation with duct height are in moderately good agreement with values given by the mode theory, using only the first mode and assuming either a square-law, fifth-root or bilinear profile.

**621.396.11.029.6:551.510.535 1770**

**Study of Long-Distance Propagation of V.H.F. Waves by Sporadic-E Ionization**—T. Kono, Y. Uesugi, M. Hirai and G. Abe. (*Jour. Radio Res. Labs, Japan*, vol. 1, pp. 1–10; March, 1954.) A report is presented of vertical-incidence measurements and propagation tests carried out in Japan between June and August, 1952, over distances of 500–1,100 km at frequencies of about 31, 43, and 65 mc. Results were analyzed statistically; empirical expressions are given for the relation between $fE_s/f \cos i$ and $F$, where $F$ is the field strength relative to the free-space field strength, $f$ is the transmission frequency, and $i$ is the angle of incidence. The probability of the calculated value of $F$ exceeding its actual value can be estimated. Results are presented graphically.

**621.396.11.029.6:[551.524+551.57 1771**

**What Role does the Nocturnal Cooling play in the Ultra-short-Wave Propagation?**—K. Hirao. (*Jour. Radio Res. Labs, Japan*, vol. 1, pp. 27–39; March, 1954.) Variations of the temperature and humidity in the lower atmosphere at the transmitter are compared with the field-strength variations at the receiving station. The frequency used was 150 mc. Results are tabulated and also presented graphically. Nocturnal cooling has both a direct effect and an indirect effect through the resulting changes of humidity.

**621.396.812.3:551.510.535 1772**

**The Autocorrelogram of Randomly Fading Waves**—R. B. Banerji. (*Jour. Atmos. Terr. Phys.*, vol. 6, pp. 50–56; January, 1955.) The received power spectrum for the case of a completely rough ionosphere having superposed steady and random motion is deduced assuming transmitting and receiving antennas omnidirectional. Autocorrelograms of fading patterns corresponding to pure drift and pure turbulence are compared. An observed fading pattern need not contain more than 250 independent points for these two extreme cases to be distinguished. See also 2120 of 1953.

## RECEPTION

**621.396.621 1773**

**Logarithmic-Amplifier Simplifications and Improvements**—D. E. Sunstein and T. H. Chambers. (PROC. IRE, vol. 43, pp. 343–344; March, 1955.) Comment on 3345 of 1954 (Chambers and Page) and reply.

**621.396.621 1774**

**Statistical Survey of the Engineering Construction of Broadcast Receivers [in Western Germany]**—W. W. Diefenbach. [*Funk-Technik* (*Berlin*), vol. 9, pp. 674–676; December, 1954.]

**621.396.621:621.376.3 1775**

**Reception of Frequency-Modulated Oscillations with Automatic Matching of Receiver Bandwidth to the Dynamic Range of the Modulation**—K. Lamberts. (*Fernmeldetech. Z.*, vol. 7, pp. 665–669; December, 1954.) For weak signals in wide-band noise, the af signal/noise ratio of a FM receiver can be improved by restricting the IF bandwidth. A circuit is described for varying this bandwidth automatically in proportion to the variations of the frequency deviation. An improvement of 14 db is obtained by 1:4 reduction of bandwidth. The system does not eliminate impulsive noise.

**621.396.621.54:621.314.7 1776**

**Transistor Broadcast Receivers**—A. P. Stern and J. A. Raper. [*Elec. Engng.* (*New York*), vol. 73, pp. 1107–1112; December, 1954.] General design principles are discussed and circuit diagrams are shown of some experimental AM superheterodyne receivers using point-contact and junction transistors. Good quality reception is attainable.

**621.396.82:621.376.3 1777**

**Reception of an F.M. Signal in the Presence of a Stronger Signal in the Same Frequency Band, and other Associated Results**—R. M. Wilmotte. (*Proc. IEE*, Part B, vol. 102, pp. 260–261; March, 1955.) Discussion on 1899 of 1954.

**621.396.828 1778**

**Interference Suppression**—R. Davidson. (*Wireless World*, vol. 61, pp. 173–176; April, 1955.) Techniques for dealing with small commutator motors are described. Details are given of recently developed lead-through capacitors.

## STATIONS AND COMMUNICATION SYSTEMS

**621.376.2 1779**

**Tables of Bennett Functions for the Two-Frequency Modulation Product Problem for the Half-Wave Linear Rectifier**—R. L. Sternberg, J. S. Shipman and W. B. Thurston. (*Quart. Jour. Mech. Appl. Math.*, vol. 7, part 4, pp. 505–511; December, 1954.) For previous work see 3028 of 1954 (Sternberg) and 2212 of 1954 (Sternberg and Kaufman).

**621.376.5:621.39 1780**

**Average Spectrum of a Periodic Series of Identical Pulses Randomly Displaced and Distorted**—R. M. Fortet. (*Elec. Commun.*, vol. 31, pp. 283–287; December, 1954.) See 544 of February.

**621.376.56:621.39 1781**

**Signal/Noise Ratio in Pulse Code Modulation**—N. L. Yates-Fish and E. Fitch. (*Proc. IEE*, Part B, vol. 102, pp. 204–210; March, 1955.) Formulas are derived for the output signal/noise ratio in simple pulse-code modulation systems for all values of input signal/noise ratios. The output ratio improves very rapidly with increasing input ratio provided the latter exceeds a certain critical value. The system is useful for links connected in tandem, since degradation of the overall performance below that for a single link may be avoided by a relatively small increase of power in each link.

621.376.56:621.39 1782

**Study of Pulse-Code Modulation**—C. Villars. (*Tech. Mitt. schweiz. Telegr.-Teleph-Verw.*, vol. 32, pp. 449–472; December 1, 1954. In French.) An account is given of an experimental installation developed in collaboration with C. Margna. A binary counter is used as coder, and 32 discrete amplitude levels are recognized on each side of zero. Two types of receiver were constructed, one in which each pulse is treated separately and the other in which pulses are treated in groups, as described by Meacham and Peterson (2366 of 1948). Measurements of signal quality and of the improvement of signal/noise ratio between the hf and af channels are reported and compared with international standard requirements; the af bandwidth is quite satisfactory but quantization noise may be excessive. For multichannel communications the system compares well with others in respect of reliability, freedom from crosstalk, and ease of providing secrecy.

621.391.1 1783

**Prospects for the Development of Transmission Paths**—E. Hölzler. (*Fernmeldetech. Z.*, vol. 7, pp. 647–651; December, 1954.) Review and comparison of various types of line and radio paths. Manufacturing difficulties appear to limit the bandwidth of tubular lines to about 10 mc. The possibilities of surface-wave lines at usw are discussed. Special types of dielectric and metal waveguides may prove suitable at frequencies above that (10–15 kmc) for which atmospheric absorption on radio paths becomes excessive.

621.395.44:621.315.28 1784

**A Transatlantic Telephone Cable**—Kelly, Radley, Gilman and Halsey. (*See* 1547.)

621.396.41:621.376.3 1785

**Linearity Requirements for Multichannel F.M. Radio-Link Systems**—G. Bosse. (*Fernmeldetech. Z.*, vol. 7, pp. 678–682; December, 1954.) The relations between the measured distortion factor and the noise in frequency-division multichannel systems are calculated by substituting an equivalent noise voltage for the sum of the voltages in the channels. Nonlinear distortion due to curvature of modulator and demodulator characteristics and to transmission-time variations is considered. Diagrams are presented from which a determination can be made of the maximum permissible distortion factor for a given signal/noise ratio, and of the optimum amount of pre-emphasis.

621.396.41:621.376.3 1786

**Problems of Frequency Modulation in Multichannel Radio Links**—H. Meinke. (*Fernmeldetech. Z.*, vol. 7, pp. 670–677; December, 1954.) The four principal criteria to be considered in deciding on the type of modulation to use in a multichannel system are (a) signal/noise ratio, (b) crosstalk, (c) bandwidth requirements, (d) equipment requirements. An account is given of experimental work on systems using individual-channel FM. Comparison is made with p.ph.m. systems. The results with the FM system are promising.

621.396.41:621.396.65 1787

**Wide-Band Radio Links: Deliberations of Study Commission No. 9 of the C.C.I.R. at Geneva (from 10th to 22nd September 1954)**—W. Klein. (*Tech. Mitt. Schweiz. Telegr.-Teleph Verw.*, vol. 32, pp. 497–499; December 1, 1954. In French.) Proposals for standardizing multichannel links are reported; both frequency-division FM systems and time-division PPM systems are considered.

621.396.41.029.62:621.396.822.1 1788

**Intermodulation Noise in V.H.F. Multichannel Telephone Systems**—J. L. Slow. (*Jour. Brit. IRE*, vol. 15, pp. 67–83; February, 1955.) Intermodulation noise due to various forms of distortion in the radio circuits of a frequency-division multiplex system is analyzed. Expressions are derived of the form $\overline{N}_n = \overline{H}_n + n\overline{P} + C$, where $\overline{N}_n$ is the $n$th-order noise power and $\overline{P}$ the multichannel speech power in db referred to 1 mw, $\overline{H}_n$ is the $n$th-harmonic ratio of a test tone in db and $C$ is a constant. The analysis is valid for FM systems handling up to 60 channels. Formulas and curves are given and single-tone and two-tone tests are described for determining the intermodulation noise due to (a) modulator/demodulator distortion, (b) phase distortion, (c) feeder mismatch.

621.396.61:621.396.66 1789

**Operational Measurements on U.S.W. [f.m.] Broadcast Transmitters**—L. Merkl. (*Arch. tech. Messen*, no. 227, pp. 269–272; December, 1954, and no. 228, pp. 7–10; January, 1955.) Measurements considered include the monitoring of voltages and tube currents, rf output power, frequency and distortion.

621.396.665.1:621.396.65:621.376.3 1790

**Transmission of Speech with Dynamic Compression**—G. Hässler. (*Fernmeldetech. Z.*, vol. 7, pp. 659–664; December, 1954.) The method of operation of the syllable compandor is described. Design advantages resulting from use of these compandors in multichannel carrier-frequency systems are indicated. The example of a FM radio link is treated numerically.

621.396.712.029.62+621.396.61 1791

**Some Aspects of V.H.F. Sound Broadcasting and F.M. Broadcast Stations**—P. A. T. Bevan. (*Electronic Eng.*, vol. 27, pp. 96–101 and 147–153; March and April, 1955.) The relative merits of systems using AM, AM with limiting, and FM are discussed, mainly on the basis of their effectiveness for suppressing various types of noise and interference; field tests indicated the superiority of FM. A detailed account is given of the Wrotham high-power experiment. The design of FM transmitters is considered, with particular reference to modulators and monitoring. The antenna and transmission-line systems, parallel operation of FM transmitters, and unattended operation of transmitters are also discussed.

621–526 1792

**Closed Expansion of the Convolution Integral (A Generalization of Servomechanism Error Coefficients)**—E. Arthurs and L. H. Martin. (*Jour. Appl. Phys.*, vol. 26, pp. 58–60; January, 1955.)

621.311.6:681.142 1793

**Precision High-Current Computer Power Supplies**—A. B. Rosenstein. [*Trans. Amer. IEE*, Part I, *Communication and Electronics*, pp. 405–409; September, 1954. Digest, *Elec. Engng.* (*New York*), vol. 73, p. 1080; December, 1954.] A unit supplying 225 V dc at 15 a uses a magnetic amplifier-regulated Se rectifier.

621.319.339 1794

**A Portable Van de Graaff Generator**—T. R. Foord. (*Jour. Sci. Instr.*, vol. 31, pp. 440–441; December, 1954.) The generator described develops a maximum open-circuit voltage of about 200 kv and has a short-circuit current of 15 $\mu$a.

621.396.63:621.314.7 1795

**Practical Local Calling Circuit**—(*Short Wave Mag.*, vol. 12, pp. 556–558; December, 1954.) The circuit diagram of an experimental local-station calling device designed for operation on the 160-m amateur band is given and discussed. The unit is basically a transistor receiver which operates a calling bell via a relay. Power consumption is of the order of 0.5 kwh per annum.

## TELEVISION AND PHOTOTELEGRAPHY

621.397.5(44) 1796

**The French Television Network**—(*Télévision*, no. 49, pp. 305–307; December, 1954.) Details are given of revised frequencies for channels 1–12, and basic operational data for the various stations are tabulated.

621.397.5:535.623].001.4 1797

**Phase Measurement for Color TV and F.M.**—K. Schlesinger. (*Electronics*, vol. 28, pp. 142–146; February, 1955.) The principles of operation with the vectorscope are described.

621.397.61:621.372.54 1798

**Filters for Television Transmitter Diplexers**—G. Meyer-Brötz. (*Fernmeldetech. Z.*, vol. 7, pp. 683–688; December, 1954.) The requirements for separating filters used with combined sound and vision antenna systems differ from those for other diplexers because the bandwidth of the vision signal is large compared with the frequency separation of the two carriers. Various types of diplexer are surveyed, and the design of notch diplexers composed of coaxial lines is discussed particularly.

621.397.611:535.623 1799

**Chromacoder Colorcasting**—C. G. Lloyd and P. H. Boucheron. (*Radio-Electronic Engng.*, vol. 23, pp. 7–9, 35; December, 1954.) A description is given of a system employing emitron storage tubes for conversion of the sequential color signal, obtained from a monochrome television camera with a rotating color-segment disk in the lens system, into a NTSC-standard signal. See also 275 of February.

621.397.7+621.397.26 1800

**The Television Transmitter and Relay Installations at Antwerp**—(*Radio Rev. TV*, vol. 6, pp. 610–613; December, 1954.) The station, installed at the top of the 23-story Torengebouw, operates as a two-way microwave link between Brussels and Breda, at the same time broadcasting the received program in band I at a mean power of 2.5 kw. Sound is transmitted by cable. An outline description of the broadcast transmitter is given.

621.397.7:778.5 1801

**Considerations on the Operation of Vidigraphs**—Y. Angel. (*Onde élect.*, vol. 34, pp. 958–973; December, 1954.) The term "vidigraph" is proposed for apparatus for the cinematographic recording of television programs from the face of a receiver tube. A particular system is described using a long-persistence screen. Problems of obtaining correct contrast are discussed.

621.397.7:778.5:621.395.625.3 1802

**The 16-mm Substandard Film with Magnetic Stripe [for sound] as used in the Südwestfunk Television Service—Equipment and Operating Methods**—H. Lauer and O. Schulze. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, nos. 9/10, pp. 203–210; 1954.)

621.397.7.029.62:621.372.51 1803

**V.H.F. Power Transmission Equipment for Band III Television Broadcast**—B. M. Sosin. (*Marconi Rev.*, vol. 17, pp. 88–100; 3rd Quarter, 1954.) A descriptive account including some technical details on the construction and the characteristics of a system for linking television sound and vision transmitters to a common antenna. The system included a vestigial-sideband filter, a frequency-discriminating combining filter, test load and feeder monitoring components.

621.397.8 1804
**Various Factors affecting Picture Quality in Television. Possibilities of Improvement**—F. Below. (*Tech. Hausmitt. NordwDtsch. Rdfunks*, vol. 6, nos. 9/10, pp. 195–202; 1954.) Deleterious effects due to bandwidth limitation and overshoot are considered. Methods of reducing defects due to the vestigial-sideband system of transmission are indicated. Crispening technique described by Goldmark and Hollywood (828 of 1952) and spectrum equalization methods described by Gouriet (1936 of 1953) are discussed. Improvements can be effected by reshaping or replacing the synchronizing pulses and by correct adjustment of level and gamma.

621.397.81:621.397.26 1805
**Propagation on Bands I and III**—F. W. R. Strafford and I. A. Davidson. (*Wireless World*, vol. 61, pp. 171–172; April, 1955.) A direct comparison has been made of propagation in the two bands by radiating 180.4-mc signals from the BBC mast at Sutton Coldfield, as well as the television waveform on 61.75 mc. Two receivers were installed in a mobile unit, and continuous records of signal strength were made. The receiving antennas were at a mean height of 25 feet, thus the difference between the local variations at the two frequencies could be investigated. Rapid variations due to reflecting objects and slow variations possibly due to ground irregularities were observed. The significance of the latter for calculations of service area is discussed. The mean level of the band-III signal decreases with increasing distance faster than that of the band-I signal, as predicted theoretically.

## TRANSMISSION

621.396.61:621.314.7:621.311.6:621.383.5 1806
**SPTTX [sun-powered transistor transmitter] Demonstration for N.P.L.**—(*Short Wave Mag.*, vol. 12, p. 557; December, 1954.) Brief note on a demonstration of the transmitter referred to in 1167 of May.

621.396.61:621.372.2 1807
**Frequency Stability of Self-Excited Transmitters with Long Aerial Feeders**—A. Käch. (*Arch. elekt. Übertragung*, vol. 8, pp. 491–498 and 553–561; November and December, 1954.) A stability criterion is derived whereby the influence of the transmitter is reduced to that of a single equipment constant which can be determined experimentally. The magnitude of this constant and the data of the feeder line uniquely determine the load conditions at the stability limit. For a given type of line the retroaction of the load on the transmitter is greatest for a line length giving a total attenuation of about 3 db. The corresponding critical mismatch at the line termination is shown in normalized load curves. Experimental methods are described for determining the transmitter constant, and an indication is given of the maximum frequency shift to be expected. Agreement between theoretical and practical results is good.

621.396.61:621.376 1808
**Phase-to-Amplitude Modulation**—B. D. Virmani. (*Wireless World*, vol. 61, pp. 183–187; April, 1955.) High efficiency can be obtained with phase-to-amplitude modulation because (a) the phase modulation is performed at low level, and (b) the tubes in the two ph.m. rf channels can be continuously driven to their limits. Details are given of a 400-w transmitter covering the frequency bands 3.5–8, 13–30 and 26—56 mc and permitting either phase-to-amplitude-modulation or ssb operation; the conventional oscillator and phase-shifting network are replaced by a polyphase oscillator, which retains the correct phase displacement when the oscillator frequency is varied.

621.396.61:621.396.712.029.62 1809
**Some Aspects of V.H.F. Sound Broadcasting and F.M. Broadcast Stations**—Bevan. (*See* 1791.)

## TUBES AND THERMIONICS

621.314.63+621.314.7 1810
**Saturation Current in Alloy Junctions**—W. M. Webster. (PROC. IRE, vol. 43, pp. 277–280; March, 1955.) Theory is developed for diodes made by alloying circular junctions on to thin base wafers. An equation is derived from which the value of $I_s$, the saturation current obtained with reverse biasing, can be calculated. Most of this current originates from thermal generation at the free surfaces of the base. $I_s$ increases linearly with base resistivity and exponentially with temperature; it also increases, but more slowly, with base thickness and surface recombination velocity. The equation for the collector of an alloy-junction transistor is basically the same as for a diode, with a correction for the emitter. For the emitter junction the equation requires modification.

621.314.63:546.289 1811
**Inductive Behaviour of *p-n* Junctions in the Forward Direction**—G. Kohn and W. Nonnenmacher. (*Arch. elekt. Übertragung*, vol. 8, pp. 561–564; December, 1954.) The observed time lag of the forward conductance of Ge diodes [3282 of 1952 (Einsele)] cannot be explained in terms of the usual equivalent circuit comprising parallel voltage-dependent resistance and capacitance. A circuit comprising resistance in series with parallel inductance and resistance fits the observations and is also consistent with the small-signal frequency variation of the diode impedance.

621.314.63:546.289 1812
**Admittance Measurements on Alloyed Germanium-Indium Rectifiers**—H. L. Rath. (*Z. Naturf.*, vol. 9a, pp. 699–700; July/August, 1954.) A method is described for determining the admittance from the characteristic variation of the capacitance with frequency at different temperatures.

621.314.632+621.314.7].002.2 1813
**Transistors and Germanium Diodes**—[*Elect. Rev.*, (*London*), vol. 155, pp. 791–795; November, 1954.] An account of production techniques used in Britain for the manufacture of point-contact crystal tubes on a comparatively large scale.

621.314.7 1814
**Transistors and their Applications—a Bibliography, 1948–1953**—A. R. Krull. (*Trans. IRE*, vol. ED-1, pp. 40–77; August, 1954.)

621.314.7 1815
**Electrical Characteristics of Power Transistors**—A. Nussbaum. (PROC. IRE, vol. 43, pp. 315–322; March, 1955.) Measurements have been made on *p-n-p* junction transistors with collector dissipation of about 20 w [3391 of 1954 (Roka et al.]. The results relating to parameters which are a function of current strength do not agree with those obtained theoretically for low-power operation [3390 of 1954 (Rittner)]. Plans for further investigations are indicated.

621.314.7:537.311.33:546.28 1816
**Electronic Behaviour of Certain Grain Boundaries in Perfect Crystals**—Mataré. (*See* 1690.)

621.314.7:621.317.755 1817
**An Alpha Plotter for Point-Contact Transistors**—T. P. Sylvan. [*Elec. Engng.* (*New York*), vol. 73, pp. 1094–1098; December, 1954.] Description, with detailed parts list, of a cro test set.

621.383.27 1818
**The Transit Time of Electrons in Photomultipliers**—E. H. Rhoderick and R. W. P. McWhirter. (*Jour. Sci. Instr.*, vol. 31, p. 475; December, 1954.) Experimental results on commercial 11- and 13-stage photomultipliers indicate transit times ranging from about 5.5 to $8\times10^{-8}$ seconds at voltages between 185 and 105 v per stage.

621.383.5 1819
**Electronic Interpretation of Inertia Phenomena in Photocells (in particular, Internal Capacitance)**—G. Blet. (*Jour. Phys. Radium*, vol. 15, pp. 823–828; December, 1954.) See 2801 of 1954.

621.383.5 1820
**Barrier-Layer Photocells: Part 1**—D. Geist. (*Arch. tech. Messen*, no. 227, pp. 281–284; December, 1954.) A brief account of the mechanism and operational characteristics of *p-n* junction and metal-semiconductor barrier-layer photocells. 35 references.

621.385.029.6 1821
**The Traveling-Wave Tube—a Record of its Early History**—R. L. Wathen. (*Jour. Frank. Inst.*, vol. 258, pp. 429–442; December, 1954.)

621.385.029.6 1822
**A Large-Signal Theory of Traveling-Wave Amplifiers**—P. K. Tien, L. R. Walker and V. M. Wolontis. (PROC. IRE, vol. 43, pp. 260–277; March, 1955.) Analysis presented by Nordsieck (2497 of 1953) is extended to cover space-charge effects at high operating levels. In addition to the space-charge parameter $QC$ and the other parameters used in linear theory, a parameter $k$ is introduced such that $1/k$ is proportional to the beam radius and gives an indication of the range of action of the space-charge forces. Computations have been made for a number of typical cases; amplitude and phase of the circuit wave are given as functions of distance. The limiting efficiency increases with electron injection speed, increases first and then decreases with increase of $QC$, and increases with $1/k$, assuming uniform distribution of field and current over the cross section. The electron motion is analyzed.

621.385.029.6 1823
**The Performance of Travelling-Wave Valves at High Input Levels**—W. Klein and W. Friz. (*Fernmeldetech. Z.*, vol. 7, pp. 349–357; July, 1954.) The helix is considered as a succession of elementary zones, in each of which linear amplification theory is applicable, and the power output is determined in successive steps. Losses are taken into account, and graphs show the dependence of other parameters on the velocity and loss parameters. The effect of an attenuating section on amplification and the optimum choice of its position are examined. The expressions developed are interpreted with reference to a particular tube and the output-power/input-power curve discussed in detail. The importance of the coupling factor $C$ in predicting tube performance is stressed. Amplification at low and high input levels is compared, and the effects of operating voltage and type of helix considered. An estimate is made of the greatest power output compatible with linear operation. Computed and measured values are in reasonable agreement in all cases.

621.385.029.6 1824
**Traveling-Wave-Tube Characteristics for Finite Values of *C***—C. K. Birdsall and G. R. Brewer. (*Trans. IRE*, vol. ED-1, pp. 1–11; August, 1954.) Values of the preformance parameters of traveling-wave tubes are presented in the form of curves for values of the gain parameter $C$ up to 0.5 and for relatively large values, up to 2, of the space charge parameter $QC$.

621.385.029.6 1825
**Cross-Wound Twin Helices for Traveling-Wave Tubes**—M. Chodorow and E. L. Chu. (*Jour. Appl. Phys.*, vol. 26, pp. 33–43; Janu-

ary, 1955.) When a single-helix structure is used with voltages above about 10 kv, a large proportion of the energy is diverted into non-interacting space harmonics, with a corresponding reduction of the impedance for the fundamental interacting mode; undesired backward waves may also be generated. Use of a twin helix eliminates these drawbacks. For a particular twin helix with radius 0.4 $\lambda/2\pi$, the impedance for the fundamental mode was more than twice that of a single helix of the same radius. Dispersion is greater with the twin helix.

621.385.029.6 1826
**Electronic Resonance Effect in Valves with Crossed Electric and Magnetic Fields**—W. Willshaw, G. Mourier and G. Guilbaud. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 283–285; January 17, 1955.] A formula is presented expressing the conditions under which electronic resonance is liable to disturb the normal operation of traveling-wave tubes using lines with periodic structure. A large effect is produced when the resonance corresponds to a space harmonic with high-intensity field. Measurements of the variation of efficiency of a carcinotron over the electronic tuning range are discussed in the light of the theory.

621.385.029.6 1827
**Excitation of the Carcinotron *M* Valve**—G. Mourier. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 406–408; January 24, 1955.] An investigation is made of the current build-up in backward-wave oscillator tubes of the type discussed by Guénard et al. (3616 of 1952); the calculations are based on the energy exchanges between beam and delay line. A coefficient $Q_a$ is introduced defining the excitation quality; an expression is given for $Q_a$ depending on delay-line length, operating wavelength, electron and wave velocities, and ratio of instantaneous current to current in the oscillating state. The excitation time varies in the same sense as $Q_a$.

621.385.029.6 1828
**Influence of Space Charge on the Excitation Current of a Carcinotron-Type Magnetron Oscillator**—B. Epsztein. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 240, pp. 408–410; January 24, 1955.] Values found experimentally for the excitation current in the carcinotron *M* tube are generally lower than values derived from the formula of Guénard et al. (3616 of 1952) and Warnecke et al. (3085 of 1954). A new formula is developed taking account of space charge effects; satisfactory agreement is then obtained between the theoretical and experimental values.

621.385.029.6 1829
**Resonant Behavior of Electron Beams in Periodically Focused Tubes for Transverse Signal Fields**—R. Alder. O. M. Kromhout and P. A. Clavier. (PROC. IRE, vol. 43, pp. 339–341; March, 1955.) Transverse-field traveling-wave tubes are discussed in which a ribbon beam is passed between pairs of plates at alternate high and low potentials; resonance is observed over a range of focusing conditions. A formula is derived for the resonance frequency, which is equal to the plasma frequency.

621.385.029.6 1830
**Magnetic Focusing of Electron Beams**—J. T. Mendel. (PROC. IRE, vol. 43, pp. 327–331; March, 1955.) Analysis is developed applicable to various types of focusing used for pencil beams as in traveling-wave tubes. The semi-shielded cathode offers practical advantages, in conjunction with either uniform or periodic focusing fields. Expressions are derived for the electron trajectories. Balance conditions yielding minimum beam ripple are determined; these require a high percentage of flux threading the cathode for a relatively small increase in magnetic field above the Brillouin value.

621.385.029.6 1831
**Hysteresis in Klystron Oscillators**—T. Moreno and R. L. Jepsen. (PROC. IRE, vol. 43, p. 344; March, 1955.) A possible explanation of electronic hysteresis is advanced based on an electronic-admittance/rf-voltage characteristic which is not monotonic.

621.385.029.6:621.372.413 1832
**Stabilization of Reflex Klystrons by High-*Q* External Cavities**—S. J. Rabinowitz. (*Trans. IRE*, vol. MTT-2, pp. 23–26; September, 1954.) The effective *Q* of the klystron oscillator is raised by associating a high-*Q* external cavity with it. A suitable design of stabilization cavity and coupling network is illustrated, and experimentally determined characteristics of some stabilized klystrons are presented.

621.385.029.6:621.396.822 1833
**Noise in Transverse-Field Traveling-Wave Tubes**—G. Wade, K. Amo and D. A. Watkins. (*Jour. Appl. Phys.*, vol. 25, pp. 1514–1520; December, 1954.) Analysis indicates that in a transverse-field traveling-wave tube the noise can only be kept low if the beam is well collimated. In a particular tube for operation at 1 kmc, with a collimator of width 0.004 inch, the theoretical noise figure is 2 db; without the collimator the noise figure would be 11 db.

621.385.029.63/.64:621.372.2 1834
**Theoretical Study of Traveling-Wave Tube**—K. Udagawa. (*Rep. Elect. Commun. Lab., Japan*, vol. 2, pp. 34–52; August, 1954.) Expressions are derived for the propagation constants of a coaxial arrangement of (a) two helices, and (b) a helix and dielectric cylinder. Series of curves are plotted showing the influence of the thickness of the dielectric cylinder on the phase velocity, and the characteristic impedance and attenuation of a traveling-wave tube helix with a dielectric support. Gain is calculated taking account of thermal effects; space-charge effects and transverse electron motion are investigated.

621.385.029.63/.64:621.396.822 1835
**Influencing Space-Charge Waves of Fluctuating Beams by Resonator Circuits**—K. Pöschl. (*Frequenz*, vol. 8, pp. 284–288; September, 1954.) Theoretical considerations show that the noise factor in a two-stage klystron can be reduced below the minimum given by Robinson (1618 of 1954) by using an additional resonator ahead of the input stage.

621.385.029.65 1836
**An Experimental Broad-Band Helix Traveling-Wave Amplifier for Millimeter Wavelengths**—S. D. Robertson. (*Trans. IRE*, vol. MTT-2, pp. 48–54; September, 1954.) A tube with a helix of diameter 0.015 inch has given 19 db gain at 6 mm $\lambda$ and 9 db at 5.2 mm. Design problems are discussed. An anode potential of 1 kv and a beam current of 10 ma are used.

621.385.029.65:621.317.361 1837
**Cold Measurements of 8 mm Magnetron Frequency and Pulling Figure**—A. E. Barrington. (*Proc. IEE*, Part B, vol. 102, pp. 247–248; March, 1955.) A square-wave signal is applied to the magnetron, and voltages proportional respectively to the input and reflected signals are applied to the *x* and *y* plates of a cro, giving a straight-line trace whose slope is minimum at resonance. The change of resonance frequency with adjustment of the position of a puller probe is observed.

621.385.032.21 1838
**Influence of Space Charge in Spherical Electron Guns**—N. B. Aizenberg. (*Zh. tekh. Fiz.*, vol. 24, pp. 2079–2082; November, 1954.) While it is usually assumed that the intensity of the field at the tip of the cathode in a spherical electron gun is proportional to the anode voltage, for large values of the discharge current the space charge must be taken into account. The effect of this charge is investigated experimentally and the minimum value of the discharge current for which the effect becomes noticeable is determined.

621.385.032.21 1839
**New Forms of Thermionic Cathode**—[*Nature*, (*London*), vol. 174, pp. 1176–1177; December 25, 1954.] Report of a colloquium held in October, 1954 and sponsored by the Institute of Physics and the Physical Society. The discussion dealt mainly with efforts to produce cathodes giving high emission over long periods.

621.385.032.21:537.29 1840
**Progress in Electron Emission at High Fields**—W. P. Dyke. (PROC. IRE, vol. 43, pp. 162–167; February, 1955.) The properties of cold and hot field-emission cathodes yielding high current densities are surveyed and methods of stabilizing their performance are discussed.

621.385.032.216 1841
**A New Type of Diffusion Cathode**—A. H. Beck, A. B. Cutting, A. D. Brisbane and G. King. [*Nature* (*London*), vol. 174, pp. 1010–1011; November 27, 1954.] Brief details are given of a cathode made by molding a mixture of Ni powder and (Ba, Sr) carbonate powder with a small percentage of reducing agent. The pulsed emission is shown as a function of temperature in comparison with the conventional oxide cathode and with published data for the L cathode. From determinations of the work function it is inferred that the emission originates from an incomplete monolayer of Ba ions with about 70 per cent surface coverage. For a fuller account, see *Le Vide*, vol. 9, pp. 302–309; November, 1954.

621.385.032.216 1842
**Contaminated-Metal Sintered Thermionic Cathodes**—R. Uzan and G. Mesnard. [*Compt. Rend. Acad. Sci.*, (*Paris*), vol. 239, pp. 1613–1615; December 8, 1954.] New techniques are outlined for preparing high-emission molded cathodes combining metal and oxide powders in such a way that the surface is entirely of metal. Good results are obtained with Ni when the oxide content is only just sufficient to provide the necessary diffusion to the surface; the sintering may be performed in vacuum or in a hydrogen atmosphere. W gives lower emission than Ni at low temperatures but greater emission at temperatures over 1,250 degrees K; sintering in hydrogen lowers the emission with W. See also 3728 of 1955.

## MISCELLANEOUS

621.396.6+621.396.712 1843
**Development Work of [West] German Broadcasting Institutes**—H. Rindfleisch. (*Elektrotech. Z., Edn A*, vol. 75, pp. 587–590; September 11, 1954.) A brief survey of equipment and techniques developed since the war, and particularly since 1951. 44 references.

621.396.97 1844
**Recording and Tabulating the Radio-TV Audience**—A. C. L. Brown. (*Electronics*, vol. 28, pp. 126–129; January, 1955.) Listener research is conducted by means of film records indicating the periods during which the receiver is tuned to different channels. The recording system is applicable for other purposes, e.g. investigation of atmospheric variations.

413=30=20 1845
**English-German Technical and Engineering Dictionary [Book Review]**—L. de Vries. Publishers: McGraw-Hill, London, 997 pp., £7. (*Wireless Eng.*, vol. 32, p. 90; March, 1955.) A companion volume to the already published German-English Dictionary. Includes many new terms in the field of radar, television, nuclear engineering, etc.

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(Continued from page 78A)

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(Continued from page 83A)

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(Continued from page 84A)

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(Continued from page 86A)

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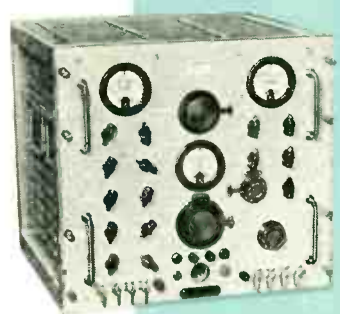

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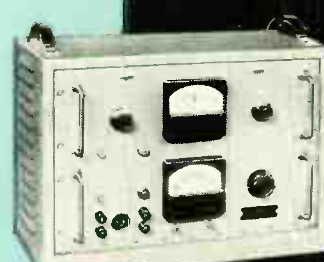
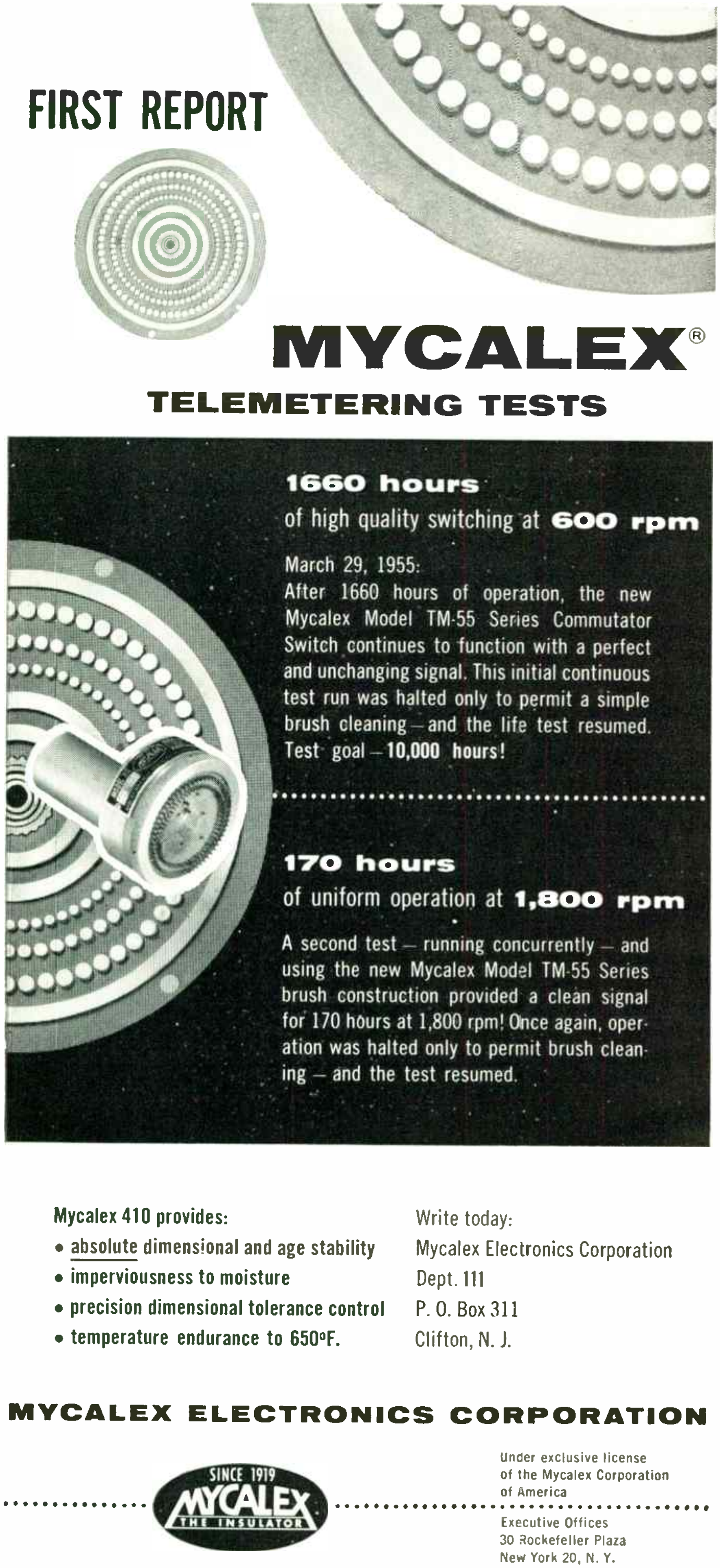


**BJ Incidental FM Meter**

Measures incidental deviation of a carrier in the presence of AM.




## Membership

(Continued from page 88A)

Stuart, H. D., 1844 Ardmore Blvd., Pittsburgh 21, Pa.
Summers, R. M., Springer Route, Ardmore, Okla.
Swan, C. B., 324 Bridge St., E., Belleville, Ont., Canada
Tamburo, S., 1032 West Grand Ave., Chicago 22, Ill.
Thurston, R. N., Bell Telephone Laboratories 1D 446, Murray Hill, N J
Troutman, P. G., 1836 Melissa, Arlington, Texas
Varn, S. P., Jr., 1545 Bolling Ave., Norfolk, Va.
Vlasak, W. R., 4417 North 21 St., Phoenix, Ariz.
Voelcker, H. B., Jr., SIGEL-CRB-3 Coles Signal Laboratory, Fort Monmouth, N. J.
Walsh, R. M., 1148 Amador Ave., Berkeley 6, Calif.
Weinberg, J., 116 Lake Ave., Boonton, N. J.
Wexler, P. S., Raytheon Manufacturing Company, 55 Chapel St., Newton 58, Mass.
Whitten, A. C., R.F.D., 2 Seneca Falls, N. Y.
Wible, P. E., Jr., 1823 Griswold Dr., Apt. L-21, Fort Wayne 3 Ind.
Williamson, W. P., III, Box 8692 Wright Air Development Center, Wright Patterson Air Force Base, Ohio
Zellner, F. L., Jr., 1143 W. 11 St., McKees Rocks, Pa.

## Industrial Engineering Notes

### AERONAUTICAL*

Chairman Louis B. Rothschild of the Air Coordinating Committee has announced that the group has unanimously approved a program to be followed in connection with the current controversy surrounding the common system of short-distance radio navigational aids. "In view of the military's planned implementation of a tactical system," the ACC announcement said, "the divergence between the common civil/military non-tactical system and the tactical military system becomes obvious. Thus, a course of action must be devised which will minimize disruption to all aviation interests during this divergence, and arrive, to the extent possible, at a common civil/military system of navigation which provides for civil/military non-tactical, as well as basic tactical operations." The controversy was touched off earlier this year when the Air Navigation Development Board issued a report in which it favored the TACAN system developed by the military over the VOR/DME system recommended by the Radio Technical Commission for Aeronautics and accepted as the common system. The matter has been probed by several committees on Capitol Hill and a resolution is now pending to establish a joint committee to investigate the matter. The ACC program was reported in part as follows: A. Interim Military Tactical Program (1) The military will proceed immediately to implement the minimum amount of TACAN necessary to meet military tactical require-

(Continued on page 92A)

* The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of April 18, 25, May 2, and 9, published by the Radio-Electronics Television Manufacturers Association, whose helpfulness is gratefully acknowledged.


## "Circuit-designed and circuit-tested" to meet today's TV requirements

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For complete information on the New Sylvania types 25DN6 and 6CS7, check the appropriate space. Or if you have interests in other equipment fields let us send you a complete listing of Sylvania "circuit-designed and circuit-tested" tubes.

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**"Circuit-designed"** for horizontal amplifier use in *off-the-line* low B+ series-string TV applications; and to eliminate "snivet" problems.

**"Circuit-tested"**—to exhibit a low plate knee characteristic and deliver high peak currents which are necessary for proper deflection.

### Type 6CS7

**"Circuit-designed"** to provide an oscillator combined with a vertical deflection amplifier with higher plate dissipation (6.5 watts).

**"Circuit-tested"** — to deliver optimum performance at higher ratings under the more stringent operating conditions of modern circuitry.

Sylvania Electric Products Inc., 1740 Broadway, New York 19, N. Y.

☐ Please send complete data on the new 25DN6 and 6CS7

Please send information on other "circuit-designed and circuit-tested" types as indicated below.

☐ Other entertainment types ☐ Control equipment types
☐ Military equipment types ☐ Test equipment types
☐ Special-purpose types ☐ ______________types

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*(Continued from page 90A)*

ments. These installations must avoid interference with the Radar Safety Beacon and DME. (2) The proposed use of TACAN must be coordinated through the ACC agencies, and aircraft flights based on the system must be capable of being integrated into the operation of the Federal Airways to minimize operational conflicts. Implementation of TACAN will be in accordance with policy to be developed by ACC. B. The Common System Program (1) The CAA continue the VOR/DME system, as hereinafter provided, until some succeeding common system has been adopted and installed on the Federal Airways system. During the transition period when a succeeding common system is being installed, priorities in frequencies or other areas of conflict shall be given to the succeeding system. (2) The agencies of the government responsible for the implementation or operation of any phase or phases of the present or a future common system program will consult with other interested agencies through the ACC in the discharge of these responsibilities. Any implementation of DME will be carried out only in proper priority relationship to the other needed improvements in the system. (3) Based on information now available, if TACAN is adopted for use in the succeeding common system, it is estimated that under such a program, VOR will be continued until 1965 and that DME will be continued until 1960. The ANDB also was directed to study several "undetermined factors" in connection with the use of TACAN and "immediately plan and direct a program to complete the development of TACAN for possible common system use." It also was directed to "conduct studies to determine the feasibility of developing a third rho-theta system which would meet all of the stated common system requirements." The ACC statement also covered the international program to be followed.

### FCC Actions

**The Federal Communications Commission has issued a public notice soliciting information which can serve as the basis for determining whether a rule-making procedure should be instituted concerning radio-astronomy frequency requirements.** The Commission stated that in view of the widespread interest and work being done in the field of radio-astronomy, both in this country and many others, it considers it expedient to develop at this time additional information and has listed some six points upon which comments are solicited from interested parties by July 1.

### Electronics

**A high level of procurement for electronic and communications equipment will be sustained by the Air Force for the next four or five years, according to present**

*(Continued on page 94A)*

## Industrial Engineering Notes

(Continued from page 92A)

plans which look toward the expansion of America's early warning system and modernization and expansion of the aircraft program. Air Force obligations during fiscal year 1956 for the procurement of electronics and communications equipment are expected to total nearly $625.9 million. This compares with obligations of about $427.6 million in 1955 and $326.4 million obligated for this type of equipment in fiscal year 1954.

### MOBILIZATION

Atomic Energy Commission Chairman Lewis Strauss has announced a new program under which organizations or individuals may be given access to non-military "confidential" and "secret" restricted data on atomic energy technology for their own private purposes. Under the new program, information classified as "confidential, restricted data" may be made available to any person who can evidence a potential use or application of the data in his business, profession or trade. The other conditions to this access of information are that the applicant obtain a simplified security "L" clearance and agree in writing to conform to all AEC security regulations. Also, so-called "study agreements" now in effect will be converted to the new type arrangements. Under similar conditions, limited access also may be granted to certain specific information classified as "secret, restricted data" if the applicant proves that the information is significantly important to his business, provided, however, that the applicant obtains a full security "Q" clearance. The government will retain royalty-free, non-exclusive rights for governmental purposes in inventions and discoveries which result from such access. Present AEC contractors will be granted access to both categories of restricted data for private purposes on the same basis and conditions and will be granted the same rights as other applicants.

### RETMA

Climaxing a three-day industry conference, the RETMA Board of Directors approved proposals for broad administrative changes in the Association's organizational structure, subject to approval by the membership at the June convention, and selected Director Leslie F. Muter, pioneer radio manufacturer and veteran RETMA Treasurer and Past President, to receive the 1955 Medal of Honor at the industry banquet on June 16 in Chicago.

### TECHNICAL

The Office of Technical Services, Commerce Department, recently released several reports dealing with scientific discoveries and developments of interest to the

(Continued on page 97A)

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## Industrial Engineering Notes

(Continued from page 97A)

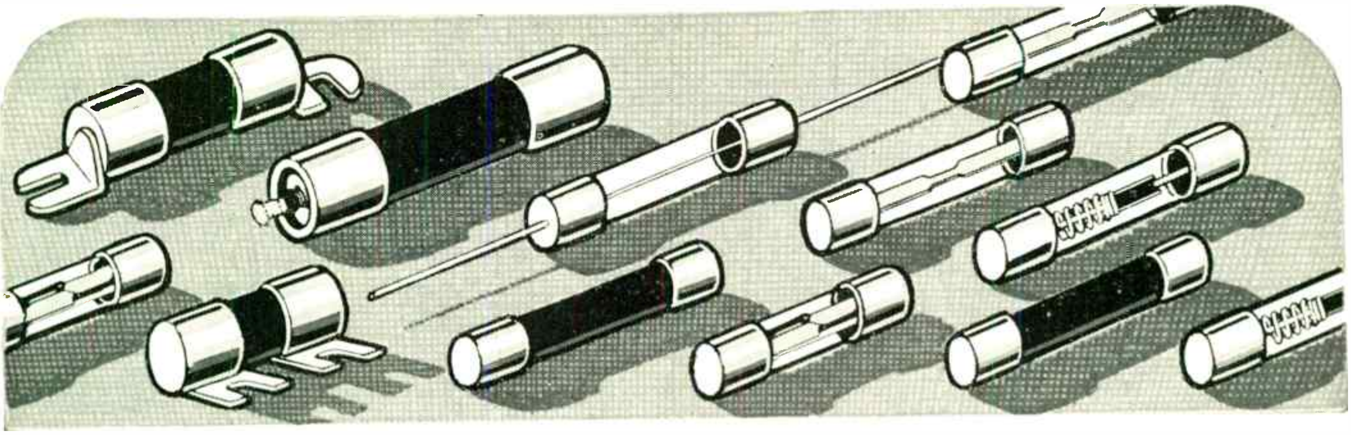
electronics industry, and in addition made available to the public the "AN" (Army-Navy) nomenclature system for communications and electronics equipment, devised by the Joint Communications-Electronics Committee of the Defense Department. "Properties of Large Slot Antennas" PB 111523, and "A High Vacuum Gold Cathode Gaseous Discharge—A Cyclotron Effect" PB 111522, are both available from the OTS, Commerce Department, Washington 25, D. C., for $1.25 each. "The Summary of Joint Nomenclature System ("AN" System) for Communications-Electronic Equipment" PB 111581, also is available from the Commerce Department, for 25 cents per copy. This report presents a coordinated system of nomenclature for communications-electronic equipment, jointly used by the three military services, in a handy chart form. The system is useful, it was said, in identifying equipment references where reference is made to a new item, or a complete listing is not at hand. The coding system used in the charts indicates at once the type of equipment, where it is normally installed, and its functional purpose. . . . **Technical details on the system for mass production of electronics—known as the "Modular Design of Electronics" and the "Mechanized Production of Electronics"—have been made available to industry in five reports published recently by the Office of Technical Services, Commerce Department.** The reports include a summary description of the system, techniques of conversion from conventional to modular design, hand fabrication techniques, mechanized production, and manufacturing cost determination. The system of producing electronics equipment mechanically is not limited to large-scale production, it was said. It may be applied equally well to model-shop or laboratory practices. It is suggested, in fact, that hand assembled electronic models be developed and performance tested before large-scale production by MPE techniques is attempted. Through the application of this system it is possible to substantially reduce the lead time normally required before full production is attained and to stock-pile production facilities, it was noted. It requires no proportionate increase in skilled manpower when the standby plant is put into operation, OTS said. The five volumes describing this system are listed below. They may be obtained from the OTS, Commerce Department, Washington 25, D. C., at the prices indicated. "Vol. 1, Summary of Modular Design of Electronics and Mechanized Production of Electronics," PB 111275, $2. "Vol. 2, Techniques for Converting from Conventional Design of Electronics to Modular Design of Electronics," PB 111276, $2. "Vol. 3, Hand Fabrication Technique and Photographic Processing for Modular Design of Electronics," PB 111277, $2. "Vol. 4, Mech-

(Continued on page 98A)

## Industrial Engineering Notes

*(Continued from page 97A)*

anized Production of Electronics," PB 111278, $4. "Vol. 5, Manufacturing Cost Determination," PB 111315, $4. . . . **The Office of Technical Services, Commerce Department, has listed studies in the field of electronics in its March 1955 issue of the "U. S. Government Research Reports."** The following reports can be purchased from the Photoduplication Section, Library of Congress, Washington 25, D. C., for the reported price: "Design and Calibration of Microwave Antenna Gain Standards" PB 116133, microfilm, $2.50; photocopy, $5.25. "Comparison of Linear and Circular Polarization at X-band by Means of a CW Doppler Radar Operating Over Water" PB 116186, microfilm, $1.50; photocopy, $1.50. "Aerodynamic and Radar Transmissivity Properties of Screen Materials" PB 116249, microfilm, $3.25; photocopy, $9. "Folded Antennas" PB 116294, microfilm, $4; photocopy, $11.50. "General Study of Rectangular Waveguide (pressurized)" PB 116171, microfilm, $2.50; photocopy, $5.25. "Noise Studies on CW Klystrons" PB 116250: microfilm, $1.50; photocopy, $1.50. "On the Theory of Wave Propagation in Non-Homogeneous Media" PB 116185, microfilm, $2.25; photocopy, $4. "On the Perturbation Theory of Electromagnetic Cavity Resonators" PB 115744, microfilm, $2; photocopy, $2.75. "Practical Transmission Line Network Design for VHF and UHF Filter Applications" PB 115942, microfilm, $5.25; photocopy, $16.50. "Microwave Noise Study" PB 116251, microfilm, $3.75; photocopy, $10.25. "Study of R-F Performance Measurements—Final Report" PB 116073, microfilm, $5.25; photocopy, $16.50. "Synthesis. Final Report" PB 116102, microfilm, $2.25; photocopy. $4. "Trouble-Shooting in Electronics Equipment—A Proposed Method" PB 116207, microfilm, $5; photocopy, $15.25. "Use of Real Gases in a Shock Tube" PB 116211, microfilm, $2.75; photocopy, $7 75. . . . **The Air Force has recently released the results of a program of basic research in the field of nonlinear servomechanisms.** The results are described in a research report made available to industry by the Office of Technical Services, Commerce Department. The report, "Research in Nonlinear Mechanics as Applied to Servomechanisms," is the result of research done by the Cook Electric Co. Laboratories under Wright Air Development Center contract. The work was aimed at an improvement of servomechanism response through the use of nonlinear techniques, development of practical nonlinear elements to instrument these techniques, extension of nonlinear servomechanism theory through theoretical analysis, and formulation of nonlinear servo theory into practical design information for use by design engineers. The report is available from the OTS, Commerce Department, Washington 25, D. C., and should be ordered by number—PB 111584—priced at $3.75.



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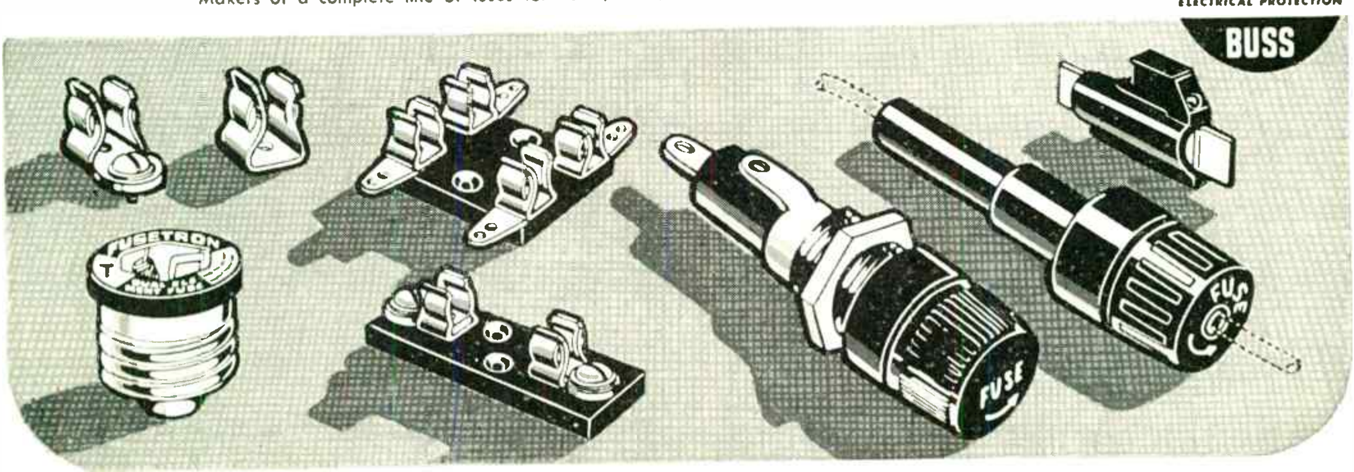

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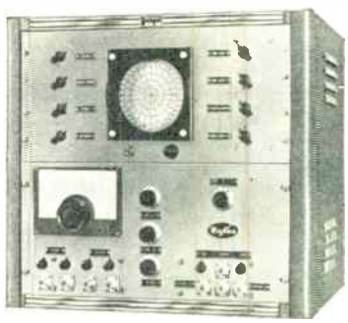

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## Professional Group Meetings

### Antennas and Propagation

Washington Chapter—March 28

"Ionospheric Propagation" by Alvin McNish, National Bureau of Standards.

### Audio

Boston Chapter—April 14

"The Past, Present and Future of Magnetic Recording" by John S. Boyers, Magnetic Memory Devices Division, National Company.

Houston Chapter—March 22

"A Multi-Loop Self Balancing Power Amplifier" by J. Ross MacDonald, Texas Instruments, Inc.

Philadelphia Chapter—March 17

"Experiences and Observations Along the Road to Improved Sound Reproduction" by E. W. Kellog, RCA (retired).

Chicago Chapter—February 18

"Toward Flutter Measurements of Magnetic Recorders" by U. R. Furst, Furst Electronics.

### Circuit Theory

Albuquerque Chapter—March 23

"Circuit Synthesis with Particular Reference to Transistors" by Walter Brown, Sandia Corporation.

Chicago Chapter—January 21

"RF Spectra as Related to Non-Linear Circuit Elements" by William Firestone, Motorola, Incorporated.

Seattle Chapter—February 10

"Digital Computers and Automatic Control" by D. R. Brown, M.I.T.

Syracuse Chapter—March 15

"Active Filters" by J. J. Suran, General Electric.

### Communications Systems

Washington Chapter—March 30

"Disaster Planning in the Field of Telecommunications" by Horace R. Hampton, C and P Telephone Company.

### Component Parts

Los Angeles Chapter—March 14

"The Metal Film Potentiometer" by D. W. Moore, Servomechanisms, Inc.

"Instrument Switches" by C. Broneer and G. Edwards, both of Aerovox Corporation.

(Continued on page 104A)

# 23 Fields of Special Interest -

The 23 Professional Groups are listed below, together with a brief definition of each, the name of

## ACTIVITIES

The IRE Professional Group has the responsibility of providing the individual with the advantages of a small, select society in the field of his specialization, with its own magazine, just as IRE provides him with the advantages of a large, general society. The advantages of the small society relate primarily to meetings and to publications. Specialized symposia may be arranged either to coincide with IRE Conventions or to occur where there are places of large activity in the field of interest.

The Group is concerned with the advancement of scientific engineering leading to increased professional standing in its field and serves to aid in promoting close cooperation and exchange of technical information among its members. It provides a forum for discussion and presentation of papers on subjects of mutual interest, and provides smaller, more compact Groups who may meet on the common basis of professional interests.

## ORGANIZATION

The IRE Professional Group is established under a constitution within the framework of the IRE. The constitution defines the technical field of interest of the Group, establishes committee structures, describes broadly its functions and procedures, and fixes a minimum level of activity. The management of an IRE Professional Group is in the hands of its Administrative Committee, the officers and members of which are elected annually. The IRE provides financial assistance to the Groups in accordance with their activity and current needs.

## PUBLICATIONS

Every Group publishes a magazine which is called TRANSACTIONS of the Professional Group, generally on a regular quarterly schedule. The TRANSACTIONS serve to preserve and disseminate the body of knowledge that constitutes the fields of interest of the Groups. All editions are distributed without additional cost to members who have paid the annual assessment.

The CONVENTION RECORD covering the sessions presented at the IRE National Convention is furnished without further charge to the members of Groups who have paid assessments.

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IRE members of any grade are eligible for membership in the IRE Professional Groups and will receive all Group publications upon payment of the prescribed assessments. An IRE member may join as many Professional Groups as serve his interests and wishes.

To join IRE Professional Groups, indicate on the application coupon in the lower right-hand corner of the opposite page the Group or Groups you wish to join. Detach completed coupon and mail with your check for assessments to The Institute of Radio Engineers, 1 East 79th Street, New York 21, N.Y.

### Aeronautical and Navigational Electronics

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

Mr. Edgar A. Post, Chairman, United Air Lines, Operations Base, Stapleton Field, Denver 7, Colo.

Fee $2. 13 Transactions, 4 Newsletters, *5, *6, *8, & *9, and *Vol. ANE-1, Nos. 1, 2, 3 and 4.

### Antennas and Propagation

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

Mr. Delmer C. Ports, Chairman, Jansky & Bailey, 1339 Wisconsin Ave., N.W., Washington 7, D.C.

Fee $4. 11 Transactions, 1 Newsletter, *4, *Vol. AP-1, Nos. 1, 2; *Vol. AP-2, Nos. 1-4; AP-3, No. 1-2.

### Audio

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

Mr. Winston E. Kock, Chairman, Bell Telephone Laboratories, Inc., Murray Hill, N.J.

Fee $2. 24 Transactions, 4 Newsletters. *5, *7, *10. *Vol. AU-1, Nos. 1-6; *Vol. AU-2, Nos. 1-5; Vol. AU-3, Nos. 1-2.

### Automatic Control

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

Mr. Robert B. Wilcox, Chairman, Raytheon Mfg. Co., 148 California St., Newton 58, Mass.

Fee $2.

### Broadcast & Television Receivers

*The design and manufacture of broadcast and television receivers and components and activities related thereto.*

Mr. Wilson P. Boothroyd, Chairman Philco Corporation, Tioga and "C" Sts. Philadelphia 32, Pa.

Fee $2. 9 Transactions. *1, *2, *3, *5, *6, *7, 8; BTR-1, No. 1.

### Broadcast Transmission Systems

*Broadcast transmission systems engineering, including the design and utilization of broadcast equipment.*

Mr. Oscar W. B. Reed, Jr., Chairman, Jansky & Bailey, 1735 DeSales St., N.W., Washington, D.C.

Fee $2. 1 Transaction, No. 1.

### Circuit Theory

*Design and theory of operation of circuits for use in radio and electronic equipment.*

Dr. Herbert J. Carlin, Chairman, Microwave Research Institute, Polytechnic Institute of Brooklyn, 55 Johnson St., Brooklyn 1, N.Y.

Fee $2. 7 Transactions. *1, *2, *Vol. CT-1, Nos. 1-4; CT-2, No. 1.

### Communications Systems

*Radio and wire telephone, telegraph and facsimile in marine, aeronautical, radio-relay, coaxial cable and fixed station services.*

Mr. Arthur C. Peterson, Jr., Chairman, Bell Telephone Laboratories, 463 West Street, New York 14, N.Y.

Fee $2. 4 Transactions. 5 Newsletters. *Vol. CS-1, No. 1; *Vol. CS-2, Nos. 1-2; CS-3, No. 1.

### Component Parts

*The characteristics, limitation, applications, development, performance and reliability of component parts.*

Mr. Floyd A. Paul, Chairman, Bendix Development Lab., 116 W. Olive Ave., Burbank, Calif.

Fee $2. 3 Transactions. *PGCP-1-2-3.

### Electron Devices

*Electron devices, including particularly electron tubes and solid state devices.*

Dr. John S. Saby, Chairman, Electronics Laboratory, General Electric Company, Syracuse, New York

Fee $2. 9 Transactions, 3 Newsletters, 2 Technical Bulletins. *1, *2, *4, *Vol. ED-1, Nos. 1-4; ED-2, No. 1.

THE INSTITUTE OF RADIO

## Professional Group Meetings

*(Continued from page 101A)*

Los Angeles Chapter—January 10

"Manufacturing and Application Techniques of Transistors" by Don Combes and Leslie King, Hydro-Aire, Inc.

Philadelphia Chapter—April 20

"Significant Testing of Super-Reliable Components" by John A. Connor and Richard H. Baker, Radio Corporation of America.

Washington Chapter—March 16

"Behavior of Ferrites in Microwave Components" by John C. Cacheris, Diamond Ordnance Fuze Laboratory.

### Electron Devices

Boston Chapter—March 30

"Low-Noise UHF Ceramic-Metal Triode" by G. C. Downing and W. C. Wicke, both of Bomac Laboratories, Inc.

### Engineering Management

Philadelphia Chapter—October 27

"Interaction Between Top-Level Management and Engineers in a Large Corporation" by J. T. Cimorelli, RCA

Chicago Chapter—February 18

"Management Considerations for New Product Introduction" by E. H. Wavering, Motorola, Inc.

New York Chapter—April 21

"Supervising Engineering Programs from the Cost Point of View" by F. X. Lamb, Weston Electric Instrument Corporation.

### Electronic Computers

Akron Chapter—April 26

"Linear Programming" by Joseph E. Flanagan, Applied Science Rep. of IBM Corporation.

Boston Chapter—April 21

"Digital Machines for Nationwide Dialing" by John Meszar, Bell Telephone Laboratories.

Boston Chapter—February 24

"Panel Discussion: Requirements and Applications of Computers in Business" by Milton Brand, Nowland and Company, and Edward L. Wallace, Harvard Business School.

Chicago Chapter—February 18

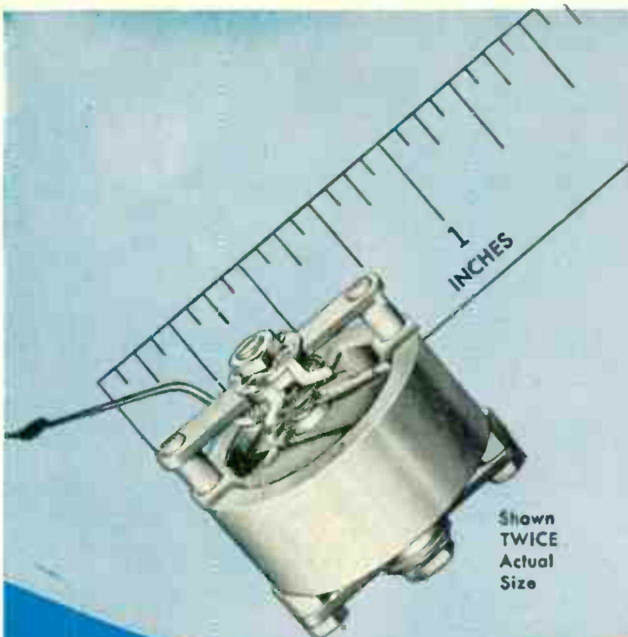
"Fundamental Considerations in the Design of Magnetic Core Storage Systems" by Robert Schuman, Argonne National Laboratories.

Chicago Chapter—January 21

"Teletype High-Speed Equipment and Systems" by W. P. Byrnes, Teletype Corporation.

*(Continued on page 151A)*

# New Simpson Core Meter Movement

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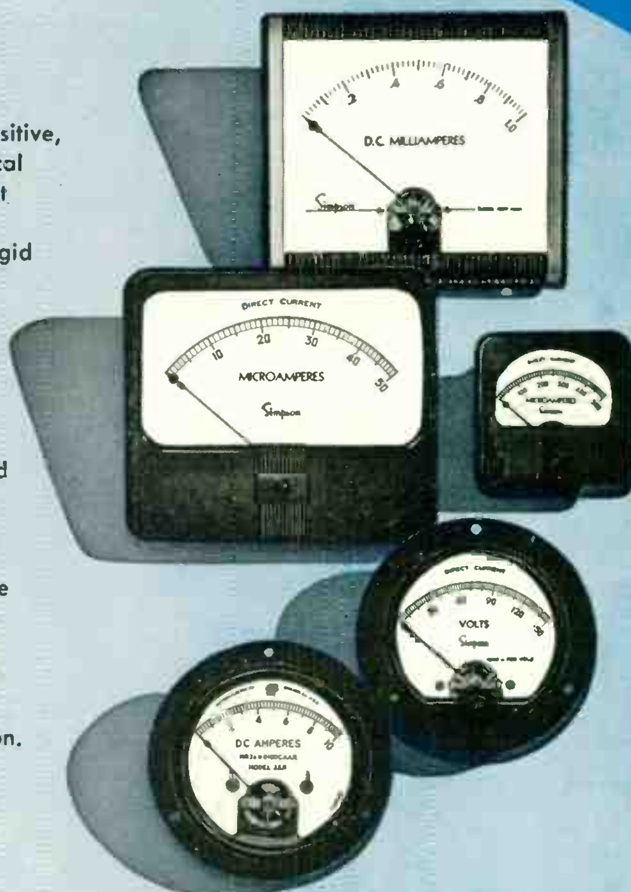
Here's the new Simpson ***Core-Type Meter Movement.*** It's a more compact, more sensitive, ***self-shielding*** movement that gives electrical measurements with laboratory accuracy, yet has the ruggedness to withstand severe shocks. Its accuracy specifications are so rigid that Simpson engineers had to devise unusual production techniques.

Let Simpson engineers design panel meters using the new core movement to your special instrument requirements. Simpson continues to maintain its large stock of standard panel meters in over 700 sizes and ranges, available through distributors.




## RUGGEDIZED METERS

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## Section Meetings

### Akron

"The Weapon Systems Concept," by Col. E. N. Ljunggren, Air Research and Development Command; April 19, 1955.

General Electric Lighting Show, and Election of Officers; May 17, 1955.

### Albuquerque-Los Alamos

"Establishment of Reliabilities for Vacuum Tubes in Complex Electronic Devices," by R. O. Frantik, and "Quality Assurance of Electron Tubes for Maximum Reliability Application," by A. F. Hurford, both of Sandia Corporation; May 12, 1955.

### Atlanta

"Frequency Control of UHF Oscillators," by E. D. Holmes, Georgia Institute of Tech.; April 29, 1955.

### Beaumont-Port Arthur

"Antenna Applications in Two Way Radio Systems," by T. J. McMullin, Communications Engineering Company; May 18, 1955.

### Binghamton

"Development of Automation," by T. W. Zebley, General Electric Corp.; May 9, 1955.

### Buffalo-Niagara

"The Evolution of Broad-Band Mixer-Duplexers," by T. N. Anderson, Airtron, Inc.; April 20, 1955.

### Cincinnati

"Engineering Problems in the Nuclear Age" by Rear Adm. H. G. Rickover, U. S. Navy; February 24, 1955.

"The New WLW Cathanode Modulation System," by R. J. Rockwell, Crosley Bcstg. Corp.; March 15, 1955.

Spring Technical Conference; April 15, 16, 1955.

### Cleveland

"The Weapon System Concept," by Col. E. N. Ljunggren, USAF; April 19, 1955.

### Dayton

"Semiconductor Physics as Applied to Junction Transistors and Rectifiers," by N. B. Nichols, Raytheon Mfg. Co.; April 26, 1955.

### Detroit

"Automatic Fabrication of Electronic Equipment," by D. F. Melton, General Mills, Inc.; April 15, 1955.

### Elmira-Corning

"Automatic Street Lighting Control," by G.W. Nagel, Westinghouse Electric Corp.; April 18, 1955.

### El Paso

"Engineering Training," by Dr. B. W. Holcombe, Texas Western College; February 21, 1955.

"Audio Engineering," by P. W. Klipsch, Klipsch and Associates; March 8, 1955.

"Intercommunication Systems, Aural and Visual," by H. Markowitz, Custom Electronics; April 21, 1955.

### Emporium

"Traveling Wave Tubes," by John Gaenzle, Sylvania Electric Products; May 17, 1955.

### Fort Wayne

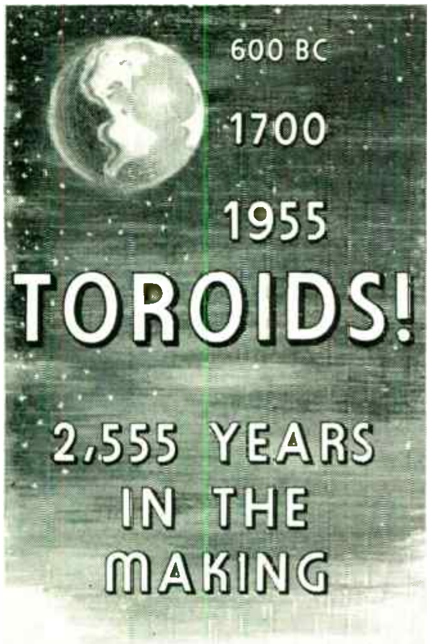
"Basic Concepts of Information Theory," by Dr. B. M. Oliver, Hewlett-Packard Corp.; March 31, 1955.

"Status of Traveling-Wave Tube Development," by A. K. Wing, F.T.L.; April 7, 1955.

### Houston

"Instrument Engineering," by J. V. Sigford, Minneapolis Honeywell Regulator Co.; May 17, 1955.

*(Continued on page 108A)*

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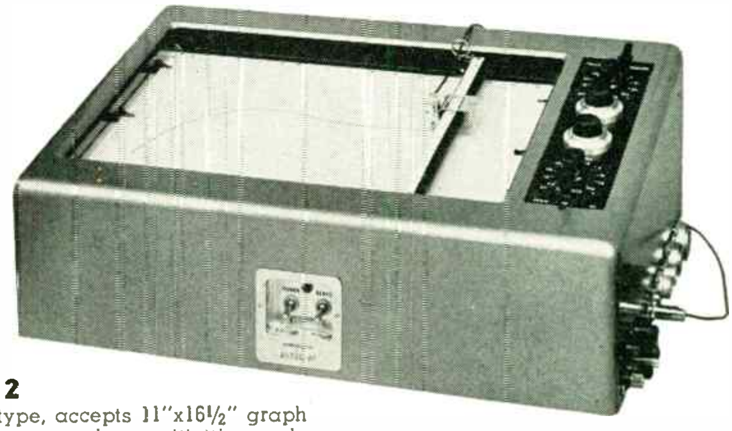

**MODEL 2**
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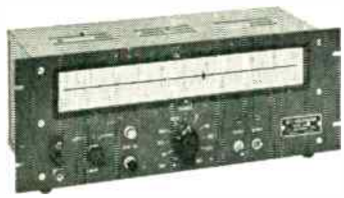

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## Section Meetings

*(Continued from page 106A)*

HUNTSVILLE

"A Contribution to Microwave Measurements," by Dr. F. J. Tischer, Redstone Arsenal; April 26, 1955.

General Electric "House of Magic"; May 13, 1955.

INYOKERN

"Digital Techniques Applied to Aircraft Fire Control Systems," by Robert McIntyre, Librascope, Inc.; March 21, 1955.

"Electrical Ceramics," by W. A. East, Consulting Engineer; April 18, 1955.

"A System for Automatic Target Acquisition by a Phototheodolite Utilizing a Remote Tracking Radar," by James Sherwin, U. S. Naval Ordnance Test Station; May 9, 1955.

ITHACA

"Pulse Code Modulation," by S. Shriner, Federal Telecommunications Labs.; May 5, 1955.

LITTLE ROCK

"High Fidelity Systems for Home Use," by J. Spilman, W. M. McClanahan and Harry Cooke; May 10, 1955.

LONDON

"The Application of Transistors in Computer Circuitry," by C. D. Florida, The Defence Research Board; April 26, 1955.

LONG ISLAND

"Nuclear Instrumentation," by R. L. Chase, Brookhaven National Laboratories; May 10, 1955.

LOS ANGELES

"The Role of Scientific Research in the Development of Missile Systems," by Dr. Ernst Krause, Lockheed Missile Systems Division, and "A Review of Progress in the Land Mobile Communication Service," by J. F. Byrne, Motorola Riverside Research Laboratory; May 3, 1955.

LOUISVILLE

"The Application of IBM to Engineering and Statistical Problems," by P. H. Sterbenz, International Business Machines Corp.; February 10, 1955.

"Recent Advances in the Reproducing Art," by A. M. Wiggins and Howard Souther, both of Electrovoice, Inc.; March 10, 1955.

"High Fidelity—Past and Future," by Marvin Camras, Armour Research Foundation; April 14, 1955.

MIAMI

Student papers by the following University of Miami students: S. Aragones, R. Watts, Dave Wensley, Ray Harper, Harry Hopcroft and R. N. Stock; April 29, 1955.

NEW YORK

"Application of Communication Concepts to Infrared Problems," by Dr. M. J. E. Golary, Squier Signal Lab.; April 6, 1955.

"An Experimental Automobile Receiver Employing Transistors," by T. O. Stanley, and "5 Watt Transistor Amplifier," by A. I. Aronson, both of David Sarnoff Research Center; May 4, 1955.

OTTAWA

"Current Problems in UHF and Microwave Multiplex Communication Systems," By Dr. H. J. Von Baeyer, Dept. of Defence Production; Election of Officers; April 14, 1955.

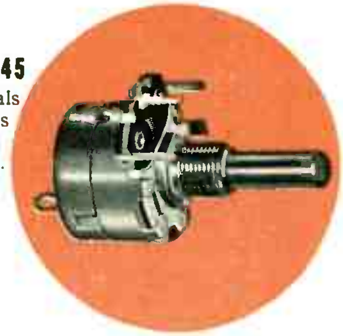
Field Trip to RCA Victor Television and Home Receiver Plant and Fort Wellington at Prescott, Ontario; May 13, 1955.

*(Continued on page 111A)*

# FOR YOUR AUTOMATION PROGRAM

## VARIABLE RESISTORS FOR PRINTED CIRCUITS

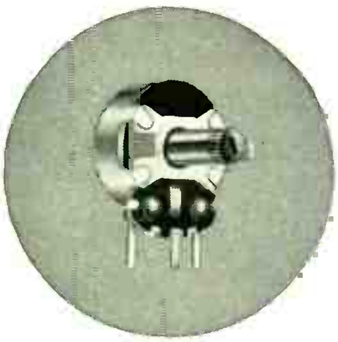
**Type UPM-45**
For TV preset control applications. Control mounts directly on printed circuit panel with no shaft extension through panel. Recessed screwdriver slot in front of control and 3/8" knurled shaft extension out back of control for finger adjustment. Terminals extend perpendicularly 7/32" from control's mounting surface.



**Type GC-U45**
Threaded bushing mounting. Terminals extend perpendicularly 7/32" from control's mounting surface. Available with or without associated switches.

**Type U70 (Miniaturized)**
Threaded bushing mounting. Terminals extend perpendicularly 5/32" from control's mounting surface.

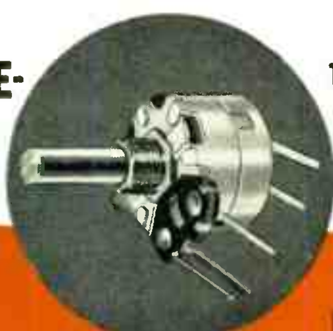

**Type YGC-B45**
Self-supporting snap-in bracket mounted control. Shaft center spaced 29/32" above printed circuit panel. Terminals extend 1-1/32" from control center.



**Type XP-45**
For TV preset control applications. Control mounts on chassis or supporting bracket by twisting two ears. Available in numerous shaft lengths and types.

**Type XGC-45**
For applications using a mounting chassis to support printed circuit panel. Threaded bushing mounting.

## VARIABLE RESISTORS FOR SOLDERLESS "WIRE-WRAP" CONNECTIONS



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**RETURN SIGNAL AMPLIFIER AND ATTENUATOR:** Covering range of 40 millivolts to 200 volts.
**DISTORTION:** At least 40 db below maximum output.
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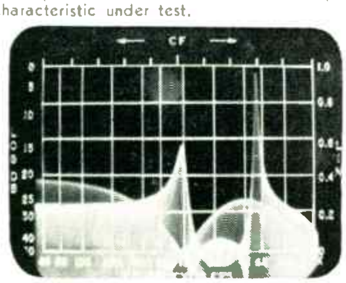
Separate output for variable marker for use with Signal Alternator Model SW-1 for alternate presentation of marker and response characteristic under test.



Response of an audio band stop filter with null marker showing frequency of one peak.

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|---|---|---|---|---|---|---|---|---|
| PM-12 | 0-50 kw | 1.15 | 60 watts | To Be Specified | 51.5 | 110 volts 60 cps | ±10% | .2-700MC |
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## Section Meetings

(Continued from page 108A)

PHILADELPHIA

"Fundamentals of Color Television," by J. W. Wentworth, RCA; March 1, 1955.

"Color Signal Generating Equipment," by Dick O'Brien, CBS; March 8, 1955.

"Color Reproducing Tubes and Associated Components," by B. Loughlin, Hazeltine Corp.; March 15, 1955.

"Color Decoding Circuits," by Jack Avins, RCA Industry Service Lab.; March 22, 1955.

"Measurement and Equalization of Amplifiers and Transmission Systems for Color TV Service," by Hugh Kelly, Bell Laboratories; March 29, 1955.

"Colorimetry Problems in Color TV and the Effect of Transmission Errors on Color Reproduction," by Harold Weiss General Electric Company; April 5, 1955.

"Automatic Electronic Production," by Dr. J. D. Ryder, President, IRE; May 4, 1955.

PHOENIX

"Series Peaked Amplifier Analyzed on Analogue Computer," by Fred Schwepp, Student, University of Arizona, and Tapescript "The Physics of Music and Hearing," by W. A. Koch, Bell Telephone Labs.; April 8, 1955.

PORTLAND

"Engineering Aspects of UHF Booster Installations," and "Engineers are People," both by Dr. G. H. Brown, David Sarnoff Research Center; April 21, 1955.

(Continued on page 112A)

## Section Meetings

*(Continued from page 111A)*

Student Papers: "A Portable Sodium-Flash Recorder," by M. H. MacKenzie, "Elementary Magnetic Digital Computer Component," by R. W. Austad, and "Air-Borne Measurements of Low Frequency Effective Ground Conductivity in Alaska," by G. M. Stanley; April 30, 1955.

ROCHESTER

"Modern Loudspeaker Design," April 21, 1955.

ROME-UTICA

"More Significant Characteristics of Nonlinear Circuits," by Dr. Ernst Weber, Polytechnic Institute of Brooklyn; May 4, 1955.

SACRAMENTO

"Planning and Constructing a Television Station," by P. K. Onnigian; KBET-TV; Election of Officers; May 13, 1955.

SALT LAKE CITY

Demonstration of Boeing Analog Computer, by Dr. R. E. Stephenson, University of Utah; April 13, 1955.

General meeting; May 17, 1955.

SAN DIEGO

"Color Broadcasting Specifications and Applications," by George Jacobs, Wrather-Alvarez; April 2, 1955.

"Bio-Technology in Engineering," by Dr. A. M. Small, Navy Electronics Lab.; May 4, 1955.

TORONTO

"Thirty Years in Canadian Radio," by Miss Jane Gray, Commentator, CHML; April 18, 1955.

TULSA

"Dynamic Instrumentation," by C. M. Hathaway, Hathaway Instrument Co.; April 21, 1955.

TWIN CITIES

"Feedback Control Systems—Past, Present and Future," by Prof. T. J. Higgins, University of Wisconsin; April 26, 1955.

WASHINGTON

"The Navy 'Jim Creek' Transmitting Station," by Cdr. R. G. Bywater, USN and H. E. Dinger, Naval Research Lab.; May 9, 1955.

WILLIAMSPORT

"Guided Missiles are Smarter than People," by D. E. Mullen, General Electric; March 11, 1955.

"Automation," by Ben Warriner, General Electric Lab.; April 21, 1955.

"Traveling Wave Tubes," by Dr. R. G. Hutter, Sylvania Electric; May 17, 1955.

*SUBSECTIONS*

BUENAVENTURA

"Early Guided Missile Development in Germany," by Dr. Willy Fiedler, U. S. Naval Air Missile Test Center; April 14, 1955.

CENTRE COUNTY

"Automation," by Charles Godwin, Cornell University; election of officers; April 21, 1955.

CHARLESTON

Tour of Television Station and Radio Studio WCSC conducted by Dick Hart, Station Engineer; April 26, 1955.

EAST BAY

"Radio Astronomy and Its Engineering Aspects," by Dr. R. N. Bracewell, University of California; May 3, 1955.

*(Continued on page 115A)*

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## Section Meetings

(Continued from page 112A)

ERIE

"The Theory and Application of Quartz Ultrasonic Delay Lines," by J. M. Wolfskill and "The Theory and Application of Quartz Crystal Units," by R. H. Tuznik, both of Bliley Electric Company; April 28, 1955.

LANCASTER

"An Excursion in Electronics," by C. N. Hoyler, RCA Laboratory, May 11, 1955.

MID-HUDSON

"Magnetic Core Circuitry," by Dr. An Wang, Wang Labs.; April 19, 1955.

MONMOUTH

"Ferrites at Microwave Frequencies," by A. G. Fox, Bell Telephone Labs.; May 18, 1955.

ORANGE BELT

"Stereophonic (3D) Sound in Your Home," by C. M. Brainard, Master Electronics Company; April 25, 1955.

PALO ALTO

"Pulse Applications of Junction Transistors," by John Linvill, Stanford University; April 21, 1955.

WICHITA

"Design of Airborne Automatic Antenna Tuners," by L. Hutton, Boeing Airplane Co.; April 27, 1955.

## Student Branch Meetings

UNIVERSITY OF AKRON (IRE-AIEE BRANCH)

"Engineering Ethics" by R. D. Landon, Dean, College of Engineering, U. of Akron, and Election of Officers; April 25, 1955.

ALABAMA POLYTECHNIC INSTITUTE (IRE BRANCH)

"Electronics in Guided Missles" by S. Johnson; April 25, 1955.

UNIVERSITY OF BRITISH COLUMBIA (IRE-AIEE BRANCH)

"Some Aspects and Future Trends of Electrical Engineering" by D. Carpenter, President, Research Industries, Ltd., and Election of Officers; March 30, 1955.

POLYTECHNIC INSTITUTE OF BROOKLYN (IRE-AIEE BRANCH) EVE. DIV.

"The New High Speed Electronic Printer" by Edmund DiGiulio, Field Engrg. Mgr., Control Instrument Co., Brooklyn, N. Y., and Business Meeting; April 28, 1955.

Film, "A is for Atom" and Election of Officers; May 11, 1955.

BROWN UNIVERSITY (IRE-AIEE BRANCH)

Election of Officers; March 16, 1955.

"An Inexpensive Voltage Regulator," by James Davis, Undergraduate, Brown U.; March 30, 1955.

CLARKSON COLLEGE OF TECHNOLOGY (IRE-AIEE BRANCH)

Election of Officers; April 21, 1955.

"Motors and Motor Control," by D. B. Seymore, Westinghouse; April 28, 1955.

"I.B.M. Computers & Accounting Machines," by Mr. Markle, I.B.M. and Film, "Piercing the Unknown," and Business Meeting; May 12, 1955.

(Continued on page 117A)

## Student Branch Meetings

(Continued from page 115A)

UNIVERSITY OF COLORADO (IRE-AIEE BRANCH)

"The Resistor—A Simple Element?" by Harry Bishop, Hytronic Measurement Assoc.; May 11, 1955.

COLUMBIA UNIVERSITY (IRE-AIEE BRANCH)

"Magnetic Recording" and "Method for Time or Frequency Compression-Expansion of Speech" (PGA Tape Scripts); April 18–19, 1955.

Film, "Research and Development at Hughes Aircraft Co."; May 3, 1955.

UNIVERSITY OF CONNECTICUT (IRE-AIEE BRANCH)

"X-Ray and High Power Electron Tubes," by Chester A. Kirka, Machlett Labs. and Election of Officers; March 31, 1955.

"Instrument Servos & Their Application to Aircraft Auto-Pilots," by F. W. Campbell, Sperry Gyroscope; April 28, 1955.

UNIVERSITY OF DELAWARE (IRE-AIEE BRANCH)

"Africa," by Professor Earl Parker Hanson; March 7, 1955.

"The Germanium Story," by Dr. S. M. Christian, "A P-N-P Alloy Junction Transistor for Radio Frequency Amplification," by Dr. C. W. Mueller and "An Experimental Transistor Personal Broadcast Receiver," by L. E. Barton; April 19, 1955.

UNIVERSITY OF DETROIT (IRE-AIEE BRANCH)

Student Paper Contest: "An Introduction to the Use of Symbolic Logic in the Design of Switching Circuits," by J. Dennis Kennedy and "Electronic Comparator," by Albert Vanschaemelhout; April 20, 1955.

"Success Story" (tape and slide talk) and "An Introduction to Miniature Low Pass Filters for Telemetering in Guided Missile Research," by Victor Schutzwhol; May 5, 1955.

DREXEL INSTITUTE OF TECHNOLOGY (IRE-AIEE BRANCH)

"Carbon in Electrical Engineering," by Bernard Silver, Electronite Carbon Co.; April 5, 1955.

"Microwaves," by Richard A. Dibos, Philco Corp.; April 28, 1955.

UNIVERSITY OF FLORIDA (IRE-AIEE BRANCH)

"The Trends in Industrial Electrical Distribution Systems," by C. F. Kucera, Allis-Chalmers, and Business Meeting; February 14, 1955.

"Research & Development of Transformers" (Film); March 14, 1955.

"Digital Computers," by Mr. Zyrak and Mr. Rich, both of Lincoln Lab., M.I.T.; March 28, 1955.

Election of Officers; May 9, 1955.

IOWA STATE COLLEGE (IRE-AIEE BRANCH)

"Hints in Interviewing for Jobs," by J. J. Jondle, Student and Election of Officers; April 27, 1955.

"Development of Radar," by C. J. Marshall, IRE Regional Director, Region 5; May 6, 1955.

LAFAYETTE COLLEGE (IRE-AIEE BRANCH)

"Tests on 69Kv Horned Air Brake Switches," by Mr. Wesley Smith and Mr. Jack McDonald, Pennsylvania Power and Light Company; March 10, 1955.

"Electronics in Medicine," by Ray Wiech, Lafayette Student '56 and Election of Officers; April 14, 1955.

"High Speed Electronic Flash Photography," by Richard B. Hanbicki, Student, Class '56 and General Meeting; May 12 1955.

(Continued on page 118A)

# Student Branch Meetings

*(Continued from page 117A)*

University of Louisville (IRE-AIEE Branch)

Films, "The Germanium Story" and "A P-N-P Alloy Junction Transistor for Radio Frequency Amplification"; May 5, 1955.

University of Massachusetts (IRE-AIEE Branch)

"High Fidelity," by D. R. Von Recklinghausen, H. H. Scott Inc. and Election of Officers; April 25, 1955.

Business Meeting; April 27, 1955.

University of Miami (IRE Branch)

Film, "Antennas and Creation of Radio Waves" and "Water Purification"; April 21, 1955.

"Telemetering of Integrated Type Meter Readings over Power Carrier Current Channels," by C. W. Cogburn and H. A. Lewis, Joint Lecturers, Florida Power and Light Co.; May 3, 1955.

University of Michigan (IRE-AIEE Branch)

Film, "Industrial Research and Development"; March 23, 1955.

"Single Side Band," by Mr. Bruene, Collins Radio Co. and Election of Officers; April 26, 1955.

Michigan State College (IRE-AIEE Branch)

"Expanding Opportunities for Tomorrow's Electrical Engineers" (tape recording and slides); April 19, 1955.

University of Missouri (IRE-AIEE Branch)

"Unlimited Horizons," by B. J. George, Kansas City Power and Light Co.; April, 1955.

"Getting Impurities out of Oils by Electric Fields," by G. C. Pauls, Petrolite Corporation; May 3, 1955, and "Join IRE as a Student," by Mr. Efthin, Secretary of St. Louis IRE, and Election of Officers, also May 3, 1955.

Montana State College (IRE-AIEE Branch)

"Hammond Organs," by K. J. Bailey, Northwest Hammond Organ Studios; April 20, 1955.

University of New Hampshire (IRE-AIEE Branch)

Student Paper Contest; April 15, 1955.

College of the City of New York (IRE Branch)

Talk on Brookhaven Laboratories by Prof. H. B. Hansteen, Chairman of E. E. Dept., CCNY and Brookhaven Labs.; April 28, 1955.

New York University (IRE Branch) Day Div.

Election of Officers; April 18, 1955.
General Meeting; April 25, 1955.

North Carolina State College (IRE-AIEE Branch)

"Symbolic Logic in Circuit Design," by Mr. O'Conner, I.B.M.; April 19, 1955.

"The Basic Principles of the Operation of Magnetic Amplifiers," by Clarence Sturzenbecker, Western Electric Co. and Installation of Officers; May 10, 1955.

University of North Dakota IRE-AIEE Branch

"Swearing as a Communicative Device," by Prof. Frederick Kroeger, Faculty, UND; April 27, 1955.

North Dakota Agricultural College (IRE-AIEE Branch)

Film, "Electric Proving Ground" and General Meeting; May 11, 1955.

*(Continued on page 120A)*

## Student Branch Meetings

*(Continued from page 118A)*

UNIVERSITY OF NOTRE DAME (IRE-AIEE BRANCH)

"The Growing Pains within the Telephone Company," by Robert Holmes, Indiana Bell Telephone Co. and General Meeting; March 31, 1955.
Student Paper Contest; April 26, 1955.

OHIO UNIVERSITY (IRE-AIEE BRANCH)

"Analog Computors," by W. C. Meilander, Goodyear Aircraft Corp.; April 19 1955.
"High Fidelity," by David H. O'Brien, Anderson High Fidelity Center and Election of Officers; May 10, 1955.

OKLAHOMA A & M COLLEGE (IRE-AIEE BRANCH)

"Real Time Digital Computer Control," by J. F. Nolan and H. E. Anderson, Lincoln Labs, M.I.T.; March 31, 1955.

OREGON STATE COLLEGE (IRE BRANCH)

"Digital Computers and Automatic Control," by David R. Brown, MIT Computer Lab.; February 15, 1955.

UNIVERSITY OF PENNSYLVANIA (IRE-AIEE BRANCH)

"An Experimental Transistor Personal Broadcast Receiver" (tapescript) and talk by L. E. Barton, RCA Labs.; May 2, 1955.

UNIVERSITY OF PITTSBURGH (IRE BRANCH)

Film, "The Inner Man Steps Out"; April 14, 1955.
"Ethics and Professional Practice," by Carl Long, Pittsburgh Chapter of PSPE; April 28, 1955.
"Closed Television at Pitt.," by Prof. W. S. Lacey, Faculty, U. of P. and Election of Officers; May 12, 1955.

RENSSELAER POLYTECHNIC INSTITUTE (IRE-AIEE BRANCH)

"Atomic Power," by Commander Naymark, USN Retired, presently with G.E.; April 26, 1955.
"Progress in Flight Control Development," by Robert Wanamaker, G.E. and Election of Officers; May 10, 1955.

SAN JOSE STATE COLLEGE (IRE BRANCH)

"Klystrons—What They are and Where They are Used," by B. Hadley, Varian Associates; February 17, 1955.
"IRE's Relation with Student Members" and "Transistor Switching Circuits" by Dr. Pettit, Prof. of E.E., Stanford U.; March 10, 1955.
"Multiplexing" and "RCA Engineering," by E. Napier, RCA; April 21, 1955.

SOUTH DAKOTA SCHOOL OF MINES & TECHNOLOGY (IRE BRANCH)

Election of Officers and "Success Story," by L. J. Linde, Allis-Chalmers (slides); April 13, 1955.

UNIVERSITY OF SOUTHERN CALIFORNIA (IRE-AIEE BRANCH)

"History of RCA," by Maurice Franko, RCA; April 13, 1955.
General Business Meeting; April 20, 1955.

VILLANOVA UNIVERSITY (IRE-AIEE BRANCH)

"The G.E. Solar Battery," by Dick Burke, "Tape Recorder Design" by Tulvio Durand and "Functional Color," by James Wresnski, all Senior E.E. Students, Villanova U.; March 30, 1955.
"Cromby Generating Station," by Wilmon Ware, Philadelphia Electric Co.; April 13, 1955.
"Safety," by M. Frank V. Connelly, Sun Oil Company; April 27, 1955.

*(Continued on page 143A)*

... with your future in mind

**_Check the chart below for positions which interest you most..._**

| FIELDS OF ENGINEERING ACTIVITY | Electrical Engineers 1-2 | Electrical Engineers 2-3 | Electrical Engineers 4+ | Mechanical Engineers 1-2 | Mechanical Engineers 2-3 | Mechanical Engineers 4+ | Physical Science 1-2 | Physical Science 2-3 | Physical Science 4+ | Chemistry Ceramics Glass Technology Metallurgy 1-2 | Chemistry Ceramics Glass Technology Metallurgy 2-3 | Chemistry Ceramics Glass Technology Metallurgy 4+ |
|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | TYPE OF DEGREE AND YEARS OF EXPERIENCE PREFERRED | | | | | | | | | | | |
| **SYSTEMS** *(Integration of theory, equipments, and environment to create and optimize major electronic concepts.)* | | | | | | | | | | | | |
| **AIRBORNE FIRE CONTROL** | | | W | | | | | | W | | | |
| **DIGITAL DATA HANDLING DEVICES** | | | C | | | C | | | C | | | |
| **MISSILE AND RADAR** | | | M | | | M | | | M | | | |
| **INERTIAL NAVIGATION** | | | M | | | M | | | M | | | |
| **COMMUNICATIONS** | | | C O | | | | | | C O | | | |
| **DESIGN • DEVELOPMENT** | | | | | | | | | | | | |
| **COLOR TV TUBES**—Electron Optics—Instrumental Analysis—Solid States (Phosphors, High Temperature Phenomena, Photo Sensitive Materials and Glass to Metal Sealing) | L | L | L | L | L | L | L | L | L | L | L | L |
| **RECEIVING TUBES**—Circuitry—Life Test and Rating—Tube Testing—Thermionic Emission | H | H | H | | H | H | | H | H | | H | H |
| **SEMI-CONDUCTORS**—Transistors—Semi-Conductor Devices | H | H | H | | | | H | H | H | | | |
| **MICROWAVE TUBES**—Tube Development and Manufacture (Traveling Wave—Backward Wave) | | H | H | | H | H | | H | H | | H | H |
| **GAS, POWER AND PHOTO TUBES**—Photo Sensitive Devices—Glass to Metal Sealing | L | L | L | L | L | L | L | L | L | L | L | L |
| **AVIATION ELECTRONICS**—Radar—Computers—Servo Mechanisms—Shock and Vibration—Circuitry—Remote Control—Heat Transfer—Sub-Miniaturization—Automatic Flight—Design for Automation—Transistorization | X | F X | M C F X | X | F X | M C F X | X | F X | M C F X | | | |
| **RADAR**—Circuitry—Antenna Design—Servo Systems—Gear Trains—Intricate Mechanisms—Fire Control | X | F X | M C F X | X | F X | M C F X | X | F X | M C F X | | | |
| **COMPUTERS**—Systems—Advanced Development—Circuitry—Assembly Design—Mechanisms—Programming | C | C F | M C F | C | C F | M C F | C | C F | M C F | | | |
| **COMMUNICATIONS**—Microwave—Aviation—Specialized Military Systems | | F | M C F | | F | M C F | | F | M C F | | | |
| **RADIO SYSTEMS**—HF-VHF—Microwave—Propagation Analysis—Telephone, Telegraph Terminal Equipment | | O | O F | | O | O F | | O | O F | | | |
| **MISSILE GUIDANCE**—Systems Planning and Design—Radar—Fire Control—Shock Problems—Servo Mechanisms | | F | M F | | F | M F | | F | M F | | | |
| **COMPONENTS**—Transformers—Coils—TV Deflection Yokes (Color or Monochrome)—Resistors | | C | C | | C | C | | C | C | | | |
| **MACHINE DESIGN** Mech. and Elec.—Automatic or Semi-Automatic Machines | | H | H | | H | H | | H | H | | | |

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**C**—Camden, N. J.—in Greater Philadelphia near many suburban communities.

**F**—Florida—on east central coast.

**H**—Harrison, N. J.—just 18 minutes from downtown New York.

**L**—Lancaster, Pa.—about an hour's drive west of Philadelphia

**M**—Moorestown, N. J.—quiet, attractive community close to Phila.

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**W**—Waltham, Mass.—near the cultural center of Boston.

**X**—Los Angeles, Calif.—west coast electronics center.

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**RADIO CORPORATION of AMERICA**

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The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. . . . .

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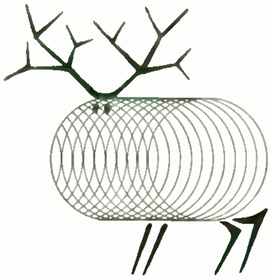
Excellent opportunities with expanding Radiation Instrument Development Group for recent graduates in E.E. or Physics and /or engineers with 1-5 years experience. Send resume and salary required to Tracerlab, Inc., Western Div., 2030 Wright Ave., Richmond 3, Calif.

### PROFESSOR

Professorial position open on the faculty of the Dept. of Electrical Engineering of a midwestern university. Faculty rank and salary will be determined upon qualifications of the applicant selected. Position starts in September 1955. A Ph.D. degree is desirable, but not necessary, if the applicant has a sufficiently strong research and publication record. Box 809.

*(Continued on page 126A)*



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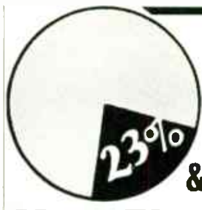



# Electronic ENGINEERS & PHYSICISTS

***More Than 23% of Your Time*** is spent at work. Doesn't it seem worth while, then, to spend a little time investigating the job possibilities at National Company where you can spend that important one quarter of your life in satisfying and rewarding work.

This future-minded firm has grown continuously in the field of electronic design and manufacture since 1914. Their products are in use in the scientific, governmental, industrial, and commercial fields. Their employees work in warm, friendly surroundings with due recognition given for effort and talent.

Their present needs are for creative people who can originate ideas and develop them through the prototype stage.

If you sincerely believe you can fit in with our ideas and have a B.S. or higher degree, you are invited to send your resume to:

**Mr. John A. Bigelow**

***National***

EST. 1914

NATIONAL COMPANY, INC.
60 Sherman St. Malden, Mass.

# Positions Open

*(Continued from page 124A)*

### SALES ENGINEER

We have an opening for a technical salesman, preferably with experience in the development and design of capacitors. Headquarters will be in Lee, Massachusetts. After a training period, the position will involve some travelling, principally in the east. The right man can expect an attractive fixed salary and expense; with an excellent opportunity for advancement. All inquiries will be treated in strictest confidence, and should be directed to the personal attention of Mr. Peter Schweitzer, Peter J. Schweitzer, Inc., 261 Madison Ave., New York 16, N.Y.

### ELECTRONIC SALES ENGINEER

A growing group of electronic manufacturers' representatives serving in the southeast since 1924 requires an electronic sales engineer. Applicant should be married and be between the age of 30-35, willing to travel two-thirds of the time. Contracts cover industrials and distributors. Prospects for partnership interest to qualifying applicant. Send complete resume with photograph. Box 812.

### PRODUCTION ENGINEER

Electronic engineer with production experience to supervise production and test personnel and to set up production and test facilities for small electronics manufacturer. Salary $4000 to $6000. Live in attractive small town with the opportunity of taking graduate work at the Pennsylvania State University. Reply to Community Engineering Corp., P.O. Box 824, State College, Pa.

### ENGINEERS

Television receiver deflection systems engineers wanted. Development and product design. Both color and monochrome. Send resumes to Dept. RT-1, Technical Employment Office, General Electric Company, Electronics Park, Syracuse, N.Y.

### ELECTRONIC ENGINEER

Electronic Engineer with audio experience desired on Consultant basis. Write, giving general information and fee expected. Box 813.

### PATENT ATTORNEY OR AGENT

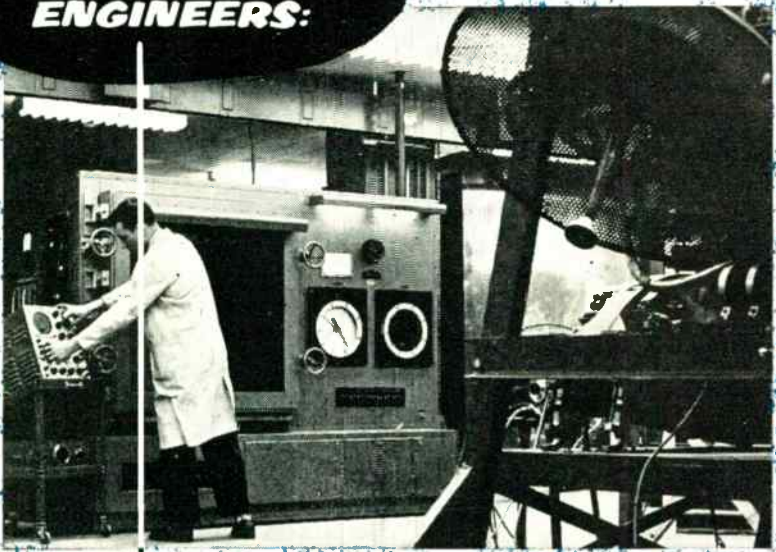
Capable and experienced man with good electronics background. Chemical experience also desirable. To engage in patent work in electron circuits and devices, in company carrying forward advanced research and development programs in monochrome and color TV, transistors and other semi-conductive devices, vacuum tubes, UHF and microwave systems, preferably with minimum of supervision. Prior experience and educational qualifications will be recognized in scope of work, responsibility and compensation. Box 814.

### ELECTRICAL ENGINEERS—PHYSICISTS

Permanent staff positions are open for BS or advance degree electrical engineers or physicists with a minimum of 3 to 5 years electro-magnetic and propagation experience in the fields of direction finding, antenna design, microwave and radar techniques. Send statement of qualifications and interest in industrial and defense research to S. J. Keane, Physics Dept., Southwest Research Institute, 8500 Culebra Road, San Antonio, Texas.

*(Continued on page 128A)*

## ELECTRONIC ENGINEERS:



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FIELD ENGINEERING
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**TO APPLY-**

**Send resume outlining education and experience to:**

Technical Director
Dept. 192
Westinghouse Electric Corporation
2519 Wilkens Avenue
Baltimore 3, Md.

**ILLUSTRATED BROCHURE WILL BE SENT TO ALL APPLICANTS.**


## Positions Open

(Continued from page 126A)

### PHYSICIST OR ELECTRONICS ENGINEER

Physicist or electronics engineer to design, construct and install set-ups to obtain data on engine ignition, performance. Diversified projects might require mechanical or electronic instrumentation, also design of auxiliary control circuits. Electric Auto-Lite Company, Toledo 1, Ohio.

### ENGINEER

Exceptional opportunity. If you have experience in design, construction and evaluation of high voltage, high frequency circuits. Requires ability to design circuits incorporating transistors, magnetic amplifiers and semi conductors. Electric Auto-Lite Company, Toledo 1, Ohio.

### INSTRUCTOR OR ASSISTANT PROFESSOR

University in southwest has an opening for either an instructor or an assistant professor of electrical engineering in the communications field. Salary and title commensurate with education and experience. Instructor with BS permitted to take courses for an advanced degree. Box 817.

### ELECTRONIC ENGINEER OR PHYSICIST

BSEE or BS Physics. 2 years experience in acoustics, electronic instrumentation or equivalent. Imaginative, resourceful person with good working knowledge of electronic circuits and physics is needed for research in underwater sound and oceanographic instrumentation. Must be unusually versatile and have a sincere interest in the marine sciences. Occasional periods at sea. Faculty rating. Moderate salary. Send complete resume. Marine Laboratory, University of Miami, Coral Gables, Florida. Att: Dr. H. B. Moore.

(Continued on page 130A)

# ENGINEERS

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MAGNETIC AMPLIFIER DESIGN AND ANALYSIS
DEVELOPMENT AND APPLICATION OF PRINTED CIRCUITS
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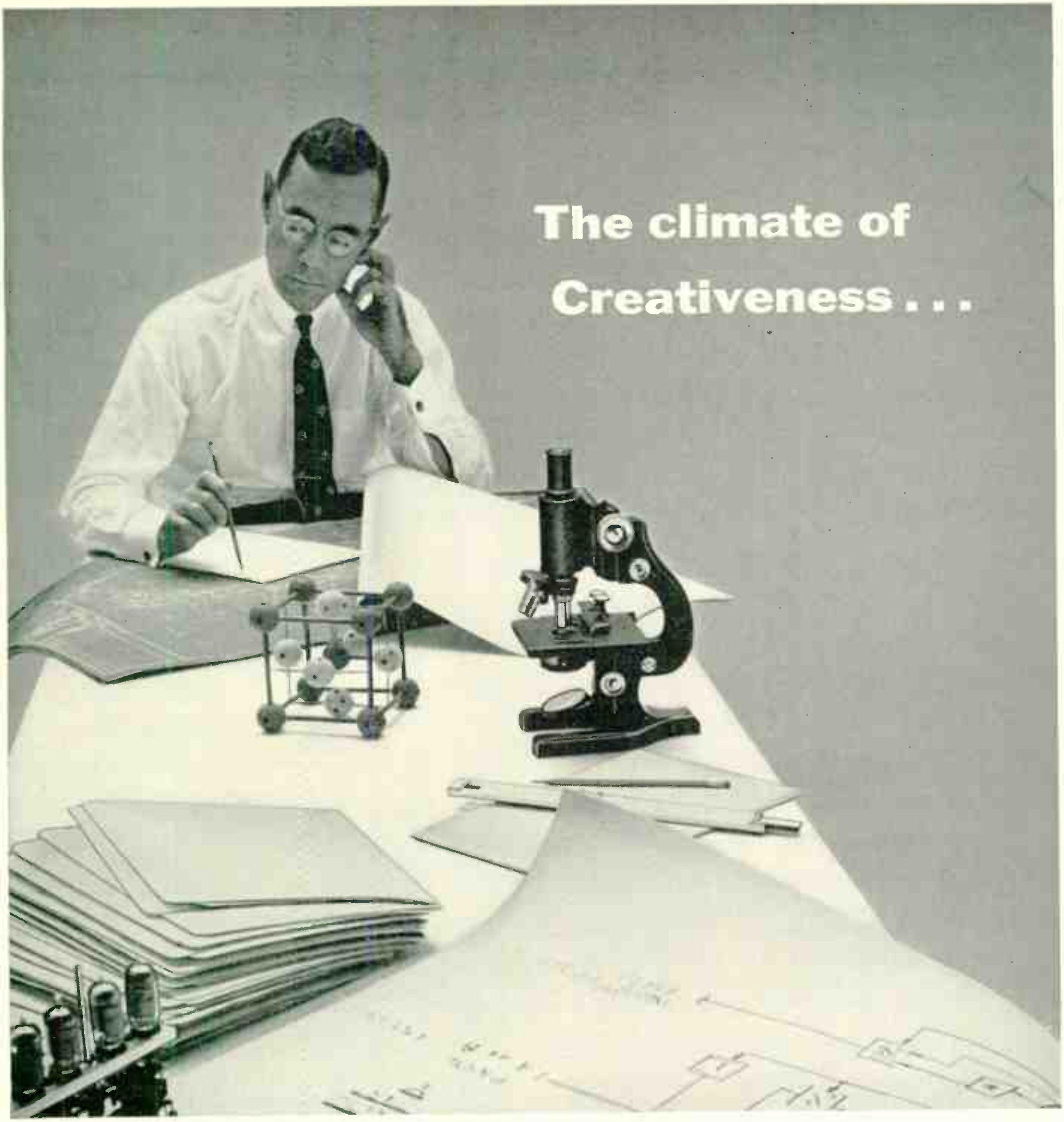


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As several surveys* have pointed out recently, too often an engineer's lot is not a happy one. Sometimes he feels he is regarded merely as an unusually versatile machine, rather than a person with skills, and aspirations, and ideas.

As far as we know, none of these surveys got around to ECA... or the statistics might have looked brighter. We are thoroughly convinced that engineers are people, they are professionals, and they are to be treated and respected as such.

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The work itself is of a kind engineers find particularly interesting, and requires a highly creative and often unorthodox approach. Engineering such as this is responsible for ECA's leadership in the development of automatic controls, electronic business machines, analog and digital computers.

That's why ECA is interested in engineers gifted with unusual talent and imagination. If you are such a man—looking for an environment in which your abilities will be appreciated and in which you can find both professional and personal satisfaction —we'd like to talk with you.

Please write details of your background and experience to Mr. W. F. Davis, Dept. 719.

*
1. Professional Engineers Conference Board for Industry survey, "How to Train Engineers in Industry"
2. University of Chicago Survey of Employee Attitudes
3. National Society of Professional Engineers, "A Professional Look at Engineers in Industry"

**ELECTRONICS CORPORATION OF AMERICA**

77 Broadway Cambridge 42, Mass.

## Positions Open

(Continued from page 128A)

### ELECTRONIC ENGINEER

Openings for engineers with 1 to 5 years experience in circuit design for communications, medical electronics instrumentation and computer fields. Small, rapidly growing electronics company with unusual profit sharing and patent program. All voting stock owned by employees; tuition reimbursement plan, medical aid plan. Write or call American Electronic Laboratories, Inc., 641 Arch St., Philadelphia, Pa., Att: Dr. Riebman.

### ASSISTANT PROFESSOR, COMMUNICATIONS, ELECTRONICS

Real opportunity for young Ph.D. interested in both teaching and research. Rapid advancement possible for man with initiative and ability. Write: Chairman, Div. of Engineering, Brown University, Providence 12, R.I.

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Instructor in electrical engineering beginning September 1955. One interested in teaching fundamental electrical engineering subjects. Salary depends on qualifications. Opportunity for advancement. New building and equipment. Apply Chairman, Dept. of Electrical Engineering, University of Nebraska, Lincoln 8, Neb.

### PROFESSOR

Position as Assistant Professor in electronics at a Middle Atlantic university. Undergraduate and graduate instruction. Doctorate in E.E. or physics preferred, but will consider experience as alternate. Salary $5000 for a nine month session. Box No. 818.

(Continued on page 132A)

# ENGINEERS

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| BUSINESS MACHINE M.E. | $10-12,000 |
| COMPUTER ENGINEERS | $ 6-12,000 |
| LOGICAL DESIGN | $ 6-10,000 |
| MATHEMATICIANS | $ 8-12,000 |
| PULSE CIRCUITRY | $ 7-12,000 |
| AVIATION INSTRUMENTS | $10-12,000 |
| TECHNICAL WRITERS, E.E. | $ 7- 8,000 |
| PHYSICISTS, SOLID STATE | $ 9-12,000 |
| SHOCK & VIBRATION | $ 7- 9,000 |
| ELECTRONIC PROCESSING | To $ 8,500 |
| AIRBORNE ELECTRONICS | $ 8-14,000 |

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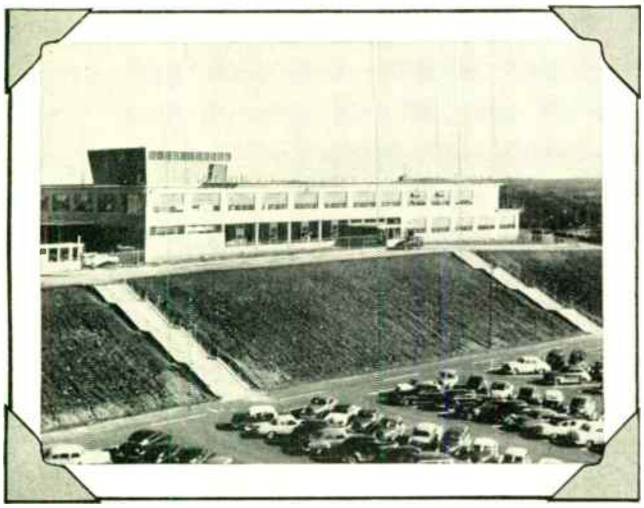

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**Waltham 54, Massachusetts**

## Positions Open

*(Continued from page 130A)*

### RADAR, SERVO COMPUTER ENGINEERS

Immediate openings on highest technical level with national leader in armament and commercial projects, in research, design and development of airborne fire control systems and guided missiles. Unlimited opportunity for rewarding career, with graduate study program, profit-sharing bonus, pension plan; accident, life, health insurance. Pleasant suburban location. Send resume to Engineering Personnel Manager, Emerson Electric of St. Louis, 8100 W. Florissant, St. Louis 21, Mo.

### PROFESSOR

Associate and Assistant Professor in Electrical Engineering Dept. of a state university located in the middle west. Specialization in electronics and microwaves required. Ph.D. or S.D. in electrical engineering or physics desired but will consider M.S. Part time research available. Box 819.

### ENGINEER

Engineer with Electrical Engineering or Physics degree and training in microwave devices. Position involves analyzing creative research for invention in the fields of microwave instruments, antennas and vacuum tubes. Patent knowledge beneficial, but not required. Submit inquiries to Engineering Personnel Division, Sperry Gyroscope Co., Great Neck, L.I., New York.

### DIRECTOR

Director, research and development guided missile field for top-rated company on East Coast. No fee. $15,000-$18,000. Guilford Personnel Service, American Bldg., Baltimore, Md.

### ELECTRONIC ENGINEERS

The U.S. Air Force has a continuing need for electronic engineers in overseas locations. Salaries range from $5000. to $8000. per annum. Application may be made to Chief, Overseas Employment Branch, Directorate of Civilian Personnel, Hqs., USAF, Att: AFPCP-D-3, Washington 25, D.C.

### SALES ENGINEER

Sales Engineer needed with BEE degree in Metropolitan New York area. Write in detail. Box 820.

### COIL ENGINEER

Coil engineer needed. Must be experienced in design of deflection yokes. Send resume of education and experience to Box 821.

# Boeing electrical engineers have vital and rewarding jobs

This is the Boeing B-52—mightiest of the world's jet bombers. The "wraps" are off it now. Some of the electrical engineers who made important contributions to its design are now at work on the B-52's further development. Many more are working on other exciting "years-ahead" airplanes and guided missiles. But these new Boeing aircraft are classified, and cannot be shown here.

Electrical engineers at Boeing have the satisfaction of vital jobs, where only constant experimentation can keep up with a progressing field. The importance of their art is steadily increasing in guided missile control, structural and flight test instrumentation, radar systems, acoustics, electro-mechanisms, antenna development, and many other specialties.

Boeing engineers can point with pride to recent developments like the B-47 and B-52 jet bombers, the KC-135—America's first jet tanker, and the IM-99 Bomarc pilotless interceptor. These Boeing engineers are now at work on widely diversified projects: rocket, ram jet and nuclear propulsion, supersonic flight, research in new materials, and many more. The result will be planes and missiles that will fly even faster, farther, and higher, and deliver an even greater punch.

These are evidences of Boeing's continuing growth—a growth made possible by uncompromising insistence on engineering excellence. Boeing employs twice as many engineers now than at the peak of World War II. But even more engineers are needed for Boeing's research, design and production teams.

If you want the satisfaction of doing an important job—if you want individual recognition and regular merit reviews—if you want to work with the finest equipment and the pick of the nation's engineers—then it will pay you to investigate career opportunities at Boeing.

**JOHN C. SANDERS, Staff Engineer—Personnel**
**Boeing Airplane Co., Dept. G-41, Seattle 14, Wash.**

Please send further information for my analysis. I am interested in the advantages of a career with Boeing.

Name ____________________

University or college(s) ________ Degree(s) ______ Year(s)____

Address ____________________

City ____________ Zone____ State ________

**BOEING**

*Aviation leadership since 1916*

SEATTLE, WASHINGTON    WICHITA, KANSAS

## 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 Institute publishes free of charge notices of positions wanted by I.R.E. 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 Institute necessarily reserves the right to decline any announcement without assignment of reason.

#### ADMINISTRATIVE ENGINEER

BSEE, LLB, Senior Member IRE, under 40. 6 years Signal Corps technical-administrative, 9 years research and manufacturing, extensive patent knowledge. Experienced in organizing and supervising group, personnel training, technical specification drafting, liaison. Desires supervisory position with coming research or manufacturing concern. Box 820 W.

#### ELECTRONICS ENGINEER

Striving, conscientious engineer for automation or missile development. BSEE 1951 Ill. Institute of Technology. 3 years experience in phono amplifiers, tape recording, multiplex communication in microwave. Box 822 W.

#### FIELD OR SALES ENGINEER

Experienced in electronics field engineering with diversified background in guided missile systems, test equipment and prototype electronics work. Retired Army Warrant Officer. Desires field or sales engineering position in Florida area. Box 823 W.

#### ENGINEER

2 years toward BEE degree. Age 27, married, 3 children. 1½ years Navy ETM 2 cl., WW II. 3 years field experience in radar and communication equipment. Presently engaged in flight simulator field. Desires any work permitting completion of studies toward degree. Box 824 W.

#### SALES—ADMINISTRATIVE ENGINEER

BSEE 1948. 7 years experience administrative engineering and military electronics equipment design in U.S. and abroad; present advancement too slow; age 34, married, 2 children; Tau Beta Pi; desires sales or administrative engineering electronics or allied field Southern California or Mediterranean area; present salary $7700. Box 825 W.

#### PATENT ENGINEER

Professor engineering, Senior Member IRE; registered to practice before U.S. Patent Office. Available on part time basis because of change in employer's operations. Full time offer considered. Box 826 W.

#### ELECTRICAL ENGINEER

BSEE 1950. Experience in petroleum geophysics, broadcast, communications radio, military electronics. Interested in electronic design, development, test or field engineering. Location preference southwestern or southern area. Will carefully consider any offer. Box 827 W.

*(Continued on page 136A)*

## Positions Wanted

### By Armed Forces Veterans

*(Continued from page 134A)*

#### ENGINEER

BSEE 1951, I.I.T. Eta Kappa Nu. Age 25. 2 years experience in adapting power system analysis to digital computers. 2 years experience at White Sands in programming and in trouble shooting a large scale digital computer. Desires position involving logical design of computers. Box 838 W.

#### ELECTRONIC ENGINEER

BEE 1953. Age 25. 5 years electronic experience, including 2 years shop and tes'ing, 1 year assistant project engineer on classified Navy project. 2 years as radio officer USAF. Desires responsible position in production or development. Box 839 W.

#### INSTRUCTOR

BSEE, MSEE. Age 29, married, 1 child. Graduate work in advanced electron tube circuits, network analysis and synthesis, and feedback systems. Mathematics minor. Licensed radio amateur. 1 year communications (R.F.) design and development; 1 year analogue computer circuit research and design; 3 years applied transistor research. Excellent references. Desires a full time teaching position in an institution that has an E.E. graduate school (Ph.D.) with privilege of engaging in six semester hours per semester of graduate study. Available September 1, 1955. Box 840 W.

#### ELECTRONICS SCIENTIST—INSTRUCTOR

Age 29, BS in mathematics, Summa cum laude. Minor in physics, graduate work in mathematics for 2 years. University mathematics instructor for 3 years. 1st class radio-telephone license with radar endorsement. Air Force electronics instructor for 3 years. Excellent references. Desires position with definite possibility of advancement. Box 841 W.

#### ELECTRICAL ENGINEER

BS in math. 1950, BEE 1954. IRE, AIEE, EIT (Ohio). Age 29, married, 1 child. 1 year experience with resistor components. Desires position in power or electronics field. Location Florida. Box 843 W.

#### COMPUTER ENGINEER

BEE, MSEE. 8 years extensive experience in analog computer field. Supervised large group of engineers in computer system and component design. Desires position of responsibility with opportunity for growth. Box 844 W.

#### ENGINEER

Age 28, married. BS in EE, working on MS. 2 years ETM 2/c in WW II; 5 years experience in VHF-UHF TV tuner design; precision radar receiver circuit design. some video and pulse work. Desires challenging position with medium size company. $7,000 minimum. Prefer east. Box 853 W.

*(Continued on page 139A)*

## Positions Wanted

### By Armed Forces Veterans

(Continued from page 136A)

**SENIOR ELECTRONIC DEVELOPMENT ENGINEER**

BS 1950. Age 25, married. 3 years video, pulse and ultrasonic systems development, project level, (patents) some guided missile system development, 2 years Army electronics instructor. Desires responsible R & D position. Box 854 W.

**ELECTRONIC ENGINEER**

BSEE January 1951. Age 27, married, 2 children. 3½ years experience in circuit development, instrument and communications systems planning, and ordnance testing. 1 year administrative and supervisory experience. Desires position offering responsibility and advancement in San Francisco area. Box 855 W.

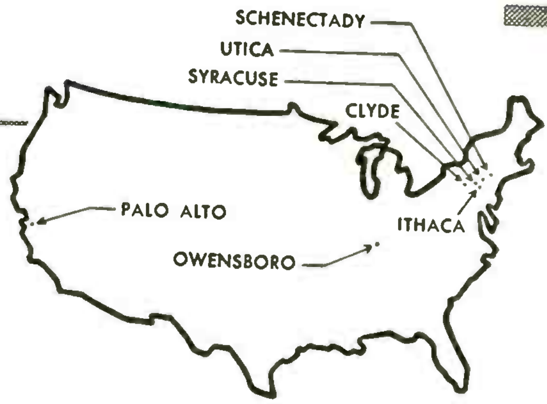
**ENGINEERING PHYSICIST**

BS Physics, MSEE. 4 years experience in industrial electro-mechanical, electronic instrumentation. 2 years instrumentation experience as assistant project engineer in nucleonics branch of Signal Corps on A.E.C. project. Employer going west. Desires eastern area position. Excellent record and references. Box 856 W.

**REGISTERED PROFESSIONAL ENGINEER**

Senior Member IRE seeks position where unique combination of technical and administrative ability and experience can be put to good use as director of engineering, technical manager,

(Continued on page 141A)




New
# GENERAL ELECTRIC
*Opportunities Throughout the Country In Advanced Electronic Developments*

At General Electric plants and laboratories from New York to California, G.E. engineers are constantly planning new and revolutionary advances in the field of electronics.

And with each new development, they are broadening the scope and opening new challenges and new opportunities in this young and fast-growing field.

Thus, the opportunity is ever-increasing at General Electric. If you are interested in taking on new challenges...in working with the finest facilities...in growing along with this leader in industry, you are invited to apply now for positions open at Schenectady, Utica, Ithaca, Syracuse, and Clyde, New York; Owensboro, Kentucky; and Palo Alto, Calif.

**ENGINEERS • PHYSICISTS**

*Positions available in the following fields:*

Advanced Development, Design, Field Service and Technical Writing in connection with:

**MILITARY RADIO & RADAR • MULTIPLEX MICROWAVE**
**MOBILE COMMUNICATION • COMMUNICATIONS**
**ELECTRONIC COMPONENTS**
**TELEVISION, TUBES & ANTENNAS**

Bachelor's or advanced degrees in Electrical or Mechanical Engineering, Physics, and experience in electronics industry necessary.

*Please send resume to:*
*Dept. 7-5 P, Technical Personnel*

**GENERAL ELECTRIC**

**ELECTRONICS PARK, SYRACUSE, N. Y.**

# tomorrow's OPPORTUNITY today

*for experienced*

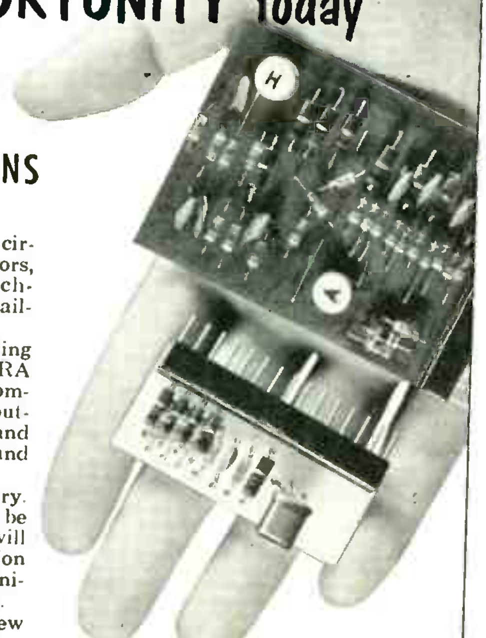
## ELECTRONIC ENGINEERS
*and*
## ELECTRONIC TECHNICIANS

If you can develop new computer circuits using magnetic cores, transistors, printed wiring, and other new techniques, we have a good position available for you.

You will work with the outstanding computer men who developed the ERA 1101, ERA 1102, and ERA 1103 Computer Systems, the Univac File Computer, ERA magnetic drum memories, and other equally famous Remington Rand systems.

Computer experience is not necessary. Your proficiency in related fields will be rewarded from the start, and you will work in the fastest-growing organization in the data-processing field. Opportunities for advancement will be numerous.

Positions are also available for new engineering graduates and technicians who want to learn digital techniques and systems. Pay, special benefits, and opportunities for advancement are most attractive.



ILLUSTRATED –
Designs for new Remington Rand ERA computers that are now under development.
Upper: general purpose digit register.
Lower: packaged transistor logic element.

*Please send an outline of your training and experience to Mr. J. N. Woodbury:*

**Remington Rand**
Engineering Research Associates Division
1902 W. Minnehaha Ave. • St. Paul W4, Minnesota

## Supervisory Engineers

for major manufacturer of
**Electronic Tubes**
Location: Eastern Seaboard

## RECEIVING TUBE DESIGN SECTION SUPERVISOR

## APPLICATION ENGR SECTION SUPERVISOR

Minimum 5 years experience in the same or related work.

- Salary open.
- Outstanding opportunities for professional progress.

Traveling and moving expenses paid. All replies confidential. Send resume to Box I.R.E. 649, 221 W. 41 St., N.Y.

## Positions Wanted

### By Armed Forces Veterans

(Continued from page 139A)

technical sales director, or similar position. Age 39. Broadly experienced in most aspects of electronics, radio and TV stations, networks, general management. Known in industry for books and articles on TV. Taught at 2 universities. Specialized in TV and films for TV. Prefer connection with broadcast organization or similar. Box 857 W.

#### ELECTRONIC DEVELOPMENT ENGINEER

BSEE, graduate work. Married. 4 years experience R & D on electronic ordnance devices. Performed original research, supervisory experience, Government contract support and liaison. Desires stimulating position in electronic weapon design group with opportunity for creative work and advancement. Box 858 W.

#### PATENT ENGINEER

Electrical engineer with patent experience in mechanical, electrical and nuclear arts. Member of Illinois and Fed. bars. Registered patent lawyer. At present available in midwest on part time basis or as independent. Box 859 W.

#### ENGINEER
#### WASHINGTON REPRESENTATIVE

Government engineer. Age 34. Nearing the Government ceiling, wants to grow. Has been Project Engineer or Supervisor at NBS, NOL, BuORD., BuShips and Army. Knows electronics for missiles, fuzes, ships, air defense and nuclear science; Government contract law and patents. Best schools and references. Box 860 W.

Farnsworth DIVISION OF IT&T

**has a future for**

# ELECTRONIC ENGINEERS

**in these specialized fields**

Career-minded men with several years specialized experience, and preferably with advanced degrees, are invited to join our rapidly expanding programs in industrial and military electronics.

***Address inquiries to:***
***Technical Employment Manager***

## TRANSISTOR CIRCUITRY

Transistorization and subminiaturization of advanced missile guidance and airborne radar systems.

## MICROWAVE ANTENNAS

Research, development, and design of airborne antennas in C, L, K, S, and X-bands for missiles and radar equipment.

## TEST EQUIPMENT

Test equipment design for radar and missiles systems and equipment manufacture.

## INDUSTRIAL TELEVISION

Development, design, and product engineering of closed-loop TV systems, including monitor, camera, and automatic controls.

## RADAR

Study, analysis, and development in highly advanced radar techniques and electronic counter-measures.

## MISSILES

Research, analysis, and development in guidance and control systems, components, and systems test equipment.

**FARNSWORTH ELECTRONICS CO.**
**Fort Wayne, Indiana**

***A division of International Telephone and Telegraph Corp.***

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# ELECTRONIC ENGINEERS

## This is the smart way to an unusual, creative career

The difference between an ordinary engineering job and a creative career at Air Associates is like night and day.

Here, assignments are exciting and challenging. Projects fire the imagination. Talent is quickly recognized, quickly rewarded.

If you are an engineer with vision, you belong here.

We're a young-minded organization with a lot of enthusiasm. As we enter our 28th year, we're happily confronted with the need for continued expansion.

You who join us can count on permanent association . . . advancement and salary dependent solely on ability . . . working side-by-side with other top-notch professionals. You will be encouraged to assume responsibility from the start. You will receive liberal company benefits. Your home will be in a quiet suburb with easy access to New York.

Openings are for section and project heads, seniors, intermediates and juniors to work on product developments of items like I.F. amplifiers, pulse circuits, servo systems, missile and aircraft control systems, advanced transmitters and receivers, and special devices of an R & D nature. For an appointment, write, wire or phone,

**James Moss, Personnel Manager**

*All Replies Answered Promptly*

**AIR ASSOCIATES, INC.**

**511 Joyce Street, Orange, New Jersey • Orange 6-4200**

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# ENGINEERS PHYSICISTS

**FOR THE NEW EXPANDING ELECTRONIC TUBE DIVISION OF**

# Westinghouse

**in Elmira, New York**

. . . where you will find challenging opportunities, and receive professional recognition. Work and live in the beautiful Southern Tier of New York State . . . in an uncrowded rural atmosphere, with complete advantages of city life.

DESIGN, DEVELOPMENT and APPLICATION ENGINEERS: Image Orthicon or Vidicon, receiving tubes.

MICROWAVE TUBE DESIGN ENGRS.: Design of Magnetrons, Traveling wave tubes, TR and ATR tubes, reference cavities.

*Above openings are for Engrs & Physicists with B.S., Master's or Doctor's degree. Also:*

ELECTRICAL ENGRS. For Equipment Design: Designing, costing, and guiding construction of processing and testing equipment, e.g. seasoning and test units, induction heaters, X-ray seasoning and test units, wave guide apparatus, high frequency oscillation test units, automatic receiving tube test circuits.

MANUFACTURING ENGRS.: Electrical and Mechanical engineers, and physicists to work on microwave, image orthicon, and other tube types.

*Interviews in your area, or travel expenses paid for interviews in Elmira. Send resume:*

**WESTINGHOUSE ELECTRIC CORP.**
Electronic Tube Div. Elmira, N.Y.

# Student Branch Meetings

*(Continued from page 120A)*

Virginia Polytechnic Institute (IRE-AIEE Branch)

Film on the Manufacturing of Paper and General Meeting; February 22, 1955.

Demonstration of Microwave Communications by representatives of Bivins and Caldwell; March 1, 1955.

Election of Officers and General Meeting; March 29, 1955.

Washington University (IRE-AIEE Branch)

Election of Officers and Business Meeting; March 31, 1955, and "A Thickness Gage Using Radioactive Isotopes," by Bush and Wallscheidt and "Analog Computer," by Rojko Also, March 31, 1955.

University of Washington (IRE-AIEE Branch)

"Civil Service Employment Opportunities," by Mr. Wihrow, Seattle City Light; April 6, 1955.

Wayne University (IRE-AIEE Branch)

"Engineering and Management," by Ray Plourde, Detroit Edison Co.; April 28, 1955.

University of Wisconsin (IRE-AIEE Branch)

"Patent Law," by John Leib, Patent Attorney for Allis-Chalmers Co.; April 21, 1955.

"The Network Analyzer," by James Skiles, Instructor at U. of Wisconsin and Election of Officers; May 9, 1955.

Worcester Polytechnic Institute (IRE-AIEE Branch)

"Hi Fi Demonstration and Lecture," by D. R. Von Recklinghausen and E. G. Dyett, Jr., of H. H. Scott Co.; April 12, 1955.

University of Wyoming (IRE-AIEE Branch)

"Success Story," by Robert L. Hudelson, Student Member and General Meeting; April 26, 1955.

# ENGINEERS and DESIGNERS NEEDED

**for:**

MISSILE GUIDANCE SYSTEMS · BOMBING NAVIGATIONAL COMPUTER SYSTEMS · NEW CIVIL AVIATION PRODUCTS · JET AND TURBO-PROP ENGINE CONTROLS · AIRBORNE FIRE CONTROLS

**GM CAREER OPPORTUNITIES IN**

Systems Engineering and Analysis
Experimental Engineering
Development Engineering
Project Coordination
Design Engineering
Product Engineering
Product Evaluation
Field Engineering

**AND WE ALSO NEED:**

**DESIGNERS · CHECKERS · LAYOUT MEN**

Positions Are Permanent
Excellent Advancement Opportunities

Every inquiry treated confidentially and given immediate attention and personal reply.

**WRITE TODAY FOR EMPLOYMENT APPLICATION**

Mr. Louis R. Berks
Supervisor of Employment

**AC SPARK PLUG DIVISION**
**Precision Instrument Plant**

**GENERAL MOTORS CORPORATION**
**Milwaukee 2, Wisconsin**

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## PHOTO-EMISSIVE EXPERIENCE

Old established firm manufacturing electronic tubes desires services of several engineers or physicists familiar with photo-emissive and photo-conductive devices. Firm is embarking on manufacture of pickup and storage tubes. Chance to get in at the start of a new operation. Location New England. State complete qualification, salary desired, citizenship status and references. Reply:

*Box 810*
INSTITUTE OF RADIO ENGINEERS
1 East 79th Street, New York 21, N.Y.

---

## *SEMICONDUCTOR DEVELOPMENT*

**DEVICE ENGINEERS**
**APPLICATIONS ENGINEERS**
**PRODUCTION ENGINEERS**
**SALES ENGINEERS**

Responsible positions are now available with one of the leading and fastest growing semiconductor manufacturers. These are outstanding opportunities for Physicists and Engineers in research, development, or production of advanced germanium and silicon diodes and transistors.

Experience in semiconductors or other components such as tubes or capacitors is desirable.

Send resume or call

**Transitron electronic corporation**
**407 Main Street, Melrose, Mass.**
**MElrose 4-9600**

---

**Bendix needs**

# ELECTRONIC ENGINEERS

**in SOUTHERN CALIFORNIA**

Unusual engineering positions in Radar, Sonar and Telemetering are available at Pacific Division, Bendix Aviation Corporation in North Hollywood, California. These positions, which are directly associated with our long-range projects for industry and for defense, are available at all levels.

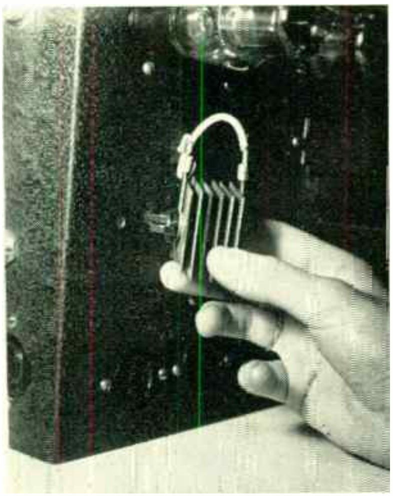
**Please address inquiries to:**

**W. C. WALKER**
*Engineering Employment Manager*

**Pacific Division**
**Bendix Aviation Corporation**
NORTH HOLLYWOOD, CALIF.

(Continued from page 20A)

## Clip-Type Rectifier



A new snap-in type rectifier featuring a clip arrangement which does not require tools for assembly, speed assembly time, and eliminates broken studs, has just been announced by the **Radio Receptor Co., Inc.**, 240 Wythe Ave., Brooklyn, N. Y.

(Continued on page 147A)

## ENGINEERS

**ELECTRONIC**

Experience in research & development of pulse circuitry, computing components and servo systems.

**SYSTEMS (E. E.)**

For work in the development of airborne navigational equipment.

**MECHANICAL**

2-3 years' experience in the design of precision instruments.

## DESIGNERS

**ELECTRONIC**
**ELECTRO-MECHANICAL**
**MECHANICAL**

1. Electronic or electrical packaging. Knowledge of sub-miniaturized techniques. Work is associated with servos, amplifiers and computers.
2. Mechanical or electro-mechanical packaging of precision flight instruments.

The modern facilities and congenial atmosphere at Kollsman, designers of America's finest aircraft instruments, provide an environment conducive to truly creative work.

Please submit complete resumes to Employment Manager.

KOLLSMAN INSTRUMENTS

**kollsman INSTRUMENT CORPORATION**

80-08 45th AVENUE, ELMHURST, NEW YORK • SUBSIDIARY OF *Standard* COIL PRODUCTS CO. INC.

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## PHYSICISTS and ELECTRICAL ENGINEERS for APPLIED RESEARCH

Expanding research program is creating new opportunities for physicists and electrical engineers to work in the fields of magnetics, electrostatics, electronics, solid-state devices, electron tube research, feed back control systems, radiation, reactor technology, mathematical physics, and allied areas. Prefer men at the M.S. or Ph.D. level of training although a limited number of positions are also available at the B.S. level. These are career-type openings offering excellent professional advancement and promotional opportunities. All applications handled promptly and confidentially. Please write today to

**The Personnel Manager**
**BATTELLE MEMORIAL INSTITUTE**
**505 King Avenue, Columbus 1, Ohio**

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## EXPERIENCED ENGINEERS

### Electron Tube Circuit Applications

New England's largest tube manufacturer has several career openings in its expansion program for qualified engineers at top Group Leader level. Require minimum of 5-10 years' experience in electron tube industry.

RAYTHEON

*Please send complete resume to Technical Placement Office*
**RAYTHEON MANUFACTURING COMPANY**
**Receiving and Cathode Ray Tube Operations**
**55 Chapel Street, Newton, Mass.**

---

## ENGINEERS

### ELECTRONIC and MECHANICAL

The Radio and Television Department of General Electric, situated in beautiful Electronics Park, is expanding its staff of development and product design engineers.

Those graduate engineers who qualify for current openings will find excellent opportunities for professional development through association with the outstanding engineers and scientists concentrating on research, development and design in all branches of the electronics industry.

Electronics Park is headquarters for the Electronics Division of GE, including the Electronics Laboratory, Radio and Television Department, Semi-conductor Products, Communications Equipment, Broadcast Equipment, Cathode Ray Tube Department, Components Department and Government Equipment Department.

*Salary scales for engineers are strictly competitive, and based on individual ability and experience. And, in addition to its comprehensive system of benefits, General Electric is noted for its stability.*

*Current openings include:*

**ELECTRONICS ENGINEERS**
**ENGINEERING SUPERVISORS**
**MECHANICAL ENGINEERS**

VHF and UHF Head-End Design
Audio and High-Fidelity Products
Advanced Development
Deflection Component Design
Deflection Systems
Color Television
Transistor Circuits
Metal-forming and Plastics

*Please send complete resume to:*
*MR. JAMES STARK*

**GENERAL ELECTRIC**

*Electronics Park*
*Syracuse, N.Y.*

**News-New Products**

*(Continued from page 145A)*

The new device, known as a Qui-klip, was developed in conjunction with the Tinnerman Products Corporation of Cleveland, manufacturers of speed-type fasteners. It does not require special sockets for mounting, only needing, according to Radio Receptor, two round holes to be snapped into place. In addition, solderless connecters are available for making electrical contact to the rectifier.

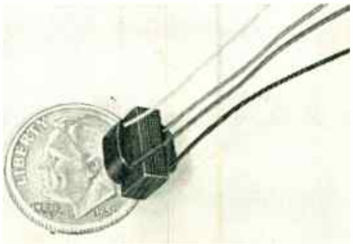
## Miniature Transformer

A new "Miniformer" miniature transformer, reduced in size and weight but of superior efficiency, has been introduced by **Gramer-Halldorson Transformer Corp.,** 2734 N. Pulaski Rd., Chicago 39, Ill. Designed primarily to meet the smaller space requirements of hearing aid components, the new "Miniformer" has additional applications where space and weight factors are of prime importance such as: computers, pocket radios, FM transceivers, telephone recorders, air borne equipment, and so forth. This unit measures $\frac{1}{4}\times\frac{5}{16}\times\frac{3}{8}$ inch and weighs 0.004 pound. The 100W72 interstage transformer, illustrated, has a match impedance of: primary, 20,000 ohms; secondary, 1,000 ohms; resistance: primary, 1030 ohms; secondary, 167 ohms. Power rating for primary inputs from 1

*(Continued on page 148A)*

**News-New Products**

*(Continued from page 147A)*



volt to 7 volts: 2.5 milliwatts. Other impedance matches are also available. Leads are color-coded, high temperature plastic insulated. For complete engineering data and a circular on this and other miniature transformers write to the company.

### 400-cps Motor

Combining good speed regulation with a very low slip, the new AC-93 miniature subfractional induction motor produced by **Dalmotor Co.**, 1360 Clay St., Santa Clara, Calif., offers favorable application characteristics for magnetic tape recorder operation; servo or actuator motor; and geophysical-equipment uses. Although the standard unit is an induction motor, its design is such that it can be supplied with minor changes, as a synchronous unit with reduced power output.

*(Continued on page 151A)*

## CIRCUIT ENGINEERS

# Why Not Work in Vacationland?

One of the many advantages in working for Sanders Associates, Inc. is the site of the plant itself: Nashua in lovely New Hampshire, New England's most beautiful state. Less than one hour from both the White and Monadnock Mountains, with cool streams, crystal lakes, lush green foliage, Nashua's many natural recreational facilities abound. Or, if you prefer the surf and the sea, you're less than one hour from world-famous Hampton and Rye Beaches.

Of distinct advantage, too, is Sanders' working environment: the small, effective engineering groups working on a variety of projects, the balance of military and commercial work, the realistic management BY engineers FOR engineers make way for quick professional growth and personal advancement.

Noteworthy "firsts" developed at Sanders include printed "strip-line" plumbing, tape resistors, the world's smallest rate gyro. On the staff are some of the top electronics experts in America, and a good many of the most promising junior men — exceptional engineers with the "something extra" that makes the difference between competence and real talent.

To complement this fine team and to permit further expansion, Sanders is adding a few engineers with at least 3 to 5 years experience in missile guidance, pulse and doppler radar, microwave antenna, airborne navigation, printed circuit and component development.

If you are an exceptional engineer — have not only talent, but ambition and drive — Sanders can provide unusual opportunity. Address inquiries to Mr. J. I. Chesterley.

SANDERS ASSOCIATES INCORPORATED 137 Canal St., Nashua, New Hampshire

---

## DEVELOPMENT ENGINEERS and PHYSICISTS

RADIO AND RADAR SYSTEM APPLICATIONS
TRANSISTORS AND MAGNETIC AMPLIFIERS
SERVOMECHANISMS AND ANALOGUE DEVICES
ELECTROMECHANICAL CONTROLS AND ACTUATORS

We are now staffing the new Electronics Laboratory of our Aeronautical Division in Anaheim, California. The selection of associates is based on consideration of their demonstrated abilities and interests in connection with our long-range laboratory plans. Careful attention to engineering and operational planning and to the selection of critical control problems assures opportunities for continuous professional development. The efforts of a relatively small but select staff are being applied on projects requiring engineering ingenuity essential to advancing the art of control in the aeronautical field.

Positions are available at all levels of laboratory work. Please direct inquiries to Vernon Vogel, Electronics Laboratory Director.

*Mr. Controls*

*Aeronautical Division*

**Robertshaw-Fulton**
CONTROLS COMPANY

SANTA ANA FREEWAY AT EUCLID AVENUE
ANAHEIM, CALIFORNIA

---

## MECHANICAL ENGINEERS

**Interested in the Electronics Field**

●

Responsible engineering positions open with a major industrial leader located in upper New York State . . . at one of the country's most advanced electronics centers.

These openings require experience in mechanical engineering of electronic equipment, or mechanical design experience and an interest in electronics.

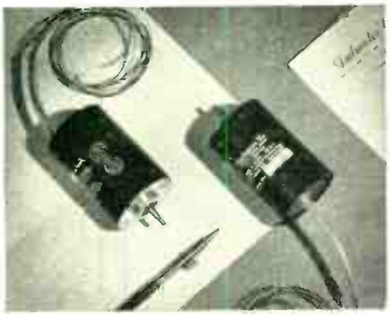
Work is in design and development, requiring no drafting board or floor work.

*To arrange for convenient interview, send resumes to:*

**Box 823**
**Institute of Radio Engineers**
**1 East 79th St.**
**New York 21, N.Y.**

## News-New Products

(Continued from page 148A)



Operating at 115 volts, 400 cps, 1 phase, the unit has an rpm of 11,800 with a 15 watt output. Locked rotor torque is 1.8 ounces/inches minimum. Maximum power input is 45 watts, and rotation is reversible.

## Professional Group Meetings

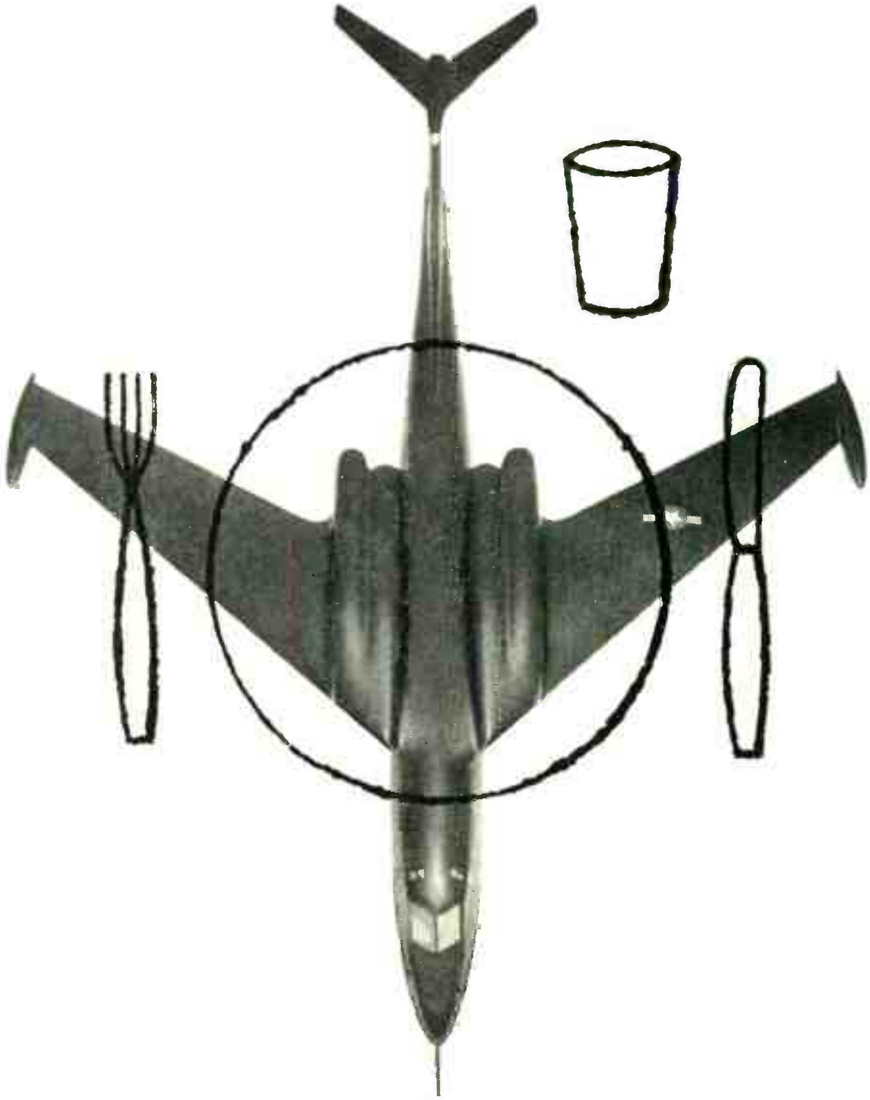
(Continued from page 104A)

Los Angeles Chapter—February 17

"How to Run Computing Installations: A Panel Discussion," moderator, Abe Faren; panel members, Don Madden; Frank Cazzone; E. M. McCormick; Fred Hollander.

(Continued on page 152A)





## Professional Group Meetings

*(Continued from page 151A)*

Los Angeles Chapter—January 20

"The Bendix Combination G.P. and Differential Analyzer" by Max Palevsky, Bendix Computer Division.

"Transistor Flip-Flops" by Chris Wanlass, Ramo-Wooldridge Corporation.

New York Chapter—February 14

"The Naval Ordnance Research Calculator (DORC)" by Byron L. Havens, Engineer in Charge of DORC Project.

New York Chapter—January 20

"Bi-Lateral Magnetic Selection Systems for Large Scale Computer Memory" by Amir H. Septhban, Monrobot Laboratories.

"High Speed Core Memory" by E. J. Otis, Air Force Cambridge Research Station.

New York Chapter—December 16

"Feasibility of an All Magnetic Computer" by Isaac Auerbach, Burroughs Research Center.

New York Chapter—November 23

"The Logical Principles Employed in Underwood Electronic Computers" by Evelyn Berezin, Underwood Corporation.

"The Type of Circuitry Employed in Underwood Electronic Computers" by Albert Auerbach, Underwood Corporation.

New York Chapter—October 26 and 27

"Project Cyclone" by Leo Bauer, Reeves Instrument Company.

### Information Theory

Albuquerque-Los Alamos Chapter—March 9

"An Example in Statistical Communication Theory" by Walter E. Brown.

### Instrumentation

Houston Chapter—March 22

"A Multi-Loop Self Balancing Power Amplifier" by J. Ross MacDonald, Texas Instruments, Incorporated.

### Medical Electronics

San Francisco Chapter—April 7

"Electronic Instrumentation in Surgery," by Bertram Feinstein, Mt. Zion Hospital.

"The New Operating Room at Mt. Zion Hospital" by Mr. Carter Collins, Consultant, Research and Development Laboratories, U. C. Medical Center.

### Microwave Theory and Techniques

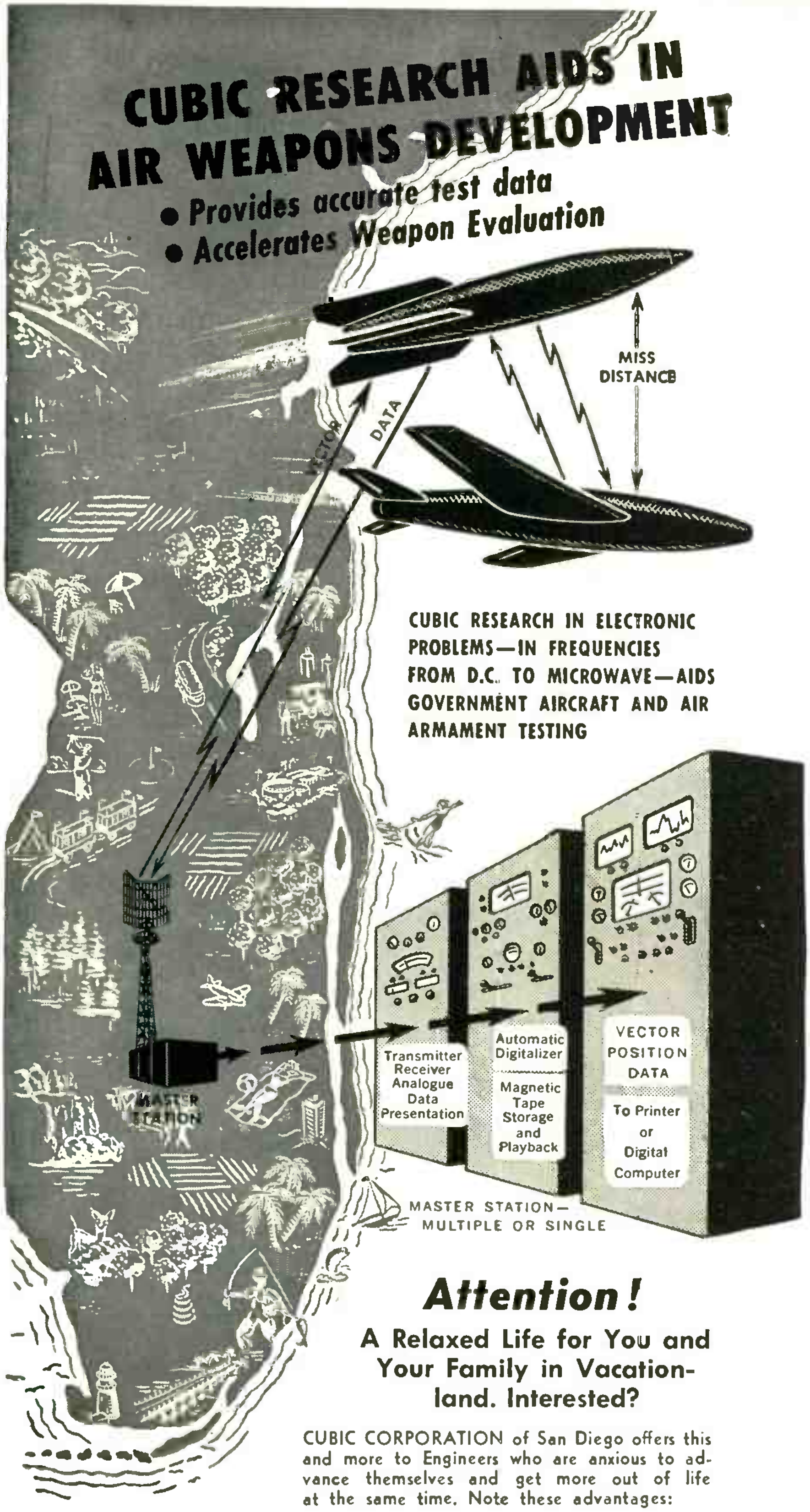
Northern New Jersey Chapter—February 16

"Microwave Applications of Ferrites" by J. H. Rowen, Bell Telephone Laboratories.

*(Continued on page 155A)*

# CUBIC RESEARCH AIDS IN AIR WEAPONS DEVELOPMENT

- Provides accurate test data
- Accelerates Weapon Evaluation




CUBIC RESEARCH IN ELECTRONIC PROBLEMS—IN FREQUENCIES FROM D.C. TO MICROWAVE—AIDS GOVERNMENT AIRCRAFT AND AIR ARMAMENT TESTING

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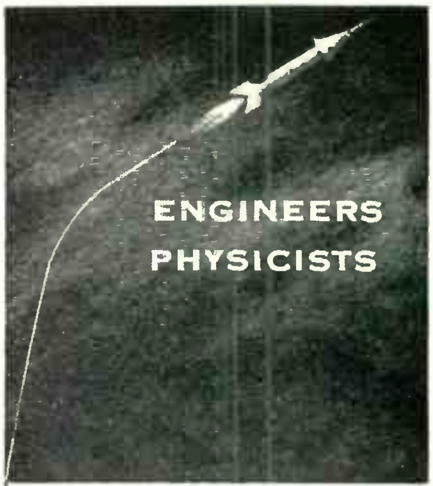


# How SYLVANIA Can Help You in the Missiles Field thru its Stability and Diversity

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INERTIAL GUIDANCE OR INFRA-RED FOR DETECTION & TRACKING
SERVO SYSTEM DESIGN & ANALYSIS
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(A suburban location just 8 miles from downtown Boston)

# PROFESSIONAL SERVICES



## Professional Group Meetings

(Continued from page 152A)

Northern New Jersey Chapter—January 19

"A Display of X-Band Impedance on an Oscilloscope" by Henry L. Bachman, Wheeler Labortories.

### Nuclear Science

Connecticut Valley Chapter—February 23

The following three films were presented: "Bikini-Radiological Laboratory," "Nuclear Reactors for Research," "Operation Ivy."

Chicago Chapter—February 18

"Trends in Reactor Development" by Lloyd V. Berkner, Associated Universities, Incorporated.

Chicago Chapter—December 17

"Radio Carbon Dating" by James R. Arnold, Institute of Nuclear Studies, University of Chicago.

Washington, D. C. Chapter—March 25

"An Accounting of the Benefits of Nuclear Energy" by Clifford K. Beck, North Carolina State College.

### Reliability & Quality Control

Washington, D. C. Chapter—April 13

"Statistical Methods for Engineers" by Leon Bass, Quality Control Div., Jet Engine Div., General Electric Company.

(Continued on page 158A)

# COMMUNICATIONS EQUIPMENT CO.

## MICROWAVE COMPONENTS

### 10 CM.—RG48/U Waveguide

**10CM ECHO BOX:** Tunable from 3200-3333 Mc. For checking out radar transmitters, for spectrum analysis, etc. Complete with pickup antenna and coupling devices ....................$17.50

**POWER SPLITTER** for use with type 726 or any 10 CM Shepherd Klystron. Energy is fed from Klystron antenna through dual pick-up system to 2 type "N" connectors ..............$12.50

**LHTR. LIGHTHOUSE ASSEMBLY.** Parts of RT39 APG 5 & APG 15. Receiver and Trans. Cavities w/assoc. Tr. Cavity and Type N CPLG. To Recvr. Uses 2C40, 2C43, 1B27. Tunable APX 2400-2700 MCS. Silver Plated ..........$15.00

**BEACON LIGHTHOUSE** cavity p/o UPN-2 Beacon 10 cm. Mfg. Bernard Rice, each ........$27.50

**MAGNETRON TO WAVEGUIDE** Coupler with 721-A Duplexer Cavity, gold plated ......$31.50

**721A TR BOX** complete with tube and tuning plungers ....................................$12.50

**McNALLY KLYSTRON CAVITIES** for 707B or 2K28. 2700-2900 Mc. ....................$4.00

**AS14A AP-10 CM** Pick up Dipole with "N" Cables ....................................$4.50

**HOLMDELL-TO-TYPE "N"** Male Adapters, W. E. #D167284 ................................$2.75

**I.F. AMP. STRIP:** 30 MC, 30 d.b. gain, 4 MC Band-width, uses 6AC7's—with video detector, A.F.C. less tubes ........................$17.50

**BEACON ANTENNA**, AS31/APN-7 in Lucite Ball. Type "N" feed ..........................$22.50

**ANTENNA**, AT49A/APR: Broadband Conical, 300-3300 MC Type "N" Feed ..........$12.50

**"E" PLANE BENDS**, 90 deg. less flanges ...$7.50

### 3 CM.—RG 52/U Waveguide

**3CM. DIPOLE FEED**, 15" L. for APS-15 ...$14.50

**MITRED ELBOW**, Cast aluminum, 1¼"x⅝" W.G. W.E. Flanges. "E" Plane ..............$3.50

**FLEX. WAVEGUIDE SECTION**, 1 ft. long. With UG-40/UG-39 flanges. Attenuation is less than 0.1 db. at 9375 mc. and VSWR is less than 1.02 ........................................$7.50

**3 CM ANTENNA ASSEMBLY:** Uses 17" parabaloid dish, operating from 24 vdc motor. Beam pattern: 5 deg. in both Azimuth and elevation. Sector Scan: over 160 deg. at 35 scans per minute Elevation Scan: over 2 deg. Tilt: Over 24 deg. .............................................$35.00

**Cross-Guide Directional Coupler**, UG-40 output flange. Main Guide is 6" Long, with 90 Deg. "E" Plane bend at one end, and is fitted with Std. UG 39/UG 40 flanges. Coupling figure: 20 db Nominal ................................$22.50

**RG52/U Waveguide** in 5' lengths, fitted with UG 39 flanges to UG40. Silver plated ...**per length $5.00**

**Rotating-Joints** supplied either with or without deck mountings. With UG40 flanges **.each. $17.50**

**Bulkhead Feed-thru Assembly** ..............$15.00

**Pressure Gauge Section** with 15 lb. gauge ...$10.00

**Directional Coupler**, UG-40/U Take off 20db $17.50

**MAGNET AND STABILIZER CAVITY** For 2J41 Magnetron ......................................$24.50

**Rotary joint** choke to choke with deck mounting ..............................................$17.50

**90 degree elbows**, "E" plane 2½" radius ..$8.50

**ADAPTER**, waveguide to type "N", UG 81-U, p/o TS 12, TS 13, Etc. ......................$7.50

**ADAPTER, UG-163/U** round cover to special BTL. Flange for TS 45, etc. ..............**$2.50 ea.**

## MICROWAVE ANTENNAS

**3 cm. Horn**, 1" x ½", with twist and 180 deg. bend. With dielectric window ..........**As shown $22.50**

**AT49/APR**—Broadband Conical, 300-3300 MC, Type N Feed ......$8.95

**Discone Antenna**, AS 125 APR, 1000-3200 mc. Stub supported with type "N" Connector ........$14.50

**AS14A/AP**, 10 CM pick up dipole assy, complete w/ length of coax and "N" connectors .........$4.50

**AS46A/APG-4** Yagi Antenna, 5 element array ..$22.50

**30' Parabolic** Reflector Spun Aluminum dish .....$4.85

**AN/APA-12**—Sector Scan adaptor for APS-2 radar—Complete Kit ..................................$37.50

**TPS-3**, 10 Ft. Dish, "Chicken Wire" Parabola. Extremely lightweight, portable ..................$125.00

**AN-154** 3 vertical dipoles working against a rectangular mesh approx. 3'x4'. Freq. 140-200 mc. with lobing switch (115v. 60 cy) and portable slatted crate. Extremely rugged ......................$27.95

**LP-24** Alford loop, for use with glide-path transmitters (MRN-1, etc.) 100-108 mc. ......................$32.50

## PULSE MODULATORS

**MIT. MOD. 3 HARD TUBE PULSER:** Output Pulse Power 144 KW (12 KV at 12 Amp). Duty Ratio .001 max. Pulse duration: .5, 1.0, 2.0 microsec. Input voltage: 115 v. 400 to 2400 cps. Uses: 1-715B, 4-829-B, 3-72's, 1-73. New ..............Less Cover—$135

**ASD Modulator Units**, mfd. by Sperry. Hard tube pulser delivers Pk. pulse of 144 kw. Similar to Mod. 3 unit. Brand new, less tubes ................$85.00

**Airborne RF head**, model ALA, delivers 50 Kw peak output at 9000 mc. at .001 duty. Complete with pulser unit and all tubes. Used, excel. .....$185.00

## VACUUM TUBES

| | | | | | |
|---|---|---|---|---|---|
| 1N21 | $ .28 | 5J33 | $ 4.75 | 700D | $ 8.75 |
| 1P5GT | .45 | 5J30 | 14.50 | 703A | 1.50 |
| 1P30 | 1.10 | C5B/5C30 | 1.10 | 704A | .95 |
| 2C21 | .35 | C6J/5C21 | 4.50 | 705A | .75 |
| 2C22/7193 | .07 | 7C4/1203A | .18 | 706A | 9.75 |
| 2C26A | .08 | 9GP7 | 3.45 | 706D | 14.75 |
| 2J21A | 2.25 | 10Y | .10 | 706EY | 9.75 |
| 2J22 | 2.50 | 15R | .15 | 706CY | 9.75 |
| 2J26 | 2.50 | 39/44 | .08 | 708A | 2.10 |
| 2J27 | 3.00 | QK59 | 39.50 | 709A | 2.45 |
| 2J29 | 18.50 | QK60 | 35.00 | 713A | .85 |
| 2J31 | 13.95 | QK61 | 50.00 | C-722A | .90 |
| 2J32 | 14.50 | QK62 | 54.00 | 725A | write |
| 2J37 | 9.50 | CEQ72 | .59 | 730A | 8.50 |
| 2J38 | 9.00 | ML-100 | 69.50 | 800 | .65 |
| 2J39 | 8.25 | HY114B | .25 | 801 | .25 |
| 2J48 | 22.50 | 227A | 2.50 | 837 | .85 |
| 2J56 | 48.50 | 268A | 2.25 | 843 | .19 |
| 2J62 | 6.75 | 316A | .50 | 861 | 12.00 |
| 3EP1 | 1.75 | 355A | 12.50 | 864 | .19 |
| 3FP7 | 1.10 | 356B | 10.50 | 876 | .75 |
| 4J34 | 23.50 | 393A | 4.50 | 884 | 1.10 |
| 4J38 | 85.00 | 417A | 6.10 | CK1005 | .35 |
| 4J42 | 47.50 | GL471A | 2.10 | 1625 | .20 |
| 5FP7 | 1.10 | WL531 | 2.75 | 1619 | .15 |
| 5GP1 | 4.50 | 532/1B32 | 1.10 | 1626 | .10 |
| 5HP4 | 3.50 | GL559 | .75 | 1629 | .10 |
| 5J23 | 34.50 | 700B | 8.75 | 8012 | 1:75 |

## POWER TRANSFORMERS

### COMBINATION—115V/60~INPUT

| | | |
|---|---|---|
| CT-133 | 150-C-150V/65MA, 6.3V/2.5A, 6.3V/0.6A | $1.79 |
| CT-127 | 900V/25MA PK. 5V/2A. 2V/7.5A .... | 2.79 |
| CT-006 | 350-0-350V/120MA. 5VCT/3A. 2.5VCT/12.5A. 2.5VCT/3.5A ................. | 4.39 |
| CT-965 | 78V/0.6A, 6.3V/2A ........................ | 1.95 |
| CT-004 | 350-0-350V/90MA. 5VCT/3A. 2.5VCT/12.5A ........................................... | 4.60 |
| CT-002 | 350-0350V/50MA. 5VCT/2A. 2.5VCT/7.5A ............................................. | 3.65 |
| CT-479 | 7000V/.018V, 2.5V/5A/17,800 V. Test .. | 22.50 |
| CT-013 | 450-0450V @ 200MA. 10V/1.5A, 2.5. 3.5A. 5V/3A ................................ | 4.35 |
| CT-403 | 350VCT .026A 5V/3A ..................... | 2.75 |
| CT-931 | 585VCT .086A 5V/3A, 6.3V/6A ...... | 4.25 |
| CT-929 | 4200V/.001A, 2.5V/2A. 6.3VCT/.6A ... | 5.35 |

### PLATE—115V/60~INPUT

| | | |
|---|---|---|
| PT-07 | 400VCT/4.0 AMPS For RA43 ....... | 17.50 |
| PT 034 | 125V/45MA (For Preamp) ........... | 1.15 |
| PT 157 | 660-0-660VAC (500VDC) or 550-0-550 VAC (400VDC) at 250 MADC ...... | 8.70 |
| PT 167 | 1400-0-1400 VAC (300MADC) or 1175-0-1175 VAC (1000VDC) at 300 MADC | 22.50 |
| PT 168 | 2100-0-2100 VAC (1750VDC) or 1800-0-1800 VAC (1500 VDC) at 300 MADC | 33.00 |
| PT 371 | 210-0-210V at 2.12 Amp. ........... | 9.45 |
| PT 133 | 3140/1570V. 2.36KVA ............... | 85.00 |
| PT 801 | 22,000V/234 MA.. 5.35 KVA ........ | 115.00 |
| PT 521 | 7500V/.06A. Half Wave ............ | 59.50 |
| PT 913 | 2500V/12 MA H'SLD ................. | 4.95 |
| PT 12A | 280VCT/1.2A ............................ | 3.95 |
| PT-38-2 | 37.5/40V AT 750 MA .................. | 2.15 |

### FILAMENT—115V/60~INPUT

| | | |
|---|---|---|
| FT-140 | 5VCT @ 10A 25KV Test ............ | 17.50 |
| FT-157 | 4V/16A, 2.5V/2.75A ................. | 2.95 |
| FT-101 | 6V/.25A ........................................ | .79 |
| FT-924 | 5.25A/21A, 2x7.75V/6.5A ............ | 14.95 |
| FT-824 | 2x26V/2.5A, 16V/1A, 1.2V/7A, 6.4V/10A 6.4V/2A ........................................ | 8.95 |
| FT-463 | 6.3VCT/1A, 5VCT/3A, 5VCT/3A ...... | 5.49 |
| FT-55-2 | 7.2V/21.5A, 6.5V/6.85A, 5V/6A, 5V/3A | 8.95 |
| FT-38A | 6.3V/2.5A, 2x2.5V/7A 5KV Test ..... | 2.79 |
| FT-650 | 2.5V/10A-3KV TEST LD-CAP ......... | 7.50 |
| FT-025 | 2.5VCT/10A, 10KV TEST ............ | 6.95 |

## FILTER CHOKES

| Stock | Description | Price |
|---|---|---|
| CH-914 | 12HY/250 MA 2500 V. Test ....... | $3.25 |
| CH-CEC 117 | 9-60H/.05-400 MA. 10 KV Test .... | $14.95 |
| CH-113 | 2.5H/700 MA. 2.5 KV Test ........ | 5.75 |
| CG-044 | 8.5H/350 MA. 3.5 KV Test ........ | 6.35 |
| CH-291 | 0.1H/12 A. DCR: 0.3 Ohms ......... | 12.50 |
| CH-322 | .35H/350 MA—10 Ohms DCR ......... | 2.75 |
| CH-141 | Dual 7H/75 MA. 11H/60 MA ......... | 4.69 |
| CH-69-1 | Dual 120H/17 MA ..................... | 2.35 |
| CH-8-35 | 2x.5H/380 MA/25 Ohms ............. | 1.79 |
| CH-776 | 1.28H/130 MA/75 ohms .............. | 2.25 |
| CH-344 | 1.5H/145MA/1200V Test ............. | 2.35 |
| CH-43A | 10HY/15 MA—850 ohms DCR ......... | 1.75 |
| CH-366 | 20H/300 MA .............................. | 6.95 |
| CH-999 | 15HY/15 MA—400 ohms DCR ......... | 1.95 |
| CH-445 | 0.5HY/200 MA. 32.2 ohms. 3000 V.T. | 1.39 |
| CH-170 | 2x0.5H/380 MA. 25 ohms ............ | 2.79 |
| CH-533 | 13.5H. 1.0 AMP DC. 13.5 KVINS .. | 29.95 |

## I. F. AMPLIFIER STRIPS

**Model 15:** 30 Mc center frequency. Bandwidth 2.5 Mc. gain figure: 65 db. Uses 5 stages of 6AC7's. Has D. C. Restorer and Video Detector. A.F.C. Strip included. Input impedance: 50 ohms. Less tubes .......$17.50

**60 MC.** Miniature IF strip, using 6AK5's 60 Mc center Freq. Gain: 95 db at Bandwidth of 2.7 Mc. New. Complete with tubes. ..........................$15.00

## SPECIALS

**500 MC. RECEIVER**, Type ASB, Uses lighthouse cavity-tuned RF Amp. and Mixer, plus GL-446 oscillator. I.F. is 60mc. New, less tubes .....................$7.50

**CAPACITOR**, Split Stator, 30-200 MMF, per section, 6 KV, working, air space—0.168", Mfg. Cardwell ..........$14.50

**NF1-1 Noise Filters**, Rated 100A./35 volts DC. AAF #3300-399857005, Mallory ...........................79¢

**POWER SUPPLY** and modulator, MP 28, For TA-12 transmitter, Dynamotor rated: Input: 28 VDC/14.8A. Output 540V/450 MA. New ................$21.50

**TU-6** Tuning Units for BC 375 transmitter. Brand New, complete with case ......................$3.95

**RANGE CALIBRATOR**, Sickles, Mod III, For accurate checking of Sweeps, Markers, and Range Scopes $185.

**RA-58** Power Supply, Input: 115 VAC, 60 cy. 1 phase. Output: 10 KV 35 MA. Ideal as Hi-Pot Supply ..$85.

**CONVERTER:** Navy Type CAJO-211444. Input: 115 VDC, Output 28V DC/20 A or 12 V DC/40 A. New, complete with Filter, Starting Box, etc. ......$67.50

**WIRE: TYPE WD3/TT.** Similar to W-130, except cotton-braid covered. On 1-mile reels .......**$5 per reel**

**PANORAMIC ADAPTER AN/APA-10.** A combined Panoramic Adapter and Scope. Has 3 inputs for feeding in receiver I.F.'s of 455KC, 5.2Mc or 30 Mc. Designed also to be used as regular scope for testing other equipment. Has both vertical and horizontal push-pull amplifier inputs, etc. Complete with 21 tubes including 3" CR tube and instruction manual. For operation on 115V 400~ ......... **$115.00**

**TEST OSCILLATOR TS-47/APR**, 40-2000+ Mc. Fundamental coverage 40-500 Mc in two ranges. Harmonics above 2000 Mc. Provides a calibrated dial accuracy ± 0.7 per cent) H.F. source for testing receiving equipment. Output 3MW or more up to 400 Mc. less on harmonics. C.W., mod. pulse or sine wave output. Operates on 115/230 60 Cy. or batteries. Part of APR countermeasures equipment. New ........................ **$165.00**

**NAVY ADF Receiver** DZ-1, made by RCA. Continuous 15 to 1750 kc in 6 bands, gear-train tuning with vernier and coarse scales, broad or sharp band-pass. CW or MCW. All controls on front panel. No headaches of mechanical or electrical control interconnections. Beautifully built with 5-gang tuning capacitor, shielded tubes and coils. 8 tubes: 3 6D6, 2—76, 2—6C6, 1—41. Complete with tubes, less power supply ........................ **$22.95**

**Hydrophone, MODEL M1-2.** A lattice of 3 crystals in a disk-like structure; 17-37 Kc. High imp. .....$27.50

**AN/CRW-3A** Remote control receiver, for operating target planes, etc. New, with soundproof mtg. box. $34.50

**TEL. REPEATER, EE 89**, complete with tubes and tech. manual ..................................$17.50

**TEL. REPEATER, EE 99**, with 12 vdc. vibrator power supply (PE 204) ............................**$49.50**

**F.T.&R. 101-A**, Two-wire applique, contains equalizing devices, and balancing circuits. Used for adapting 2-wire military circuits to 4-wire systems ..$47.50

**RADAR TRAINER**, Model 15-C, Brand new, complete with pulse generator, power supply, main unit, all cables and instruction book ....................$125

**ID-60/APA-10 PANADAPTER.** may be used with IF of 455, 5.25, or 30 mc. operates from 115V, 400 Cy. $150

**UNDERWATER MICROPHONES**, Model JR—Consists of a mosaic of 7 crystals. May be used at 17-37 Kc. omnidirectional, 50 ohm output ...............$27.50

**SA4A/APA-1 Motor Driven Coaxial Ant. Switch DPDT. Continuous Operation from 24VDC. Completely Enclosed** .........................................$24.50

**MP-22 MAST BASE Mobile Antenna Mount** ... 4.59

**SAIA/APN-1 Altitude Limit Switch for APN-1 Altimeter** ................................................ 7.95

**ALTITUDE INDICATOR for APN-1** .......... 12.50

**C-387-D Final P.A. Coil for BC610 2-3.5 MC, Variable Link** ................................................ 4.89

**RA-74 Power Supply for Super Pro** ........... 69.50

**C48/ARC-5 Control Box for ARC5** .............. 4.49

**J-22/ARC-5 Junction Box** ........................ 3.49

**T-21 Artillery Microphone complete with preamp. tube. Condenser type** ............................$24.50

## DYNAMOTORS

| | INPUT | | OUTPUT | | |
|---|---|---|---|---|---|
| TYPE | VOLTS | AMPS | VOLTS | AMPS | Price |
| 35X.059 | 19 | 3.8 | 405 | .095 | $4.35 |
| PDSX-15 | 14 | 2.8 | 220 | .08 | 8.95 |
| DA-7A | 28 | 27 | 1100 | .400 | 15.00 |
| DM 33A | 28 | 7 | 540 | .250 | 3.95 |
| 23350 | 27 | 1.75 | 285 | .075 | 3.95 |
| B-19 | 12 | 9.4 | 275 | .110 | 6.95 |
| | | | 500 | .050 | |
| DA-3A* | 28 | 10 | 300 | .260 | 6.95 |
| | | | 150 | .010 | |
| | | | 14.5 | 5. | |
| PE 73 CM | 28 | 19 | 1000 | .350 | 17.50 |
| BD 69‡ | 14 | 2.8 | 220 | .08 | 8.95 |
| DAG-33A | 18 | 3.2 | 450 | .06 | 2.50 |
| DM 25† | 12 | 2.3 | 250 | .05 | 6.95 |
| BDAR 93 | 28 | 3.25 | 375 | .150 | 6.95 |

‡ Less Filter. * Replacement for PE 94.
† Used. Excellent.

PE 94-, Brand New .........................5.95

## INVERTERS

**800-1B Input 24 vdc. 62 A. Output: 115 V. 800 cy. 7A. 1 phase. Used. excellent** ....................$18.75

**PE-218H: Input: 25/28 vdc. 92 amp. Output: 115V 350/500 cy 1500 Volt-ampere. NEW** ..............$32.50

**MAIL ORDERS PROMPTLY FILLED. ALL PRICES F.O.B. NEW YORK CITY. SEND M.O. OR CHECK. ONLY SHIPPING SENT C.O.D. RATED CONCERNS SEND P.O. PARCELS IN EXCESS OF 20 POUNDS WILL BE SHIPPED VIA CHEAPEST TRUCK OR RAILEX.**

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# INDEX AND DISPLAY ADVERTISERS



## DISPLAY ADVERTISERS

## DISPLAY ADVERTISERS



### Professional Group Meetings

*(Continued from page 155A)*

VEHICULAR COMMUNICATIONS

Detroit Chapter—April 20
"The Design of Receivers for Split Channel Operation" by R. L. Casselberry, General Electric Company.

Houston Chapter—April 26
"Progress in Mobile Communications Equipment Design" by J. A. McCormick, Mobile Communications Equipment, General Electric Company, Syracuse.

## DISPLAY ADVERTISERS

## DISPLAY ADVERTISERS


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In harness racing as in capacitors you pick the leader by looking at the record. That's why at Cornell-Dubilier, we're mighty proud of our record of new capacitor designs, consistent dependability and outstanding field performance —a record no other company can even come close to approaching. That's the record of

## C·D...45 YEARS OF FAMOUS FIRSTS

Typical of these "Famous Firsts" are the examples shown here... just three of the hundreds of money-saving answers in capacitors a C-D engineer can show you. Write to Cornell-Dubilier Electric Corp., Dept. M75, South Plainfield, N. J.

**CONSISTENTLY DEPENDABLE CORNELL-DUBILIER CAPACITORS** ®

PLANTS IN SO. PLAINFIELD, N. J.; NEW BEDFORD, WORCESTER AND CAMBRIDGE, MASS.; PROVIDENCE AND HOPE VALLEY, R. I.; INDIANAPOLIS, IND.; SANFORD AND FUQUAY SPRINGS, N. C.; SUBSIDIARY, RADIART CORP., CLEVELAND, OHIO.

THERE ARE MORE C-D CAPACITORS IN USE TODAY THAN ANY OTHER MAKE

# NEW!

# R-F Bridge

## 400 Kc to 60 Mc

The Type 1606-A Radio-Frequency Bridge is a new, improved model of the popular Type 916-A R-F Bridge which has served the communications field for well over a decade.

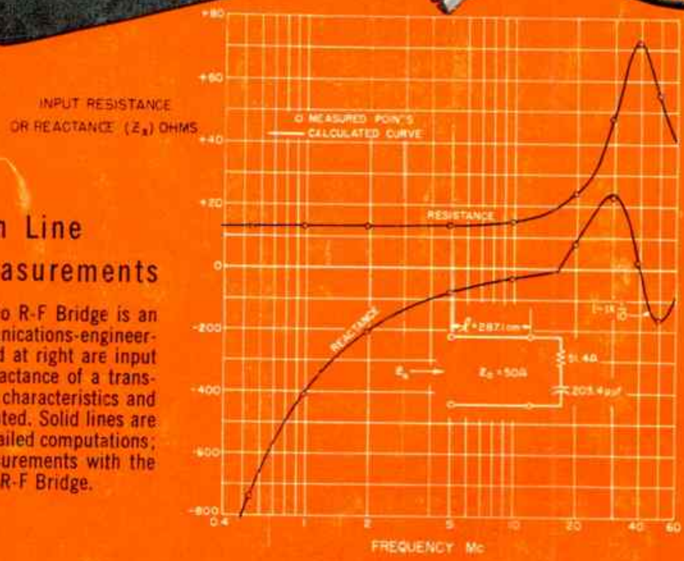
This G-R instrument reads both resistive and reactive components of impedance directly on separate dials. All variable elements are precision capacitors making possible highest accuracies in measurements, particularly at the upper radio frequencies. For direct impedance measurements of antennas, transmission lines, circuit elements and other low impedances, this versatile r-f tool is invaluable. With an external parallel capacitor, tuned circuits and other high impedances also can be measured.

The new General Radio R-F Bridge has greater sensitivity, uses new variable capacitors whose rotors and stators are milled from solid blocks of aluminum for minimum loss, is less than one-half the volume of its predecessor and contains only one bridge transformer for coverage of the whole frequency range. Dial locks are provided to prevent accidental movement of the initial balance control, and a simplified system of connection to the unknown is made available.

Those who have worked with the first G-R Radio-Frequency Bridge will find the new model an even more convenient and useful instrument. New materials, techniques and operating improvements, many suggested by customers, make this the finest R-F Impedance Bridge yet made available.


### Transmission Line Measurements

The General Radio R-F Bridge is an invaluable communications-engineering tool. Indicated at right are input resistance and reactance of a transmission line with characteristics and termination indicated. Solid lines are obtained from detailed computations; circles from measurements with the new Type 1606-A R-F Bridge.




NEW!

**Type 1606-A Radio-Frequency Bridge** .......................**$535**

**Type 1606-P1** Luggage-type Carrying Case . . . facilitates transportation and insures safe storage.......................**$15**

**Frequency Range:** 400 kc to 60 Mc

**Resistance Range:** 0 to 1000 ohms

**Reactance Range:** ±5000 ohms direct reading at 1 Mc — range varies inversely as frequency

**Basic Accuracy:** ±(1% + 0.1 ohm) for Resistance and ±(2% + 1 ohm) for Reactance; slightly less accurate at higher frequencies — high-frequency corrections provided

**Accessories Supplied:** Coaxial cables for connecting generator and detector, two leads of different lengths for connecting unknown impedance to bridge, Type 874-PB58 Panel Connector

**Dimensions:** 12½ x 9½ x 10¼ inches **Net Weight:** 23 lbs (29 lbs with carrying case)

**GENERAL RADIO Company**

275 Massachusetts Avenue, Cambridge 39, Massachusetts, U. S. A.

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in Electronics

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