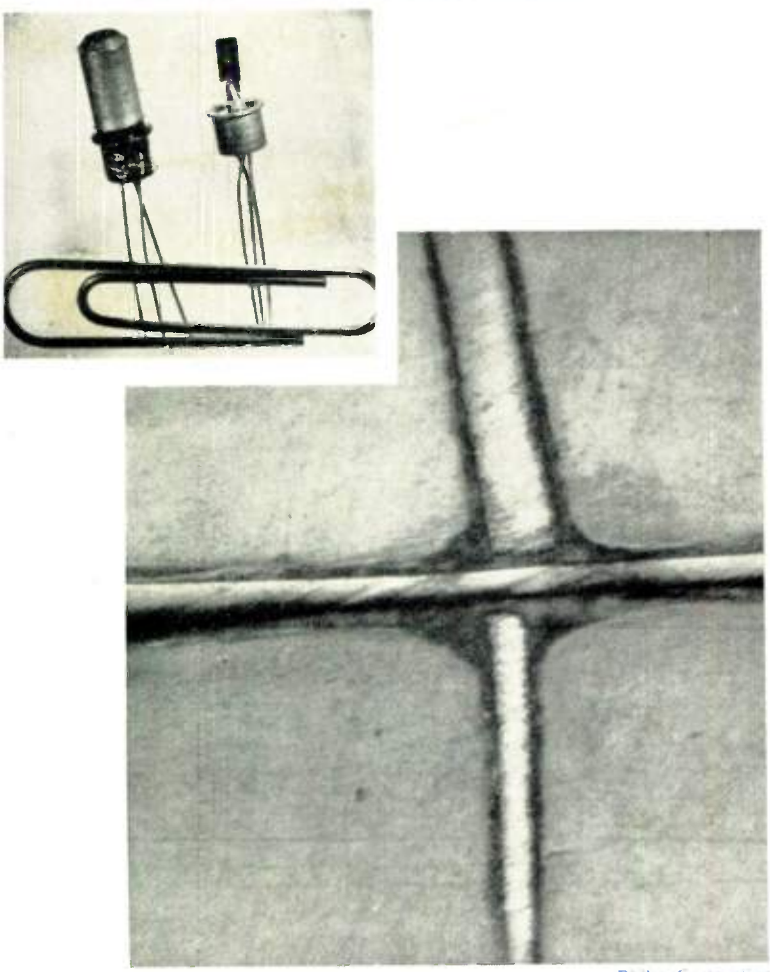
JUNE · 1955

Proceedings OF THE IRE

SILICON SURFACE-BARRIER TRANSISTOR



Philco Corporation

This magnified cutaway view shows a silicon wafer that has been electrolytically etched to a thickness of 0.5 mils. The top and bottom leads terminate at the scarcely visible electroplated emitter and collector contacts which have been deposited on each side of the wafer. The inset at top shows the completed transistor unit with and without the container top.

Volume 43 Number 6

IN THIS ISSUE

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IRE Standards on Television: Definitions of Color Terms, 1955

The Institute of Radio Engineers

LARGEST PRODUCERS IN THIS FIELD FOR TWO DECADES...

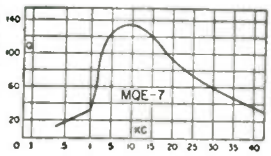
HIGH Q INDUCTORS

FOR EVERY APPLICATION

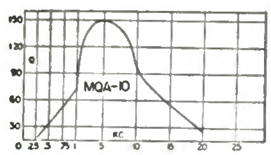
FROM STOCK... ITEMS BELOW AND 650 OTHERS IN OUR CATALOGUE B.

MQ Series Compact Hermetic Toroid Inductors

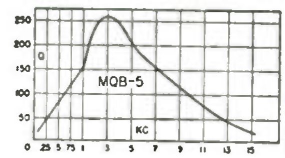
The MQ permalloy dust toroids combine the highest Q in their class with minimum size. Stability is excellent under varying voltage, temperature, frequency and vibration conditions. High permeability case plus uniform winding affords shielding of approximately 80 db.



MQE
15 stock values from 7 Mhy. to 2.8 Hy.



MQA
19 stock values from 7 Mhy. to 22 Hy.



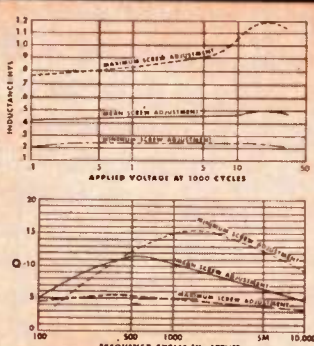
MQB
12 stock values from 10 Mhy. to 25 Hy.

MQ drawn case structure.

	Length	Width	Height
MQE	1/2	1-1/16	1-7/32
MQA	11/16	1-9/32	1-23/32
MQB	1-5/16	2-9/16	2-13/16

VIC case structure

Length	Width	Height
1-1/4	1-11/32	1-7/16



Type	Mean Hys.	Type	Mean Hys.
VIC-1	.0085	VIC-12	1.3
VIC-2	.013	VIC-13	2.2
VIC-3	.021	VIC-14	3.4
VIC-4	.034	VIC-15	5.4
VIC-5	.053	VIC-16	8.5
VIC-6	.084	VIC-17	13.
VIC-7	.13	VIC-18	21.
VIC-8	.21	VIC-19	33.
VIC-9	.34	VIC-20	52.
VIC-10	.54	VIC-21	83.
VIC-11	.85	VIC-22	130.

VIC Variable Inductors

The VIC Inductors have represented an ideal solution to the problem of tuned audio circuits. A set screw in the side of the case permits adjustment of the inductance from +85% to —45% of the mean value. Setting is positive.

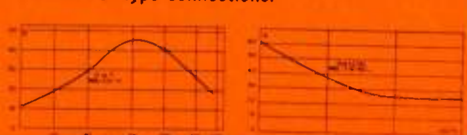
Curves shown indicate effective Q and L with varying frequency and applied AC voltage.

MQL Low Frequency High Q Coils

The MQL series of high Q coils employ special laminated Hipermalloy cores to provide very high Q at low frequencies with exceptional stability for changes of voltage, frequency, and temperature. Two identical windings permit series, parallel, or transformer type connections.

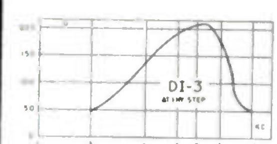
MQL-1	2.5/10 Hys.
MQL-2	5/20 Hys.
MQL-3	50/200 Hys.
MQL-4	100/400 Hys.

MQL case
1-13/16 dia. X 2-1/2" H.



DI Inductance Decades

These decades set new standards of Q, stability, frequency range and convenience. Inductance values laboratory adjusted to better than 1%. Units housed in a compact die cast case with sloping panel ideal for laboratory use.

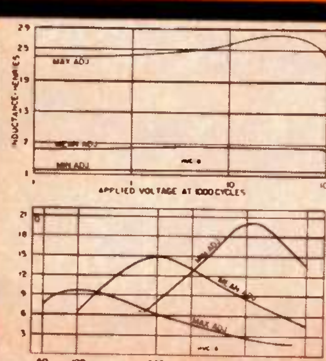


DI-1	Ten 10 Mhy. steps.
DI-2	Ten 100 Mhy. steps.
DI-3	Ten 1 Hy. steps.
DI-4	Ten 10 Hy. steps.

DI DECADE
Length 4½"
Width 4⅜"
Height 2⅜"

HVC Hermetic Variable Inductors

A step forward from our long established VIC series. Hermetically sealed to MIL-T-27 . . . extremely compact . . . wider inductance range . . . higher Q . . . lower and higher frequencies . . . superior voltage and temperature stability.



Type No.	Min. Hys.	Mean Hys.	Max. Hys.
HVC-1	.002	.006	.02
HVC-2	.005	.015	.05
HVC-3	.011	.040	.11
HVC-4	.03	.1	.3
HVC-5	.07	.25	.7
HVC-6	.2	.6	2
HVC-7	.5	1.5	5
HVC-8	1.1	4.0	11
HVC-9	3.0	10	30
HVC-10	7.0	25	70
HVC-11	20	60	200
HVC-12	50	150	500

HVC case structure.

Width	Length	Height
25/32	1-1/8	1-7/32

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Send your specifications for prices.

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150 Varick Street, New York 13, N. Y. EXPORT DIVISION: 13 E. 40th St., New York 16, N. Y. CABLES: "ARLAB"

PROCEEDINGS OF THE I.R.E. June, 1955, Vol. 43, No. 6. 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.

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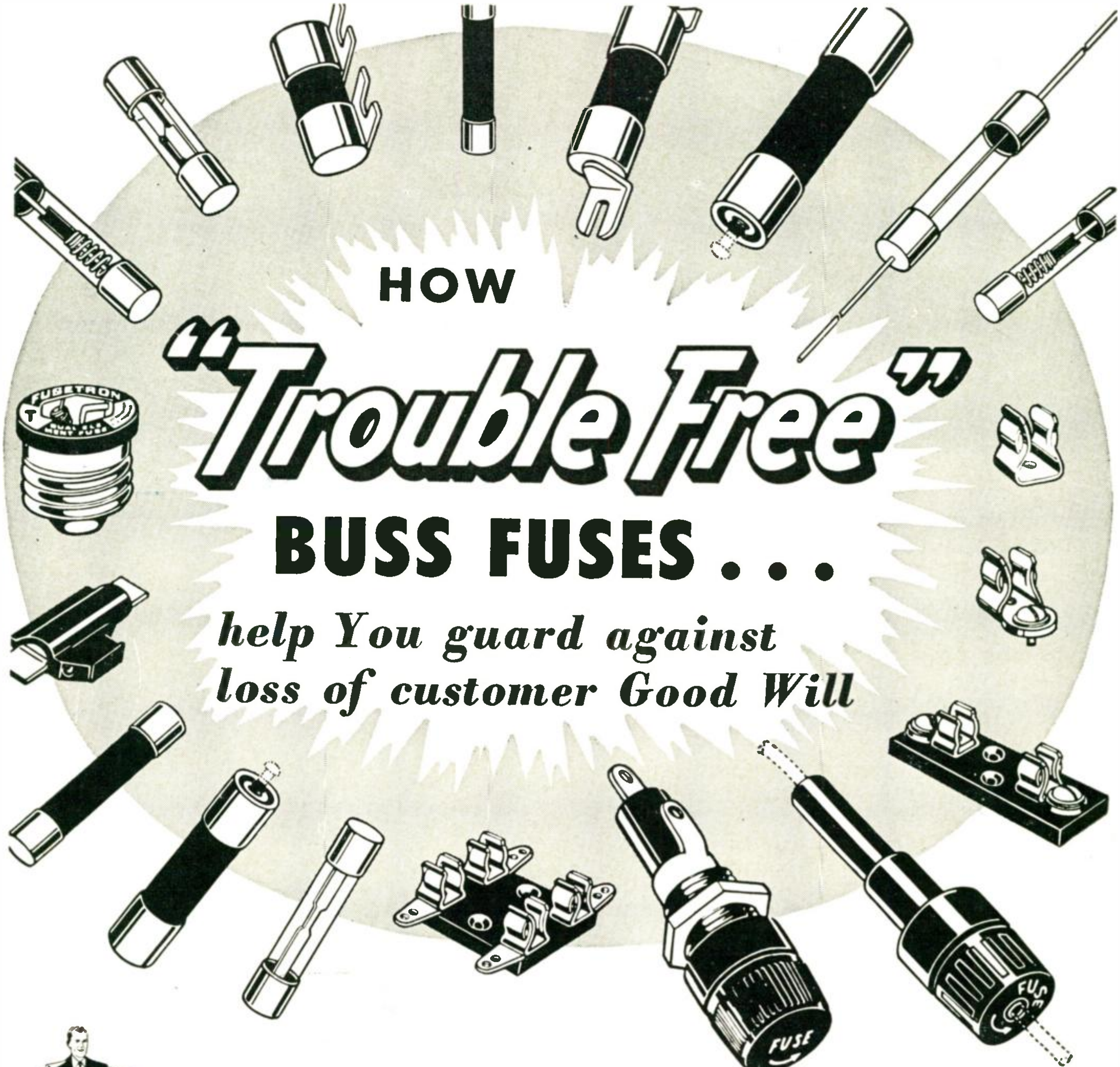

Dependable BUSS fuses can help safeguard your product or service against loss of user satisfaction because . . .

When there is trouble on the circuit, BUSS fuses open and prevent further damage to equipment—saving users the expense of replacing needlessly burned out parts.

However, BUSS fuses won't give a "false alarm" by blowing when trouble doesn't exist. Users are not annoyed with shutdowns caused by needless blows.

To make sure of unfailing dependability — every BUSS fuse, normally used by the Electronic Industries, is tested in a sensitive electronic device that rejects any fuse not correctly calibrated, properly constructed and right in all physical dimensions.

Save engineering time on new products

The BUSS fuse research laboratory and its staff of engineers are at your service to help you with problems involving electrical protection. They can help you select the right fuse for the job . . . if possible, a fuse available in local wholesalers' stocks, so that your device can easily be serviced.

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(Div. McGraw Electric Co.)
University at Jefferson St. Louis 7, Mo.

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IRE-655

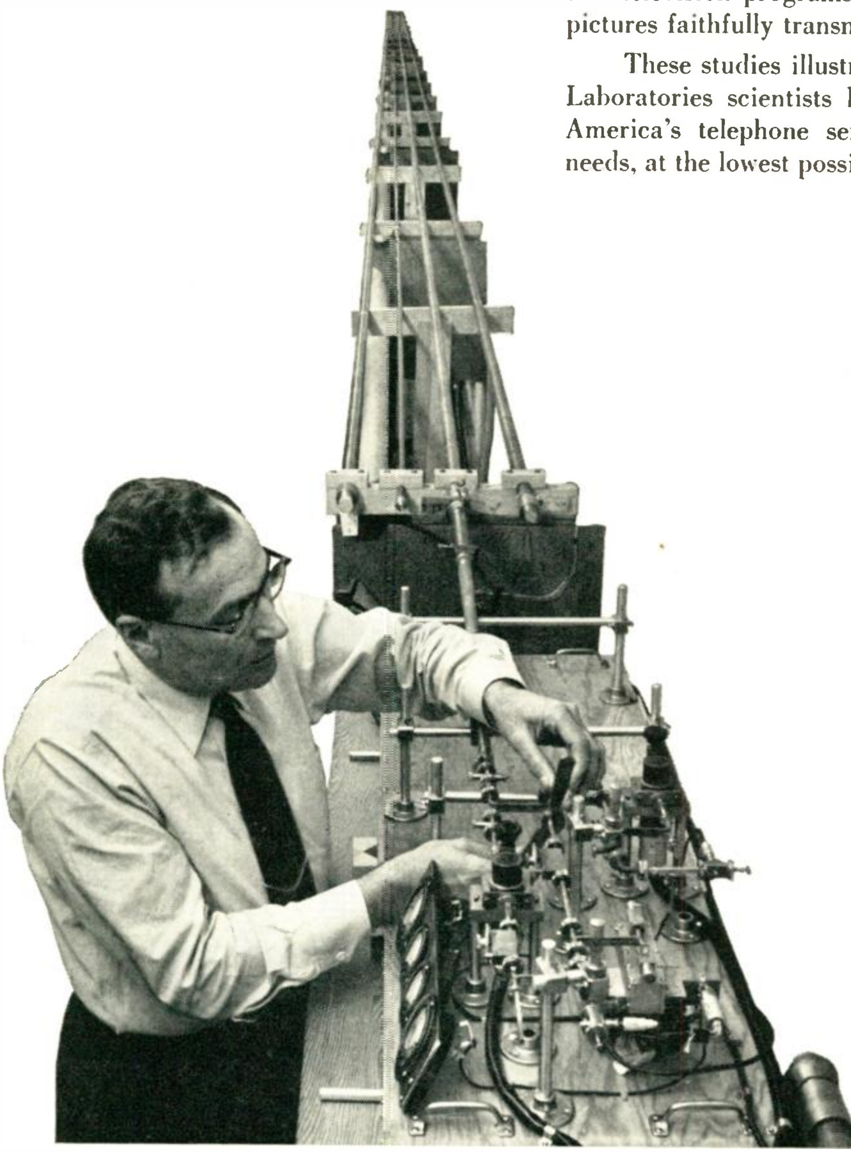
Pipes of Progress

Hundreds of thousands of telephone conversations or hundreds of television programs may one day travel together from city to city through round waveguides—hollow pipes—pioneered at Bell Telephone Laboratories.

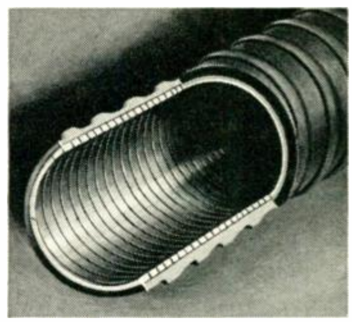
Round waveguides offer tremendous possibilities in the endless search for new ways to send many voices great distances, simultaneously, and at low cost. Today, Bell Laboratories developments such as radio relay, coaxial cable and multivoice wire circuits are ample for America's needs. But tomorrow's demands may well call for the even greater capacity of round waveguides.

Unlike wires or coaxial, these pipes have the unique property of *diminishing* power losses as frequencies rise. This means that higher frequencies can be used. As the frequency band widens, it makes room for many more voices and television programs. And the voices will be true, the pictures faithfully transmitted.

These studies illustrate once more how Bell Telephone Laboratories scientists look ahead. They make sure that America's telephone service will *always* meet America's needs, at the lowest possible cost.



Testing round waveguides at Bell Telephone Laboratories, Holmdel, New Jersey. Unlike coaxial cable, waveguides have no central conductor. Theoretically, voice-capacity is much greater than in coaxial cable.



New type of waveguide pipe formed of tightly wound insulated wire transmits better around corners than solid-wall pipes.

New type waveguide is bent on wooden forms for study of effect of curvature on transmission. The waveguide itself is here covered with a protective coating.

Bell Telephone Laboratories

Improving America's telephone service provides careers for creative men in scientific and technical fields.

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NEW EXPANDED LINE OF HERMETICALLY SEALED

COMPRESSION TYPE END SEALS

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FOR HERMETICALLY SEALING CONDENSERS, RESISTORS AND OTHER TUBULAR COMPONENTS

E-I standardization now makes it possible to offer designers and engineers the economy of standard components in a wide selection of types and sizes. These rugged compression type end seals are available in a broad range of dimensions, in either flared tube or pierced terminals, with single or multiple lead terminations. Inquiries invited.

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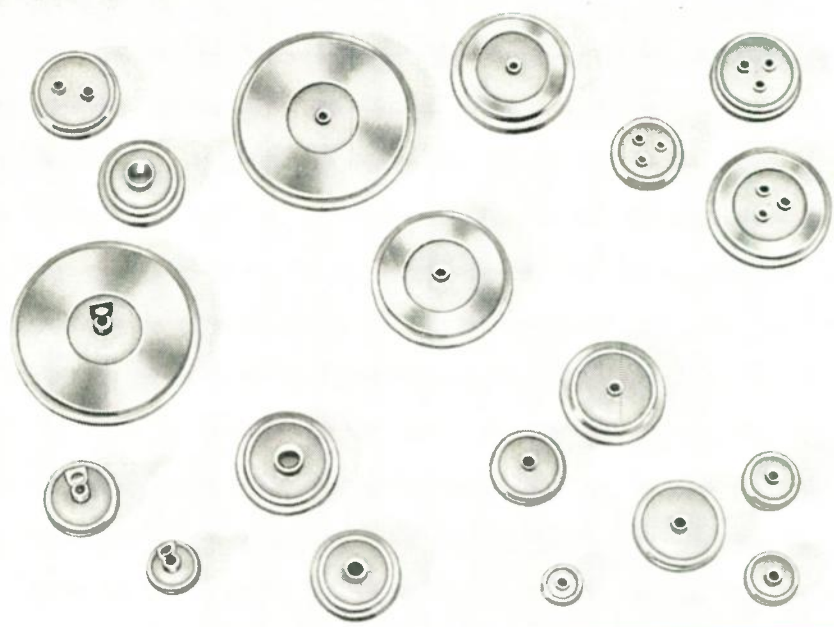
In this exclusive E-I compression construction, the glass remains under constant compression and is therefore extremely strong. These seals possess extraordinary immunity to shock, vibration and pressure changes. For all practical purposes E-I Compression Seals are indestructible. No special skill is required to apply and assembly is rapid as all metal parts are tin dipped for easy soldering.



E-I Leadership—in the field of hermetic sealing assures dependability, economy and fast delivery . . . specify E-I for multiple headers, octal plug-ins, transistor bases and closures, sealed terminals, end seals and color coded terminals.

E I ®

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Meetings with Exhibits

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

Δ

Aug. 24-26, 1955
Western Electronic Show & Convention, Civic Auditorium, San Francisco, Calif.
Exhibits: Mr. Mal Mobley, 344 N. LaBrea, 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
IRE Annual Electronic Conference, Kansas City, Mo.
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 tSatler, Boston, Mass.
Exhibits: Mr. J. D. Porter, Digital Computer Lab., Barta Building, M.I.T., Cambridge, Mass.

Nov. 28-30, 1955
IRE Data Processing Symposium, Hotel Biltmore, Atlanta, Ga.
Exhibits: Dr. B. J. Dasher, School of Electrical Engineering, Georgia Tech., 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.

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

WIDE RANGE

WIDE SWEEP-

Sweeping

OSCILLATOR

The KAY CALIBRATED *Mega-Sweep*

SG-92/U

- **Continuously Tunable Thru Video VHF and UHF Frequencies, 50KC-950MC Range**
- **Sweep Widths to 40 MC**
- **Single Dial Tuning**

Used with a standard cathode ray oscilloscope, the Kay Calibrated *Mega-Sweep* will display the response characteristic of wide band circuits over the frequency range of approximately 50 kc to 950 mc. It features a calibrated dial indication of the approximate output frequency. The center frequency of the sweeping output voltage may thus be set to an accuracy of about 10%. The calibrated *Mega-Sweep* is the ideal instrument for use in alignment of amplifiers and filters . . . also as an FM source of wide range for instructional and lab purposes.

SPECIFICATIONS

Freq. Range: 50 kc to 1000 mc.

Freq. Sweep: Sawtooth, adjustable to 40 mc. Repetition rate, 50 to 100 c/s.

RF Output: *High*, approx. 100 mv max. into open circuit. *Low*, 5 mv into open circuit.

RF Output Control: Microwave attenuator continuously variable to 26 db.

Output Waveform: Less than 5% harmonic distortion at max. output.

Meter: Provides crystal detector current for peak output.

Regulated Power Supply: 105-125 v., 50 to 60 cps. Power Input, 100 watts.

Send for Catalog 110-A

$495 f.o.b. factory

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KAY

Mega-Sweep

Widest range of the Kay line of sweeping oscillators. Provides continuous frequency coverage up through UHF-TV bands—50 kc to 1000 mc. Widely used in radar system development and in alignment and testing of TV and FM systems and components, as well as wide band IF and RF amplifiers and filters. Freq. range, 10 mc to 950 mc. *Write for Catalog 100-A.* Price, **$465** f.o.b. factory.

KAY

111-A CALIBRATED

Mega-Sweep

Higher output model calibrated *Mega-Sweep*, with zero level baseline. Higher output facilitates frequency response testing of UHF converters or tuners. Wider sweep width permits multi-channel response viewing. Zero level baseline is convenient means of measuring gain of test circuit.

SPECIFICATIONS

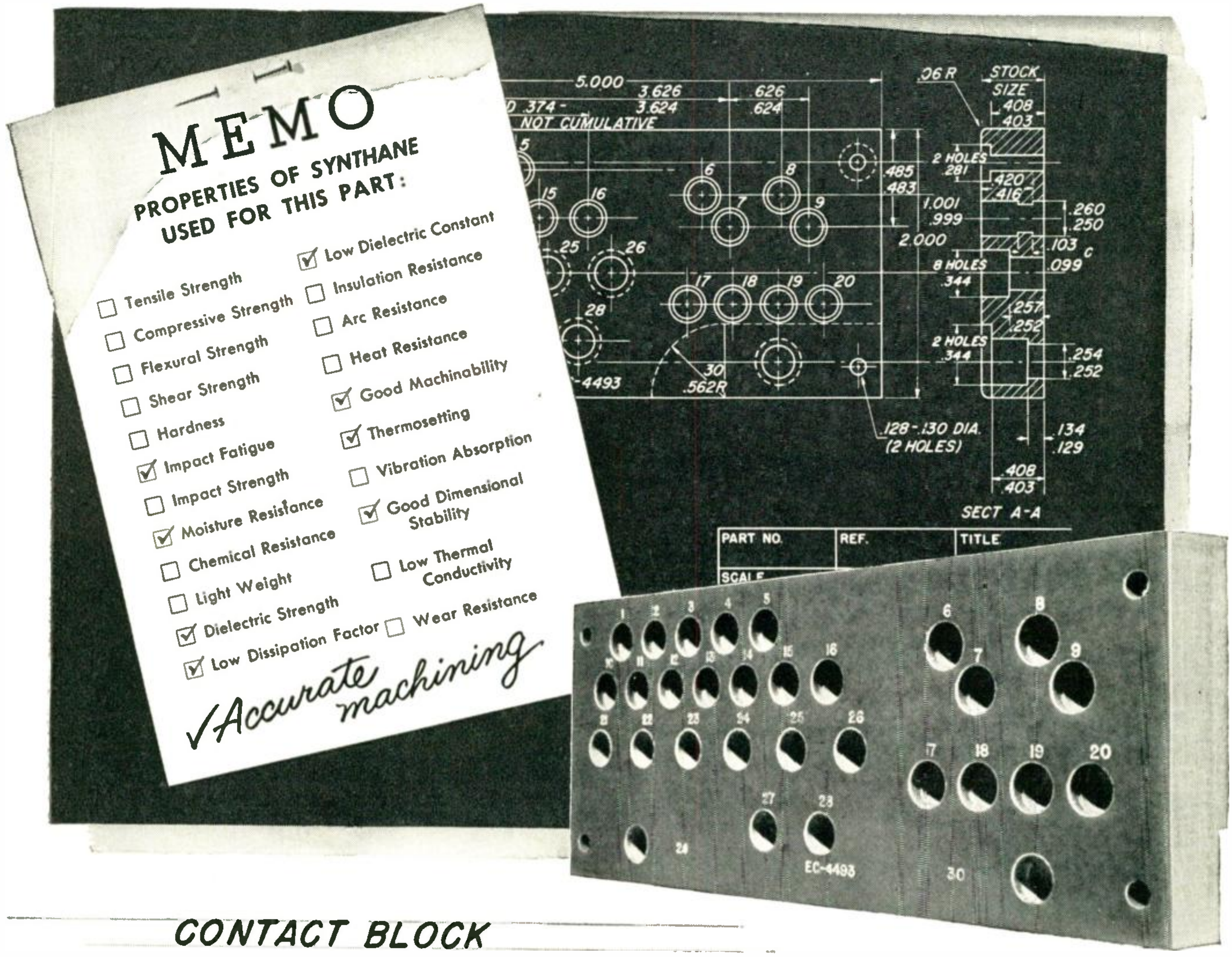
Frequency Range	Output Impedance	Output Voltage *(Into Load)*
1. 10 mc—950 mc	70 ohms unbalanced	0.15 Volts
2. 450 mc—900 mc	300 ohms balanced	0.3 Volts

Sweep Width: Continuously variable to approx. 40 mc max.

Write for Catalog 111-A Price, $575 f.o.b. factory

KAY 112-A CALIBRATED *Mega-Sweep*
Same as 111-A, except total frequency range is 800 mc to 1200 mc. *Catalog 112-A.* Price, **$575** f.o.b. factory.

YOU FURNISH THE PRINT, WE'LL FURNISH THE PART



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This contact block—for an electronic device—illustrates the rising demand for materials with many properties in combination. High dielectric strength, mechanical strength and dimensional stability are essential for the application; accurate machining is a must for proper mating of components.

The customer supplied the blueprint; Synthane Corporation did the rest—first producing the proper grade of material and then fabricating—accurately and without waste or delay.

The more than 33 grades of *Synthane* laminated plastics offer *you* a very wide range of properties in combination—physical, mechanical, electrical, and chemical. And good service and quality characterize *Synthane* fabrication. The coupon will bring you further information and technical data covering *Synthane* sheets, rods, tubes, and molded parts, and *Synthane* service.

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Using cheap capacitors with insufficient dielectric thickness is not a real saving. Take insurance against repair expenses by choosing STEAFIX silvered mica capacitors. Absolutely stable, totally damp-proof, liberally designed, STEAFIX capacitors give you years of service life without trouble and thus prove not only the best but also the most economical.

STEAFIX capacitors resist time and weather

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the european specialist of the mica capacitors

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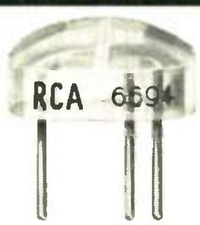
NEW MULTIPLIER PHOTOTUBE FOR GAMMA-RAY SPECTROSCOPY AND SCINTILLATION COUNTERS

RCA-6655 . . . a 10-stage, flat-face, head-on type with 1-11/16"-diameter cathode; focusing electrode for optimizing magnitude, uniformity, or speed of response; 3000 to 6500 angstroms spectral-response range (max. at approx. 4400); cathode luminous sensitivity of 50 µamp/lumen; short pulse resolving time at relatively low supply voltage of 1000 volts; and sturdy structure to withstand the rigors of field use.



NEW PHOTOCONDUCTIVE CELL FOR LIGHT-CONTROLLED RELAYS, COMPUTERS, AND LIGHT METERS

RCA-6694 . . . very tiny, cadmium-sulfide, head-on type featuring high luminous sensitivity, very low dark current, extremely low background noise, and signal output directly proportional to incident light intensity. Characteristics not substantially affected by wide temperature changes. Spectral response covers range from 3500 to 5500 angstroms with peak at about 5000. Luminous intensity sensitivity is 3 µamp/ft-c at 90 volts.

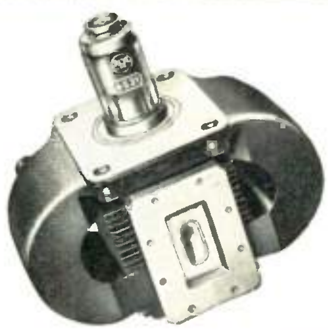
NEW HALF-WAVE VACUUM RECTIFIER FOR PULSED-RECTIFIER SERVICE

RCA-3B2 . . . a glass-octal, high-voltage rectifier of the heater-cathode type for use in the scanning systems of modern black-and-white and color-TV receivers. Rated at a maximum peak inverse plate voltage of 35,000 volts (absolute), maximum peak plate current of 80 ma., and maximum average plate current of 1.1 ma.

FOUR NEW ALLOY-JUNCTION TRANSISTORS

Hermetically sealed, germanium, p-n-p types offering extreme stability and uniformity of characteristics—initially and during life. RCA-2N104 for low-power, audio-frequency applications; RCA-2N109 for large-signal applications, such as class B audio service; RCA-2N77 and -2N105 for hearing-aids.

New RCA Transistors shown twice actual size



NEW "C" BAND MAGNETRON FOR AIRCRAFT WEATHER RADAR

RCA-6521 . . . for service as pulsed oscillator at fixed frequency of 5400 ± 20 megacycles. It is designed and conservatively rated to insure long, reliable performance. Operates with high efficiency at pulse durations up to 2.2. microseconds. Has peak input power rating of 25.6 kw, peak anode voltage rating of 16 kv, and peak anode current rating of 10 to 16 µamperes. Peak power outputs up to about 100 kw may be obtained.

RADIO CORPORATION of AMERICA
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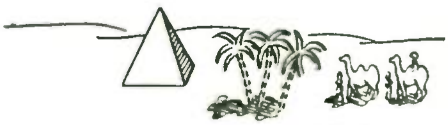
HUMID-TROL

LONG TERM STORAGE SEAL

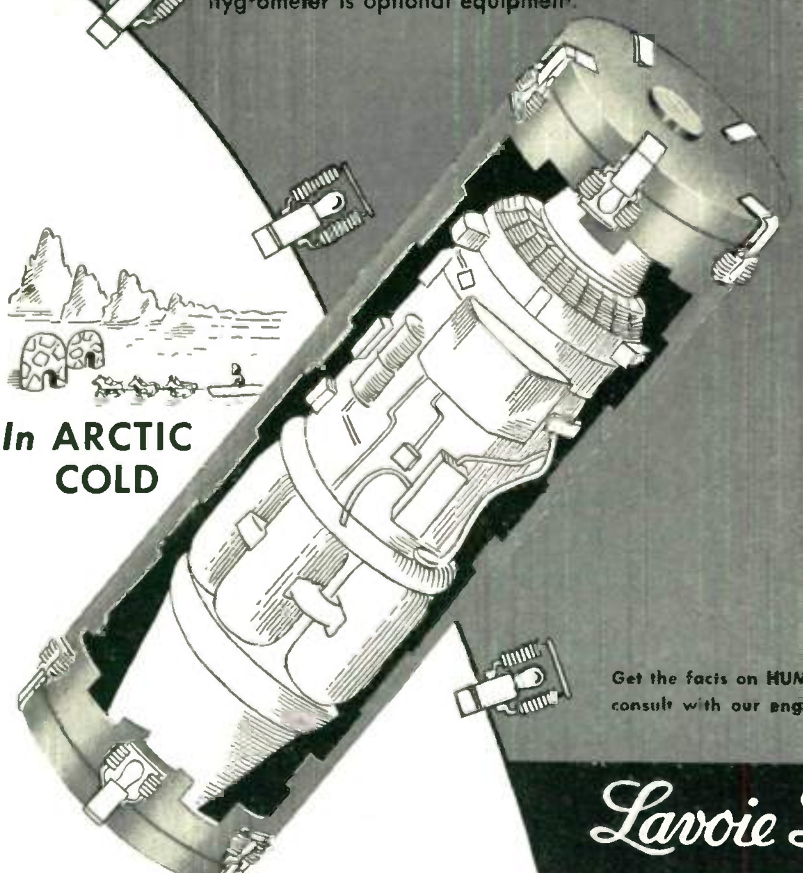
Maintains Any Desired Moisture Level
from **1%** *to* **90%** *RH* **- ALWAYS!**



In DESERT HEAT

In the search for better packaging for our own electronic devices, Lavoie research scientists have developed a combination of material and method that has given us a long term storage container in which moisture is reduced to the unprecedented low of 1%... and maintained, without variation, over long periods of time. Once attaining the desired results for our own purposes, further tests indicated a much broader latitude in which the HUMID-TROL method maintained, indefinitely, a specified degree of relative humidity for long term storage of a variety of products.

We now offer Lavoie HUMID-TROL to industry seeking long term protection against unstable humidity, dehydration, corrosion, fungi and other causes of deterioration. HUMID-TROL has no size limitations and may be used to protect small components as well as large assemblies, including engines, missiles and even heavy tanks. HUMID-TROL withstands submersion; altitude up to 100,000 feet. Re-sealable and re-usable for storage at −80°F or +180°F... withstands thermal shock and it may be customized to meet special requirements of various industries. A built-in hygrometer is optional equipment.



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RIFLES AND SMALL ARMS
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NAVIGATION EQUIPMENT
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Get the facts on HUMID-TROL; write for complete data or consult with our engineers regarding your requirements.

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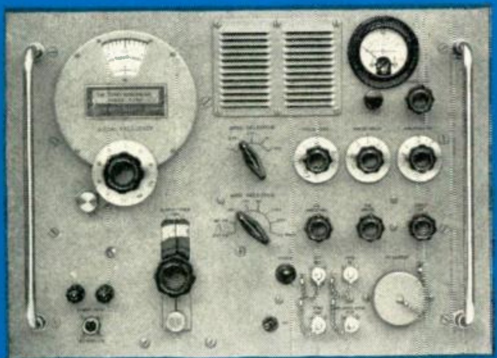
SIGNAL GENERATORS



NEW! -hp- 628A shf Signal Generator

15 to 21 KMC, 10 dbm output
SWR 1.2, high accuracy
No calibration charts
Pulse, FM, square wave modulation

Typical -hp- signal generators

-hp- 608D vhf Signal Generator

-hp- 612A uhf Signal Generator

-hp- 624C X Band Test Set

ΔRack mounted instrument available for $15.00 less.

New *-hp-* 628A is the first commercial signal generator to bring the wide range, high power, convenience and accuracy of lower-frequency signal generators to the 15 to 21 KMC range.

Operation of the instrument is typical of *-hp-* generators. Frequencies are directly set and read on one dial. No calibration charts are required. Output voltage is directly set and read. Output is 10 to 20 db better than previous spot-frequency generators. SWR is better than 1.5 at full power, and better than 1.2 at levels of +7 dbm and less. Internal pulse, FM or square wave modulation is provided, together with provision for external pulsing or FM'ing. Model 628A, $3,000.00.

Instrument	Frequency Range	Characteristics	Price
-hp- 608C	10 to 480 MC	Output 0.1 μv to 1 v into 50 ohm load. Pulse or CW modulation. Direct calib.	$ 950.00
-hp- 608D	10 to 420 MC	Output 0.1 μv to 0.5 v. Incidental FM 0.002 entire range.	1,050.00
-hp- 612A	450 to 1,200 MC	Output 0.1 μv to 0.5 v into 50 ohm load. Pulse, CW or square wave modulation. Direct calibration.	1,200.00
-hp- 614A	800 to 2,100 MC	Output 0.1 μv to 0.223 v into 50 ohm load. Pulse, CW or FM modulation. Direct calib.	1,950.00
-hp- 616A	1,800 to 4,000 MC	Output 0.1 μv to 0.223 v into 50 ohm load. Pulse, CW or FM modulation. Direct calib.	1,950.00
-hp- 618B	3,800 to 7,600 MC	Output 0.1 μv to 0.223 v into 50 ohm load. Pulse, CW, FM or square wave modulation. Direct calibration.	2,250.00
-hp- 620A	7,000 to 11,000 MC	Output 0.1 μv to 0.071 v into 50 ohm load. Pulse, FM or square wave modulation. Separate power meter and wave meter section.	2,250.00
-hp- 623B	5,925 to 7,725 MC	Output 70 μv to 0.223 v into 50 ohm load. FM or square wave modulation. Separate power meter and wave meter section.	1,750.00
-hp- 624C	8,500 to 10,000 MC	Output 3.0 μv to 0.223 v into 50 ohm load. Pulse, FM or square wave modulation. Separate power meter and wave meter section.	2,265.00Δ

Data subject to change without notice.

WRITE FOR COMPLETE SPECIFICATIONS

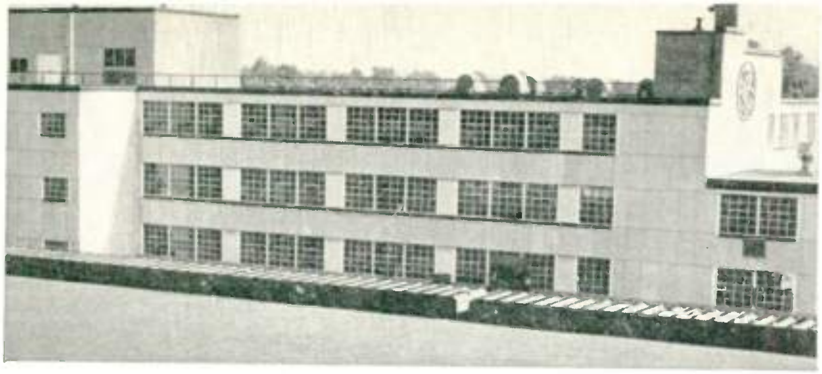
HEWLETT-PACKARD COMPANY

DEPT. W, 3340D PAGE MILL ROAD, PALO ALTO, CALIFORNIA, U.S.A.
Cable "HEWPACK"

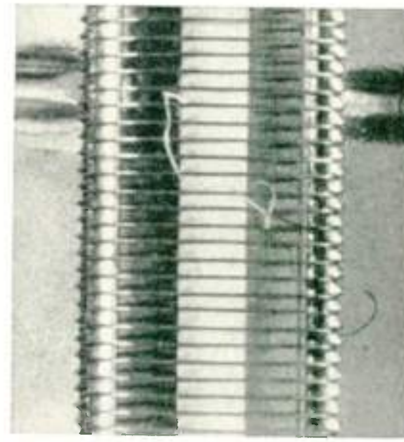
FIELD REPRESENTATIVES IN ALL PRINCIPAL AREAS

hp COMPLETE COVERAGE, HIGHEST QUALITY

AT G.E., FURTHER



SEPARATE LARGE G-E FACTORY BUILDING at Owensboro, Ky., is devoted to the production of 5-Star miniature and subminiature tubes for military, communications, and industrial uses. Dust control and air-conditioning during assembly and inspection help to assure G-E 5-Star Tube reliability.



LINT CAUSES GRID SHORT. This unretouched photograph of a tube grid magnified some 40 times, shows a stray particle of lint which can easily cause an inter-electrode short-circuit. Dust often causes a similar conductive path to form between closely-spaced tube elements.

LINT, DUST ARE SEALED OFF FROM 5-STAR PARTS. G-E tube grids are given a special cleaning, and then are rinsed, dried, inspected, and sorted. Afterwards G-E employees enclose them in treated paper bags, and fold and staple the ends of the bags so no lint or dust can enter.

AIR-LOCKS AT ALL DOORS. Employees and other persons entering, pass through air-locks, with a grating underfoot through which powerful suction removes any loose dirt from their shoes. All outer garments, lunches, and personal articles are left outside, in a separate cloak-room. Traffic is closely controlled, and those permitted to enter must wear lint-free clothing. Incoming mail is left in the air-locks, for pick-up and delivery later on by employees who work inside the "Snow White" area.

"OPERATION SNOW WHITE" INCREASES TUBE RELIABILITY

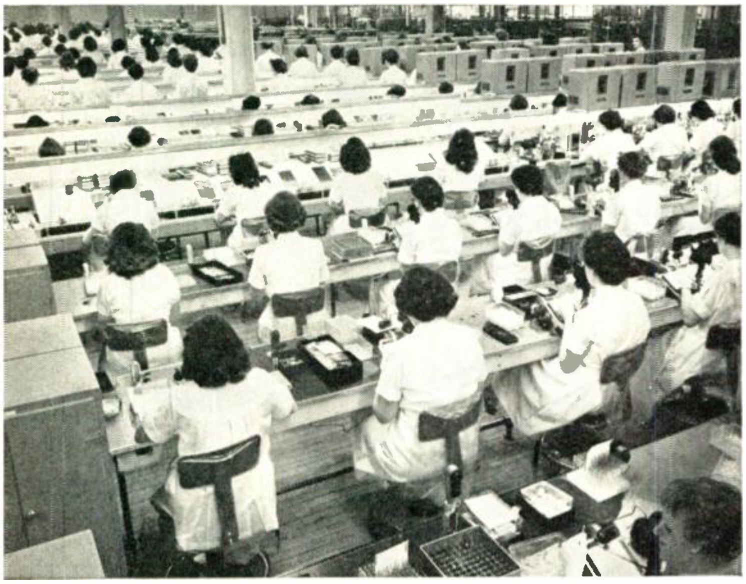
5-Star Tube inoperatives are greatly reduced by ridding assembly and inspection areas of lint and dust!

Optimum cleanliness during manufacture has resulted in a two-thirds drop in G-E 5-Star Tube inoperatives—mainly caused by intermittent short-circuits from lint. 100% factory tests prove this gain in dependability.

G-E "Operation Snow White" shields the work of 750 skilled employees from lint, dust, and dirt; involves the operations of a whole tube factory; helps assure the reliability of 5-Star types—both miniatures and subminiatures—in many million critical military, communications, and industrial tube sockets.

The extensive and important story is told briefly in the pictures and text on these pages. Ask for additional 5-Star Tubes information!

Learn why G-E 5-Star Tubes—specially designed, built, and performance-tested—are the most dependable tubes you can specify and install! *Tube Department, General Electric Company, Schenectady 5, New York.*



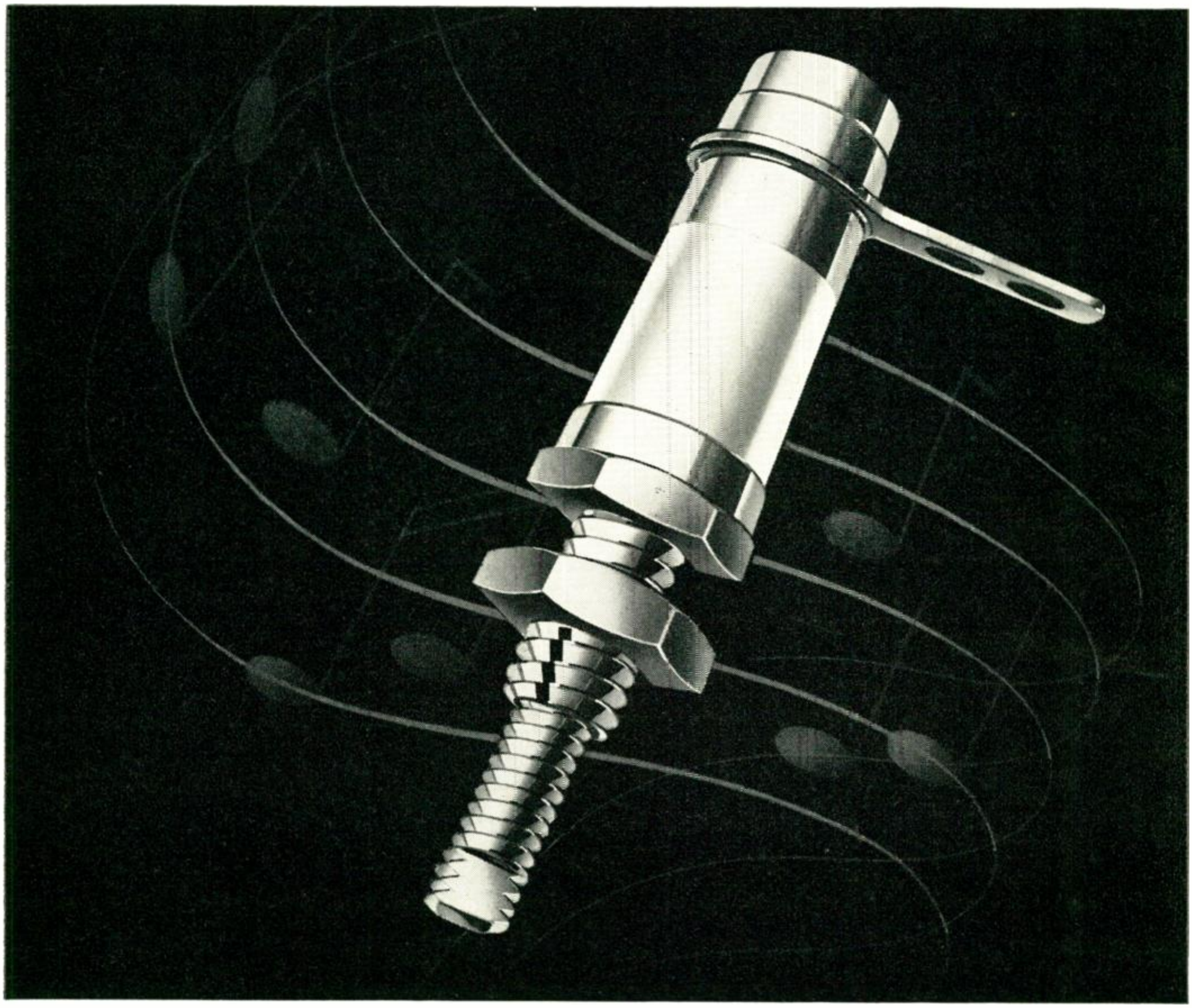
WHITE LINT-FREE UNIFORMS of Nylon and Dacron are worn by the 750 selected, trained employees who assemble and inspect G-E 5-Star Tubes. The entire working area, a part of which is shown above, is pressurized to keep out dust, with air that first has been thoroughly filtered, then dehumidified, and cooled.

ASSEMBLY, INSPECTION UNDER GLASS. G-E 5-Star Tube assembly and microscope inspection are carried out under special protective hoods that are glass-paneled for work observation. Employees wear rubber finger cots—changed every hour—to avoid contaminating the tube parts with any dirt or moisture.

Progress Is Our Most Important Product

GENERAL ELECTRIC

162-1B3



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.

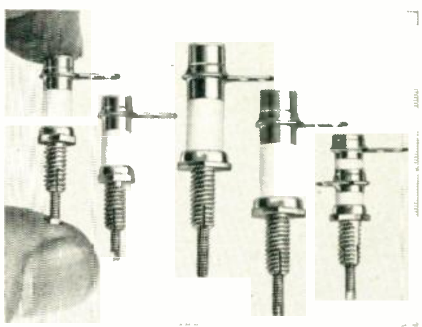
This tuning sleeve is at ground potential and can be locked firmly to eliminate undesirable capacity change. Each CST-50 is provided with a ring terminal with two soldering spaces.

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.



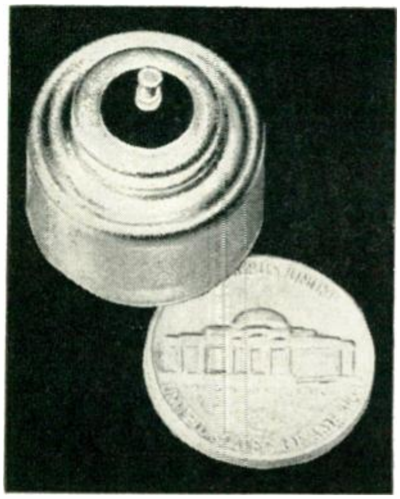
CTC

CAMBRIDGE THERMIONIC CORPORATION

makers of guaranteed electronic components, custom or standard

High Temperature Tantalum Capacitors

Cornell-Dubilier Electric Corp. has announced the development of a new Tantalum slug type electrolytic capacitor designed to operate under wide temperature ranges.



These new type "TH" Tantalums are rated from −55°C to +125°C. Units rated to +175°C can be supplied on specific order. Standard case size $\frac{3}{8}$ inch × $\frac{7}{8}$ inch to 120 $\mu\mu$f; only slightly larger to 240 $\mu\mu$f. Series combinations can be supplied at higher capacities and voltage ratings. These new capacitors are suited for operation under conditions of high G shock, high thermal cycling, and severe vibration.

Standard units range from 25 to 120 $\mu\mu$f with a voltage range of 18 to 100 volts dcw. Higher capacitances and voltages to 630 volts dcw, can be supplied. For further information send for Engineering Bulletin No. 529.

Optical Sub-assemblies

Zenith Optical Laboratory, Copiague, L. I., N. Y., has in operation an extensive department devoted to the design and fabrication of Lenses, Prisms, Optical Flats and Windows, as well as complete Optical Sub-assemblies. The fabricating, grinding, and polishing activities are directed by J. Balaze, an experienced production engineer in the optics field.

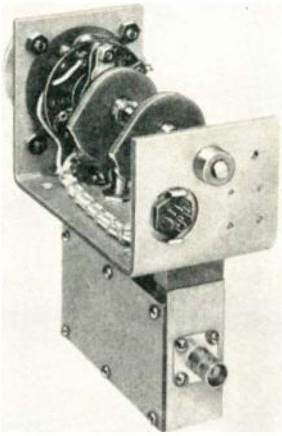
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Another feature of the expanded Zenith facilities is their high-vacuum coating department. Among the materials evaporated are Magnesium Fluoride, Aluminum, Silver, and Gold. Dichroic Coatings and Vignetted Coatings are also produced to specifications.

An enlarged designing and engineering department is available for the solution of optics problems. Co-operation on prototypes will be gladly extended. While their factory is geared for production runs, small quantity orders are readily accommodated. For additional information write to M. Hoffman, Mgr.

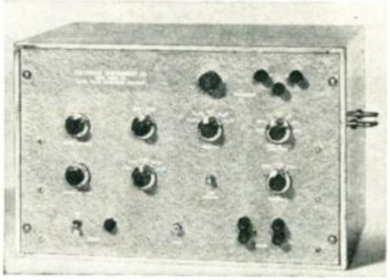
Remote Control RF Attenuator

A new remote control adaption of its standard rf attenuator has just been developed and made available by the **Daven Co.,** Dept. RH, 195 Central Ave., Newark 4, N. J. The new unit, Series 544, is cam operated and solenoid driven. Its salient feature is a special arm which permits remote selection of any of the following cam operated positions on the attenuator, first position and second position, entirely out of circuit. First attenuation position in the circuit, and second attenuation position out of the circuit. Both attenuation positions in the circuit. This cycle can be repeated indefinitely.



This new type features all of the characteristics of the standard Daven unit that is, it is flat from dc to 225 mc and maintains a resistance accuracy to ±2 per cent. Power rating is 0.25 watt. The unit operates on 115 volts and measures $5\frac{3}{8}$ inches high × $5\frac{1}{2}$ inches long × $2\frac{3}{8}$ inches wide. Although the unit shown has 3 steps, variations can be provided with up to 5 steps plus zero attenuation. A maximum of 20 db per step can be supplied. For further information write to the company.

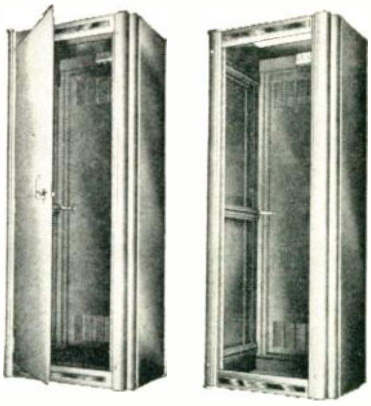
Transistor Curve Tracer



Polyphase Instrument Co., Bryn Mawr, Pa., announces its Model TA-3A Transistor Analyzer, a transistor family curve tracer. It displays on an oscilloscope the R_{12}, R_{22}, and H_{12} families in the grounded base connection and the R_{22} family in the grounded emitter connection. Either N, P, NPN, or PNP transistors can be tested. Collector current power supply has a peak rating of 100 ma at 100 v enabling the instrument to be used for testing power transistors as well as conventional transistors.

Universal Cabinet Racks

The Par-Metal Products Corp., 32–62 Forty Ninth St., Long Island City 3, N. Y., line of completely assembled racks now includes the universal type which has provisions for mounting chassis supports, shelves, sliding shelves, and standard sliding devices. These racks, in addition to being available with or without front doors, are also available with or without detachable side panels, and may be used singly, or in rows.



The racks are made in standard units of 19 and 24 inch wide panels, in 18 and 24 inch depths, and in $48\frac{1}{8}$, $67\frac{3}{8}$, $76\frac{1}{8}$ and $83\frac{1}{8}$ inches standard heights. Panel spaces are 42, $61\frac{1}{4}$, 70 and 77 inches; clear inside depths are $16\frac{1}{4}$ and $22\frac{1}{4}$ inches for units with front and rear doors, and $16\frac{7}{8}$ and $22\frac{5}{8}$ inches for units with rear doors only.

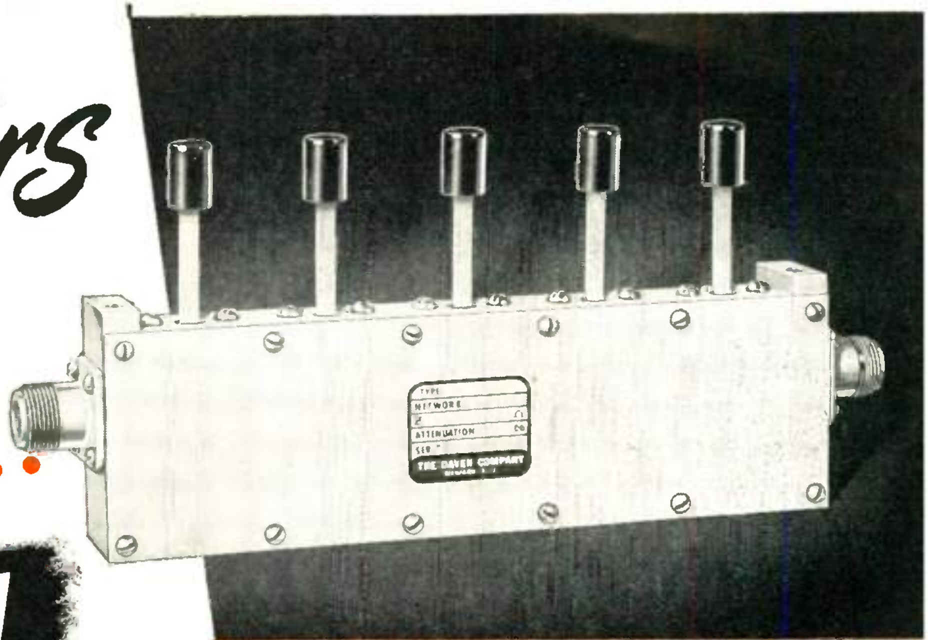
A unique feature of the racks is that intermixing of similar height racks with any combination of 19 and 24 inch wide panels are possible.

The racks range in price from $70.00 to $180.00. For additional information write for a copy of the latest Par-Metal catalog.

(Continued on page 20A)

IN *Attenuators*

WHY DOES ONE NAME.. DAVEN STAND OUT?



Series 550-RF Attenuator

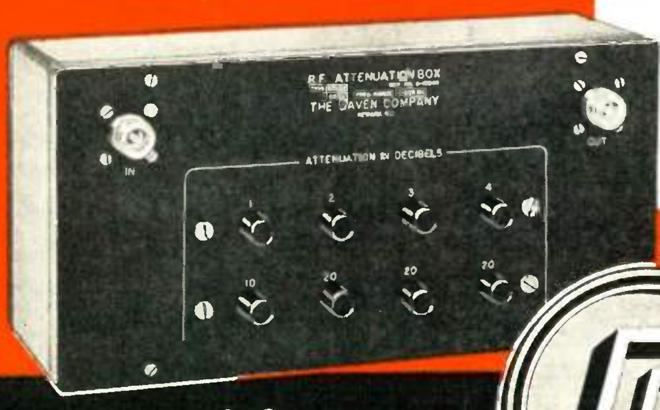
In addition to Daven being the leader in audio attenuators, they have achieved equal prominence in the production of RF units. A partial listing of some types is given below.

DAVEN Radio Frequency Attenuators, by combining proper units in series, are available with losses up to 120 DB in two DB Steps or 100 DB in one DB Steps. They have a zero insertion loss and a frequency range from DC to 225 MC.

Standard impedances are 50 and 73 ohms, with special impedances available on request. Resistor accuracy is within ± 2% at DC. An unbalanced circuit is used which provides constant input and output impedance. The units are supplied with either UG-58/U* or UG-185/U** receptacles.

Because DAVEN makes the most complete, the most accurate line of ATTENUATORS in the world!

TYPE	LOSS	TOTAL DB	STANDARD IMPEDANCES
RFA* & RFB 540**	1, 2, 3, 4 DB	10	50/50Ω and 73/73Ω
RFA & RFB 541	10, 20, 20, 20 DB	70	50/50Ω and 73/73Ω
RFA & RFB 542	2, 4, 6, 8 DB	20	50/50Ω and 73/73Ω
RFA & RFB 543	20, 20, 20, 20 DB	80	50/50Ω and 73/73Ω
RFA & RFB 550	1, 2, 3, 4, 10 DB	20	50/50Ω and 73/73Ω
RFA & RFB 551	10, 10, 20, 20, 20 DB	80	50/50Ω and 73/73Ω
RFA & RFB 552	2, 4, 6, 8, 20 DB	40	50/50Ω and 73/73Ω



Series 640-RF Attenuation Network

These units are now being used in equipment manufactured for the Army, Navy and Air Force.

Write for Catalog Data.

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)

Logarithmic-Linear Amplifier

Logarithmic functions, such as gain and attenuation expressed in decibels, are rapidly plotted on linear paper using a standard recording milliammeter in conjunction with the Logarithmic-Linear Amplifier, Model 120A developed by **Color Television, Inc.**, 974 E. San Carlos Ave., San Carlos, Calif. Complete antenna radiation patterns have been obtained in less than two minutes using the Log-Linear Amplifier in automatic antenna pattern recording systems the firm claims.

Designed primarily for use with an rf source modulated at 1000 cps, the input provides a metered, variable 0 to 10 ma dc supply for bolometer operation. Alternatively, the instrument can be supplied with a crystal input.

The output is presented on a 4-inch meter and is also available at a meter jack for driving either a 1-ma dc recording milliammeter or a high-impedance recorder. The logarithmic channel has a dynamic voltage range of 100 db, equivalent to a 50-db power change in a square-law detector. The corresponding meter scale reads attenuation from 0 to 50 db in equally spaced 1-db graduations. When the linear channel is used, the output is read on a typical db scale with 3 db at approximately mid-scale so that antenna half-power points can be accurately determined.

Microwave Antenna Test Rooms

McMillan Industrial Corp., Ipswich, Mass., has recently put on the market a new type of self-contained "free space" room for microwave antenna testing. These are prefabricated units which can be easily assembled for either indoor or outdoor installation, and can be used as boresight tunnels or test ranges as well as for pattern or other measurements where microwave energy absorption is required.

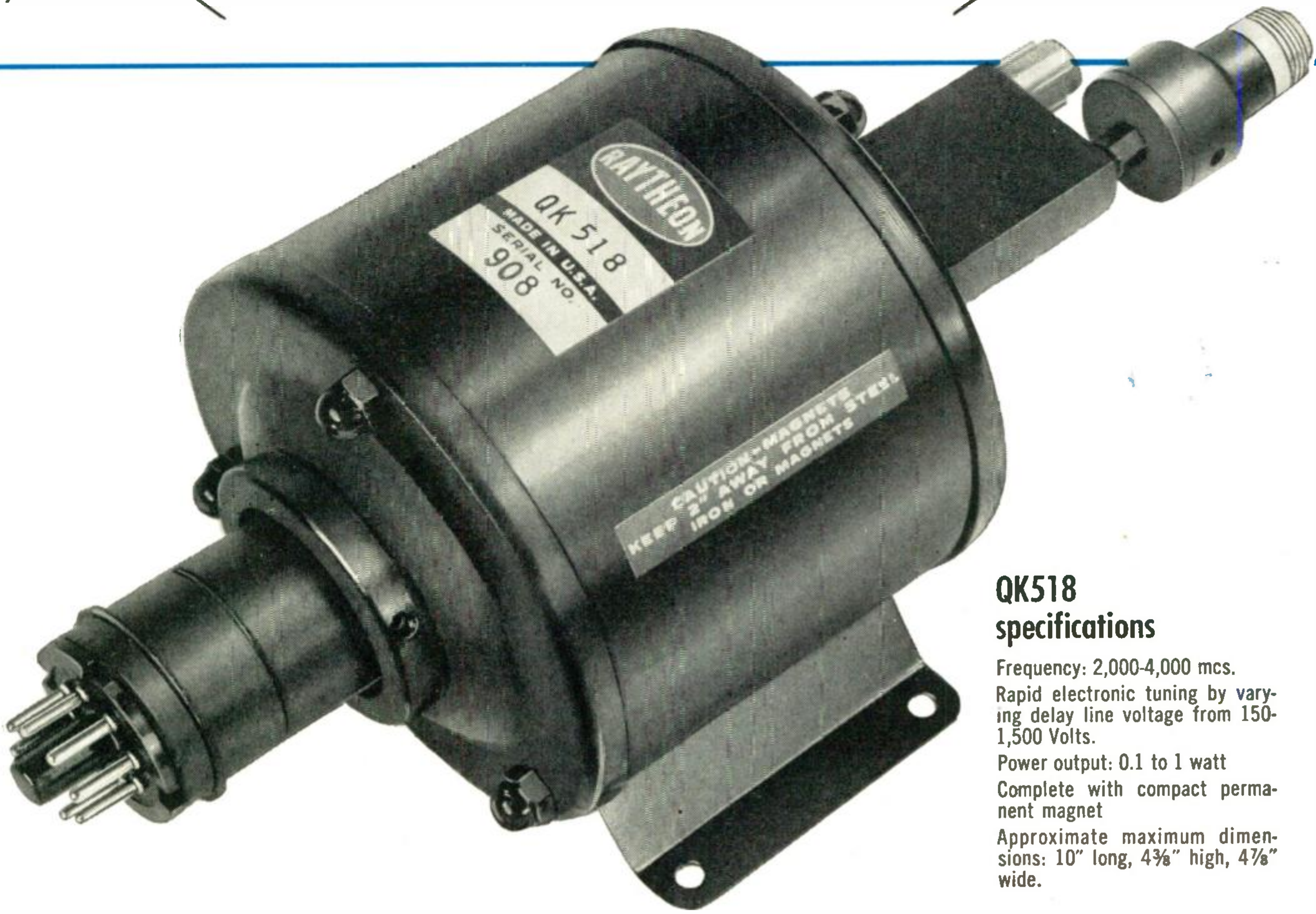
Each construction is completely self-supporting and needs no other structural members. It consists of a metal framework and plywood panels with microwave absorbing material permanently attached. Standard panel sizes are 4×6 and 4×8 feet, but any size panel can be built for a

(Continued on page 90A)

Raytheon — World's Largest Manufacturer of Magnetrons and Klystrons

VOLTAGE TUNABLE

1,000 mc. ⟵⟶ 16,000 mc.



QK518 specifications

Frequency: 2,000-4,000 mcs.

Rapid electronic tuning by varying delay line voltage from 150-1,500 Volts.

Power output: 0.1 to 1 watt

Complete with compact permanent magnet

Approximate maximum dimensions: 10" long, 4⅜" high, 4⅞" wide.

NEW Raytheon Backward Wave Oscillator Series

for wide, rapid electronic tuning—1,000 mc. to 16,000 mc.

The tubes in this revolutionary new line of Raytheon Backward Wave Oscillators give you four outstanding performance advantages:

1. Electronically tunable over an *extremely* wide range of frequencies
2. Frequency insensitive to load variations
3. High signal-to-noise ratio
4. Can be operated under conditions of amplitude or pulse modulation

These new tubes are finding fast-growing applications in microwave equipment, including radar and signal generators.

Write today for free Data Booklet on the QK518 (above) which is available for delivery. We'll also be happy to answer any questions you may have on this new line.

RAYTHEON MANUFACTURING COMPANY

RAYTHEON

Microwave and Power Tube Operations, Section PL-30, Waltham 54, Mass.

Excellence in Electronics

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

NEW SIZES AVAILABLE NOW!

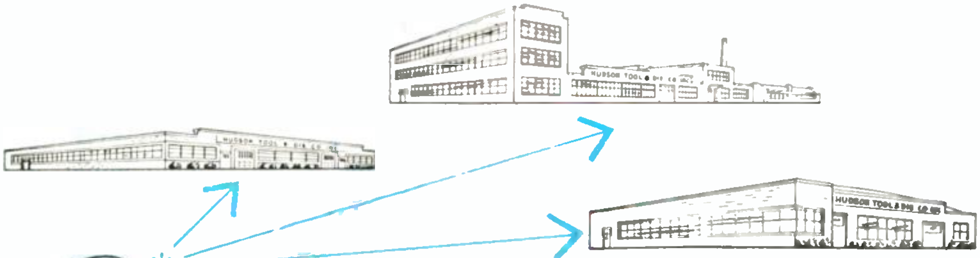
STANDARD ROUND CASES

HU-693:		
OUTSIDE DIA.	1 5/16"	
LENGTH	1 7/16"	
HU-703:		
IOUTSIDE DIA.	1 23/32"	
LENGTH	1"	
HU-694:		
OUTSIDE DIA.	1 49/64"	
LENGTH	5 1/4"	
HU-695:		
OUTSIDE DIA.	2 15/32"	
LENGTH	6"	

STANDARD RECTANGULAR CASES

HU-714:	
DIMEN. A	11/32"
DIMEN. B.	17/32"
LENGTH	1 1/8"
HU-690:	
DIMEN. A	3/4"
DIMEN. B	1 41/64"
LENGTH	1 1/32"
HU-704:	
DIMEN. A	59/64"
DIMEN. B	1 25/64"
LENGTH	1 3/16"
HU-705:	
DIMEN. A	29/32"
DIMEN. B	1 15/32"
LENGTH	1 5/16"
HU-720:	
DIMEN. A	1 1/32"
DIMEN. B	1 3/16"
LENGTH	2 1/2"
HU-710	
DIMEN. A	2 1/4"
DIMEN. B	3 3/4"
LENGTH	4 1/4"
HU-716:	
DIMEN. A	2 5/8"
DIMEN. B	2 57/64"
LENGTH	2 5/8"
HU-701:	
DIMEN. A	2 13/16"
DIMEN. B	3 5/16"
LENGTH	4 7/8"
HU-712:	
DIMEN. A	3 1/16"
DIMEN. B	3 9/16"
LENGTH	4 7/8"

CLIP THIS PAGE AND FILE WITH YOUR NEW HUDSON CATALOG!



3* FULLY-EQUIPPED PLANTS

Offering the Most Complete Line of Precision Drawn Cases and Covers in the Industry!

...THREE COMPLETE SERVICES!*

Hudson Standard Metal Closures

Over 1000 economical standard types mean HUDSON can supply precision components at commercial prices. A wide variety of optional features make it possible to solve all but the most unusual closure requirements with standard types selected from HUDSON stocks.

Hudson Quality Metal Stampings

Metal parts produced to your exact specifications at prices that reflect the economies of mass production methods. Hudson can work to close tolerances and maintain uniformity throughout production runs. Quotations supplied promptly on receipt of drawings.

Hudson Sheet Metal Facilities

Depend on HUDSON for expert fabrication of simple or complex sub-assemblies. Facilities include certified welding of alloys, silver soldering, brazing and chrome plating.

NEW CATALOG READY, NOW!

The Hudson story is contained in one handy catalog. Full descriptions of all standard items and complete information on Hudson metal working facilities. Call or write for your copy, now!

Precision Components of Steel, Aluminum, Copper, Brass, Mu Metal

HUDSON TOOL & DIE COMPANY • INC
118-122 SOUTH 14th ST., NEWARK 7, N. J.

"Eimac 450TH in excellent operating condition after 49 months of dependable service"

The 450T power triode first designed and produced by Eimac in 1937 is still a standard Eimac product widely acclaimed by the communications industry. It symbolizes the uncompromising quality, performance and reliability that has made Eitel-McCullough, Inc. the world's largest manufacturer of transmitting tubes.

WBKV 1470 K.C.
WEST BEND BROADCASTING CO.
Serving the ... Region

TELEPHONE 1640 WEST BEND, WISCONSIN

December 6, 1954

Eitel-McCullough, Incorporated
San Bruno, California

Gentlemen:

On November 30, 1954 I retired from service an Eimac 450TH, Serial Number C9-2416.

This tube was used as the Power Amplifier in the Gates BC-500D transmitter operated at this station.

It was installed in the transmitter and put into operation on November 1, 1950. The total hours of operation obtained before replacement amounts to 17,892.

Now to further explain the reason for this letter I must tell you that the tube is still in excellent operating condition as far as I am able to tell by its daily performance. It was replaced just on general principles after 49 months of dependable service.

Very truly yours,

Paul J. Ripple
Chief Engineer

PJR/ps

Mr. Paul J. Ripple, chief engineer, Station WBKV, West Bend, Wisconsin, holds Eimac 450T that he retired, "just on general principles after 17,892 hours of dependable service."

Eimac

EITEL-McCULLOUGH, INC.
SAN BRUNO · CALIFORNIA

4 FAMOUS *Collins* COMPONENTS ARE READY TO WORK FOR YOU

AUTOTUNES AND AUTOPOSITIONERS

By means of the Collins Autotune, shafts or devices requiring accurate positioning can be automatically reset to any of several variable positions. Autopositioners are used where up to 20 or more fixed positions are needed. Rotational reset accuracy .05°; Autotune Channels: 8-12 each independently variable over entire range, this may vary from a fraction of one turn to as many as 20 turns. Output torque is available in the range from ½-24 pound-inches. Operation time: as low as 1 sec.; System weights: as low as 2 lbs.; Power supply: 28 v dc, 110 v ac 50/60 cycles or other conventional sources.

MECHANICAL FILTERS

The Collins Mechanical Filter is an electro-mechanical bandpass filter, smaller than the usual i-f transformer, but providing better i-f selectivity than several stages of conventional tuned i-f circuits. The Mechanical Filter, readily adaptable to existing or new i-f designs, can be supplied with bandwidths from 500 cps to 12 kc for center frequencies from 100 kc to 500 kc, with —60 db bandwidths less than 3 times the —6 db bandwidths. Response variation within the passband is less than 3 db. Performance is dependable from —40°C to +85°C with relative humidities up to 95%.

PRECISION TUNED OSCILLATORS

Collins Precision Tuned Oscillators are permeability tuned and incorporate a precision lead screw. Mechanically rugged and sealed against atmospheric changes, these accurate R.F. sources are individually compensated for temperature and voltage variations. Fundamental frequencies in the range of 450 kc to 4 mc are covered and frequency multiplication may be employed to attain complete frequency coverage. R.F. output 1-30 v rms depending on model. Output is linear with lead screw rotation making dial design easy. Average short term (24 hr.) stability under specified condition is in the order of .003% after warmup.

HYSTERESIS SYNCHRONOUS MOTORS

Ideal for driving timing mechanisms, magnetic storage drums, recording charts and automatic frequency controls, Collins precision built hysteresis motors feature high starting torques and superior efficiency. Synchronous performance is possible from zero to as high as 1000 cycles per second. Diameter 2", length 2.3", torque up to 2 oz.-in. Some models have spilt windings for operation directly from plate circuits of 2 phase, direct-coupled push-pull amplifiers thus eliminating output transformers. Other models for 60 cps and 400 cps fixed-frequency operation from conventional power sources

For complete information on any of these Collins Components contact your nearest Collins office.

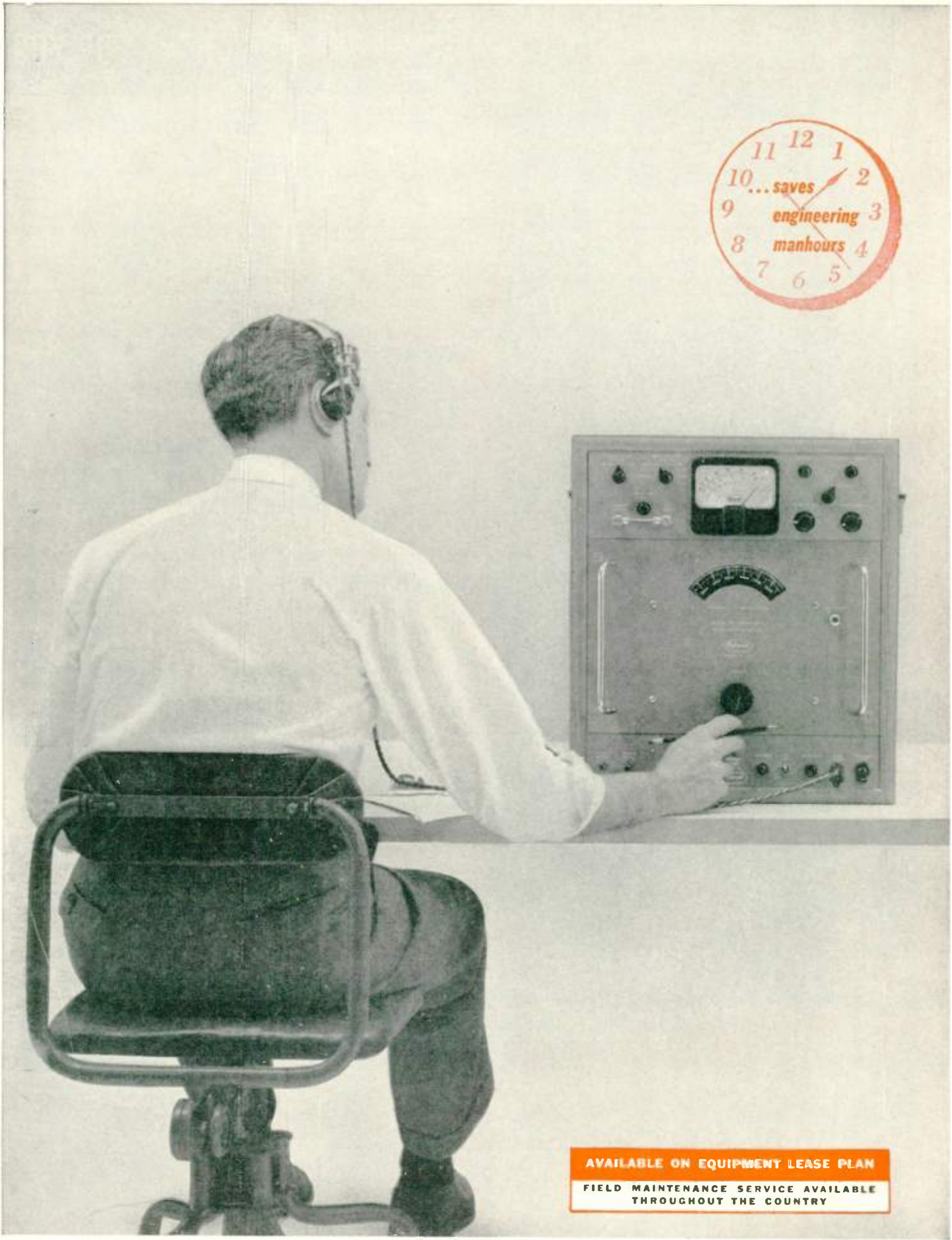
COLLINS RADIO COMPANY

CEDAR RAPIDS, IOWA

261 Madison Avenue, NEW YORK 16, NEW YORK
1200 18th Street N.W., WASHINGTON, D. C.
1930 Hi-Line Drive, DALLAS 2, TEXAS
2700 W. Olive Avenue, BURBANK, CALIFORNIA

COLLINS RADIO COMPANY OF CANADA, LTD.
74 Sparks Street, OTTAWA, ONTARIO

COLLINS



REPRESENTATIVES: • Albuquerque • Atlanta • Baltimore • Bayonne • Bridgeport • Buffalo • Chicago • Dayton • Fort Worth • Los Angeles • New York

Industrial Engineering Notes*

Aeronautics

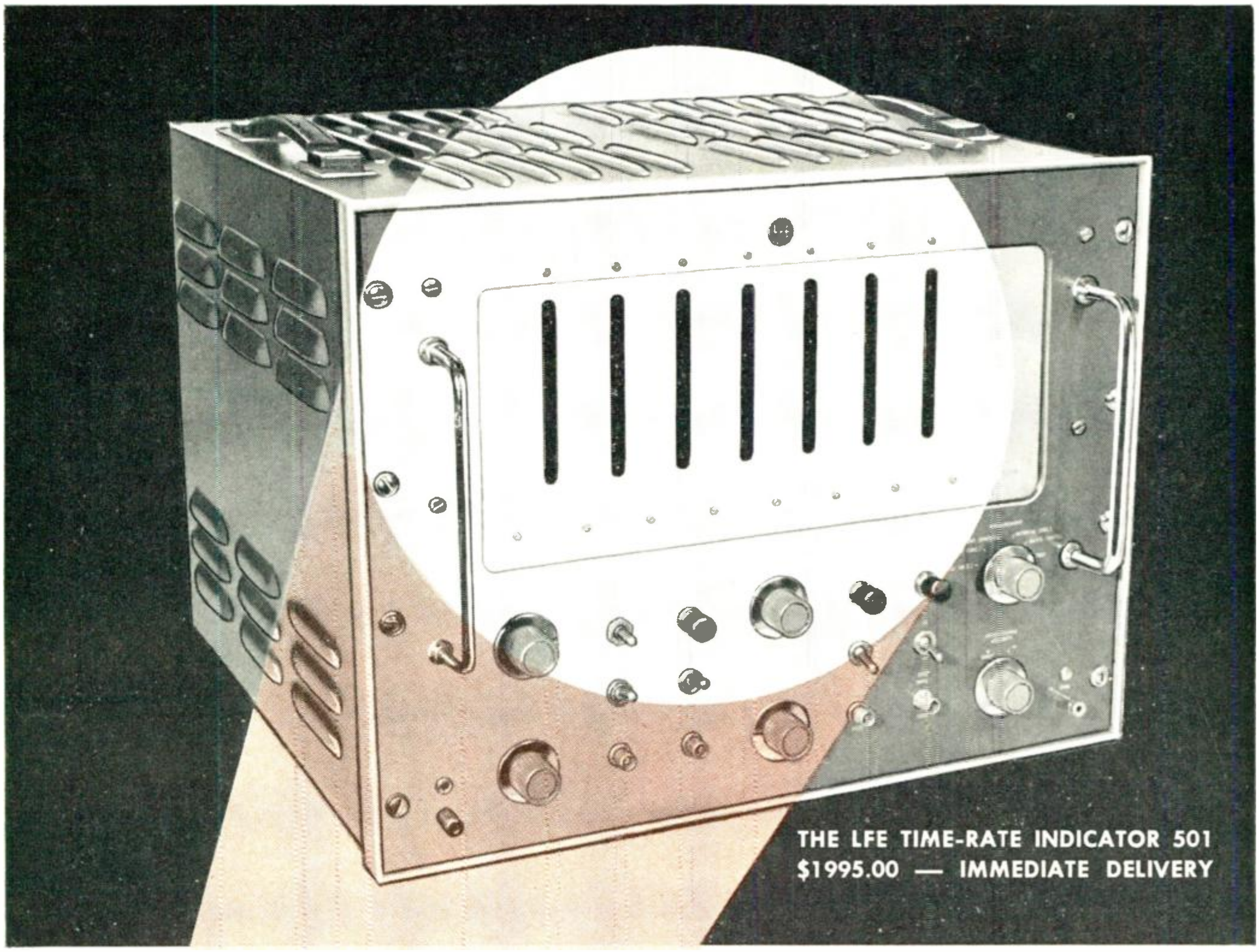
The Civil Aeronautics Board has announced an amendment to the environment test procedures for airborne radio equipment. The change was made in Section 16.30-3 of the CAB regulations. The section was changed as regards the reference number of the Radio Technical Commission for Aeronautics' paper covering environmental test procedures and clarifies the modification procedures of that section. The changes appeared in the Federal Register for March 25 (Vol. 20, No. 59) which is available from the Superintendent of Documents, Government Printing Office, Washington 25, D. C. . . . **Another congressional probe, with a possible effect on the electronics industry, looms on Capitol Hill.** The investigation, if it gets past the preliminary stages, will go into the relative merits of the two systems of radio navigational aids—VOR/DME and TACAN. The new controversy was touched off when the Air Navigation Development Board, following an extensive study of the matter, issued a report in which the majority of the board favored the TACAN system. The Air Navigation Development Board is composed of representatives of CAA, DOD and user groups and is charged with developing common air navigational aids for use by both civilian and military air carriers. The VOR/DME system now is used extensively on civil airways and VOR installations are virtually completed and DME ground installations are about 60 per cent completed. Tens-of-thousands of VOR equipments also have been installed in aircraft. The VOR/DME system was recommended in 1948 by Special Committee 31 of the Radio Technical Commission for Aeronautics. The TACAN system was developed by the Department of Defense.

RETMA Activities

The distribution of all RETMA television test charts now is being handled by the Engineering Department in New York. The following material is available at the prices noted (for overseas orders, add $1): *Resolution Chart with Supplemental Gray Scales,* $4.00; *Resolution Chart without Supplemental Gray Scales,* $2.00; *Supplemental Gray Scales for Resolution Chart,* $2.00; *Linearity Chart,* $2.00; *Registration Chart,* $3.00; *Linear Gray Scale,* $6.50; *Logarithmic Gray Scale,* $6.50. . . . **Establishment of a joint technical consulting group comprising representatives of the Federal Communications Commission and the radio-TV-electronics industry, to be supported by appropriate joint industry-FCC task forces as the occasion arises, was advocated by Dr. W. R. G. Baker, Director of the RETMA Engineering Department,**

(Continued on page 32A)

* The data on which these Notes are based were selected by permission from *Industry Reports,* issues of March 14, 28, and April 4 and 11, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.



Digitized research or control!

LFE's model 501 Time-Rate Indicator brings the advantages of high speed electronic counting to both *instrumentation* and *automation*. The broad applications of the 501 include research in the fields of frequency measurements, data recording, and process control. This unique counting tool will do more for less money than any similar device on the market.

Write to Laboratory for Electronics for their informative folder "From Instruments to Automation".

Check these features.

Ability to count pulses up to the rate of 10,000,000 per second with a relative accuracy up to ±1 count.

Ability to measure frequency in megacycles up to 10 mc with accuracy up to 1 part in 10 million with ability to totalize measurement.

Ability to measure periodic functions in decade units from 0.1 μsec to 100 sec.

Ability to make time interval measurements in the range from 1 μsec to 0.1 μsec.

Ability to do frequency ratio measurements between two input frequencies.

Built-in temperature compensated crystal controlled timing pulse generator.

Built-in wide band, high gain amplifier covering a bandwidth of 10 cps to 10 mc and with a sensitivity of 20 mv rms.

Decade Scalers from $30.00 Plug-in counter units including decade scalers from 20 kc to 10 mc, binary scalers that double the range of any particular unit, or pre-set scalers from 20 to 100 kc.

Creative developments in the field of electronics . . .

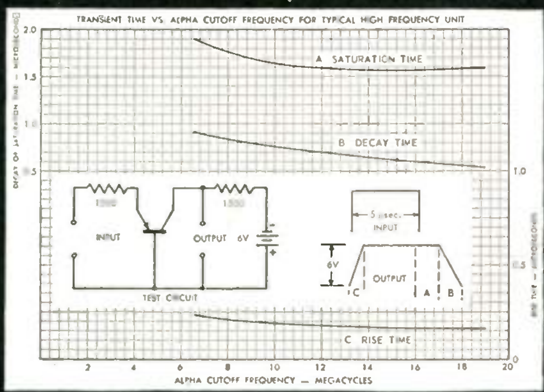
LABORATORY FOR ELECTRONICS, INC.

75 Pitts Street **Boston 14, Mass.**

New Information OF INTEREST TO
COMPUTER ENGINEERS AND DESIGNERS

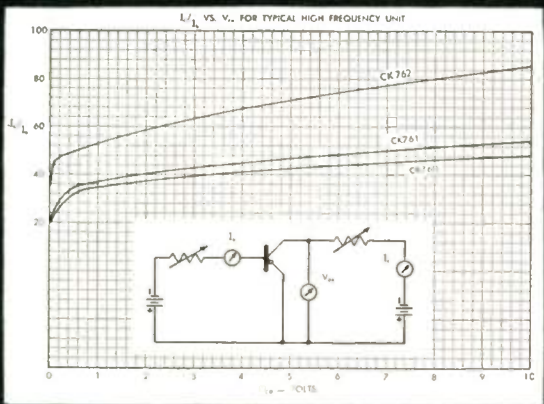
on the NEW RAYTHEON

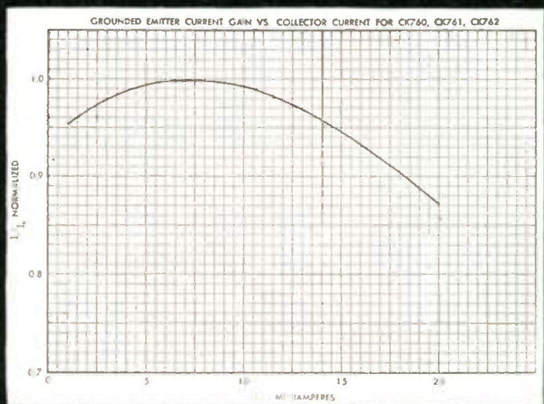
RF TRANSISTORS



RAYTHEON RF TRANSISTORS ARE

- *completely interchangeable* without selection of components
- *successfully field tested for an entire year.* In mass production for several months.
- made by the Raytheon perfected *fusion* process that has already produced nearly *two million* transistors





HIGH FREQUENCY TRANSISTORS — HERMETICALLY SEALED CASE

Type	Collector		Emitter		Extrin. Base Resis. ohms	Base Current Ampl. Factor	Alpha Freq. Cutoff mc.	Max. Junc. Temp. °C	Temp. Rise °C/mW	Coll. Capac. μμf
	Volts	Cutoff μA	mA	Cutoff* μA						
2N112/CK760	−6	1	−1.0	0.5	75	40	5	85	0.62	14
2N113/CK761	−6	1	−1.0	0.5	75	45	10	85	0.62	14
2N114/CK762	−6	1	−1.0	0.5	75	65	20	85	0.62	14

*Cutoff current measured at Vc = −12 volts

Note: above characteristics are average except where noted

RAYTHEON TRANSISTORS
more in use than all other makes combined

RADIATION COUNTER TUBES

CK1020 Thin wall beta, gamma counter, 900 volt
CK1021 Thin wall beta, gamma counter, 900 volt
CK1026 Halogen quenched, gamma counter, 900 volt
CK1049 Halogen quenched beta, gamma counter, 900 volt

Other counter types can be made to your order.

VOLTAGE REGULATOR TUBES

OA2	150 volts, 5-30 ma.
OB2-OB2WA	108 volts, 5-30 ma.
CK5787-CK5787WA	98 volts, 1-25 ma.
CK6542	150 volts, 5-25 ma.

GAS Filled TUBES

PERFORMANCE TESTED and backed by over THIRTY YEARS OF EXPERIENCE in the manufacture of gas tubes

VOLTAGE REFERENCE TUBES

CK5651-CK5651WA	85 volts, 1.5-3.5 ma.
CK5783-CK5783WA	85 volts, 1.5-3.5 ma.
CK6213	130 volts, 1-2.5 ma.

RAYTHEON

COLD CATHODE RECTIFIER TUBES

CK1042	2800 volt inverse, 8 ma. dc.
CK5517	2800 volt inverse, 12 ma. dc.
CK6174	2800 volt inverse, 3 ma. dc.

CORONA VOLTAGE REGULATOR TUBES

CK5962	700 volts, 2-55 μa
CK6437 (CK1037)	700 volts, 5-100 μa
CK1038	900 volts, 5-100 μa
CK6438 (CK1039)	1200 volts, 5-100 μa

500 to 3000 volt ratings available on special order.

Listed are representative tubes in each group. All are stable, rugged, reliable — worthy of your complete confidence.

All except Radiation Counter Tubes shown actual size.

RAYTHEON®

Excellence in Electronics

RAYTHEON MANUFACTURING CO.

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

Special 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

Industrial Engineering Notes

(Continued from page 28A)

at a Symposium on Spurious Radiation during the recent IRE convention. Dr. Baker also stressed the necessity for FCC consideration of the interference problem "as one in which the interfered service as well as the interfering service has a distinct possibility to seek a solution." Explaining the joint FCC-industry consulting group proposal, Dr. Baker said: "It would bring together the policy requirements of the FCC and the vast technical and economic resources of the industry, and would result in rules which would have the immediate acceptance of the entire industry." Dr. Baker recalled that this procedure was tried in September 1950 when a RETMA engineering committee met unofficially with representatives of the FCC to discuss radiation limits for FM and TV receivers.

Technical

The Office of Technical Services, Commerce Department, has listed studies in the field of electronics in its February 1955 issue of the "U. S. Government Research Reports," formerly titled the "Bibliography of Technical Reports." The following government-sponsored research reports can be purchased from the Photo-duplication Section, Library of Congress, Washington 25, D. C., for the reported price: "Research and Development on Paper Dielectric Capacitors—Final Report," PB 115977, microfilm, $3.75; photocopy, $10.25. "Search for New Nongaseous, Nonliquid Rectifying Systems," PB 115424 microfilm, $3.75; photocopy, $10.25. "Control of Self-Saturating Magnetic Amplifiers Using Rectified A-C with Varying Angle of Truncation," PB 115875, microfilm, $4.75; photocopy, $14. "Electronic Wave Spectrum Analyzer and its Use in Engineering Problems," PB 115917, microfilm, $4.50; photocopy, $12.75. "High Power Pulse Line Switching Devices," PB 115980, microfilm, $4; photocopy, $11.50. "Probe Measurements of Potential Within High-Density Electron Beams," PB 115783, microfilm, $8.50; photocopy, $30.25. "UHF Filtering Networks," PB 115939, microfilm, $3.25; photocopy, $9. "Resonatron Cavity Resonators," PB 116039, microfilm, $5.75; photocopy, $17.75. "Reflection Measurements on Wire Grids and Mesh Angles at 2,000 and 3,000 mc," PB 115920, microfilm, $2.75; photocopy, $6.50. "Research in Physical Electronics—Quarterly Report 7," PB 115548, microfilm, $3.75; photocopy, $10.25. "Study of the Generation and Detection of Electromagnetic Waves in the Millimeter Wave Region—Final Report," PB 115914, microfilm, $6.25; photocopy, $20.25. "Study of the Reflexion of Waves from an Inhomogeneous Medium by Means of a New First Approximation to a Solution of the General Linear, Second-Order Differential Equation and by Means of Iterations with Convergence of the Second Order," PB 115907, microfilm, $3; photocopy, $7.75. . . . **Several new advances in the electronics field**

(Continued on page 44A)

NEW Tektronix Portable Oscilloscope

gives you Laboratory Performance ...in the FIELD!

The Tektronix Type 310 is fully capable of performing much of your laboratory work, yet has the physical characteristics desirable for work away from your bench. It handles easily and fits into tight spots, simplifying field maintenance of complex electronic equipment. The high performance of the Type 310 can help you speed up your field work...its low weight and small size can ease your equipment handling problem.

Complete accessibility to tubes and components is maintained by a unique step-chassis construction, hinged at the rear. Accurate calibration and excellent linearity permit reliable quantitative measurements—you read time and amplitude directly from the screen. Functional panel design and versatile control system contribute to operator convenience, making this new oscilloscope an easy-to-use field and lab instrument.

IN THE FIELD

IN THE LAB

TYPE 310 CHARACTERISTICS

Portability
Overall dimensions—6¾" wide, 10" high, 17" deep.
Weight—only 23½" pounds.

Transient Response
Risetime—0.09 μsec.

Sensitivity
DC to 4 mc—0.1 v/div to 50 v/div in 9 calibrated steps, 0.1 v/div to 150 v/div continuously variable. AC-Coupled—3 db down at 2 cycles. AC-Coupled only, 2 cycles to 3.5 mc—0.01 v/div to 0.1 v/div in 3 calibrated steps.

Versatile Triggering
Internal, external, line...ac- or dc-coupled, and AUTOMATIC TRIGGERING.

Flat-faced CRT
3WP with 1.8-kv accelerating potential.
Edge-lighted graticule with ¼" divisions.

Wide Sweep Range
0.5 μsec/div to 0.6 sec/div, continuously variable. 18 calibrated sweeps from 0.5 μsec/div to 0.2 sec/div. Accurate 5-x magnifier extends calibrated sweep range to 0.1 μsec/div.

Horizontal Input
Sensitivity—1.2 v/div.

Voltage Calibrator
Square wave, approximately 1 kc—0.05 v to 100 v in 11 steps.

Jewel Warning Light
Indicates when controls are at non-calibrated settings.

Power Requirements
105 to 125 v, 60 to 800 cycles, 175 watts.

DC-Coupled Unblanking

All DC Voltages Electronically Regulated

Type 310 Cathode-Ray Oscilloscope—$595
f.o.b. Portland (Beaverton), Oregon

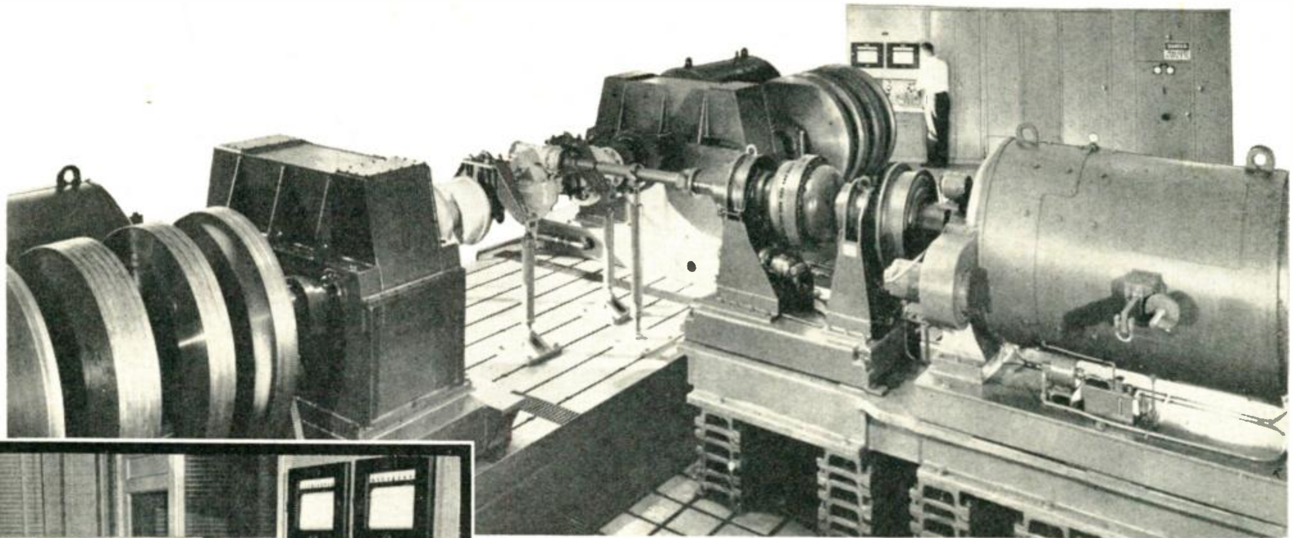
Please call your Tektronix Field Engineer or Representative for complete specifications, or write to:

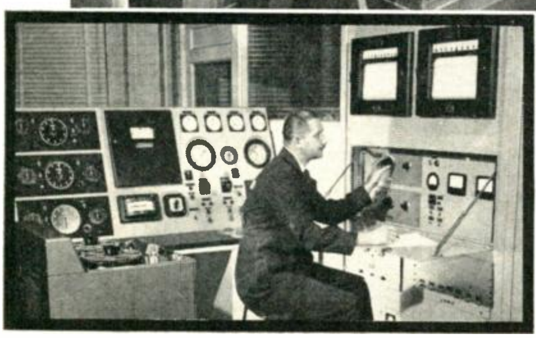
Tektronix, Inc.

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

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

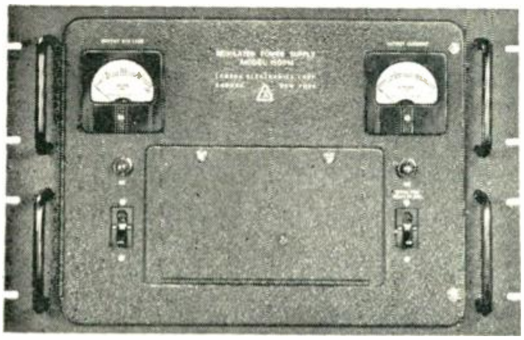
ANOTHER APPLICATION BY THE INSTRUMENTATION DIVISION OF

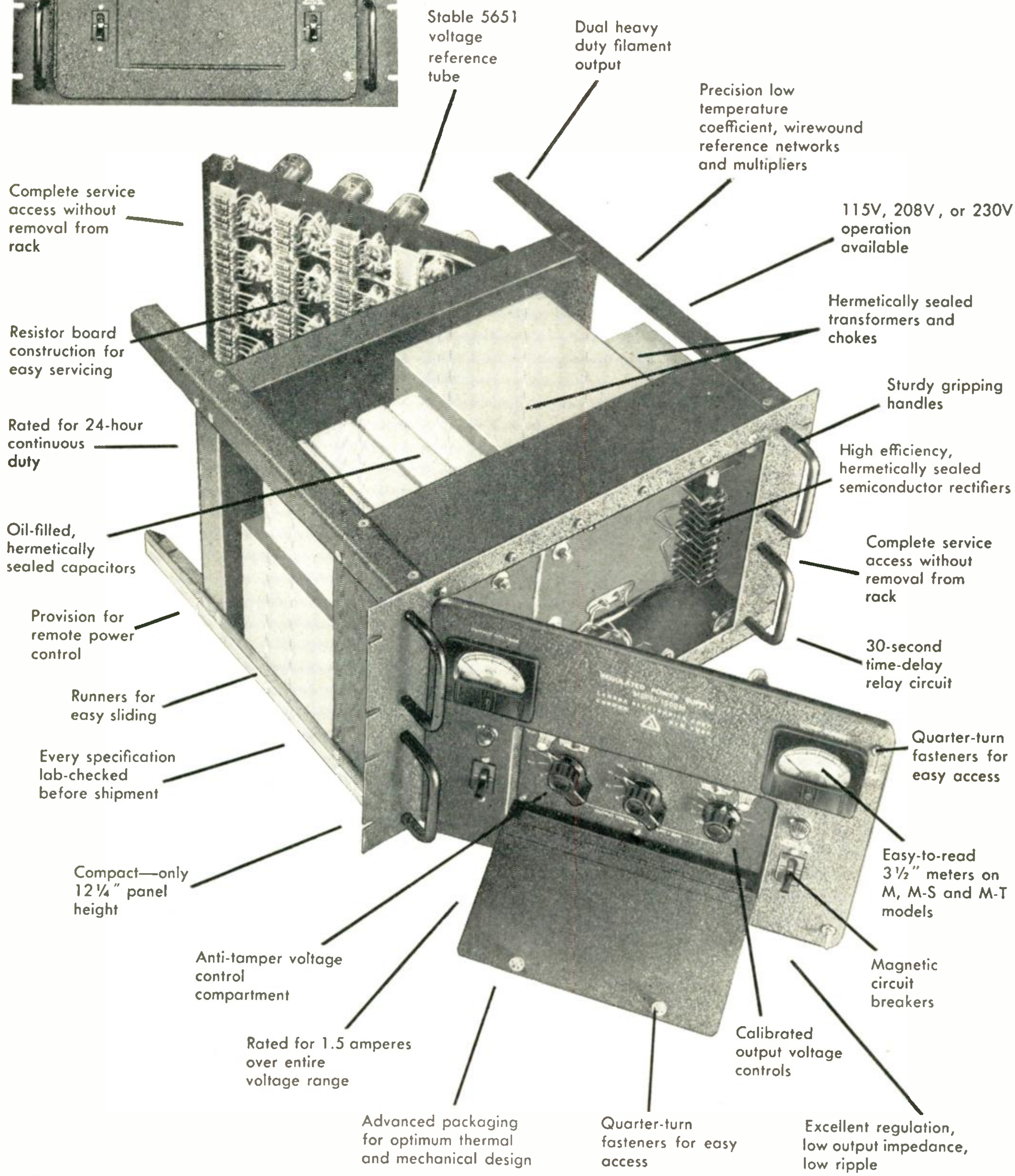
AMPEX CORPORATION • 934 CHARTER STREET, REDWOOD CITY, CALIFORNIA

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.

NEW! LAMBDA 1.5 AMP

VERSATILE, HEAVY-DUTY





REGULATED D.C.

POWER SUPPLY SERIES

COMPONENT POWER SUPPLY FOR ORIGINAL EQUIPMENT

CHOICE OF 28 OUTPUT VOLTAGE RANGES, THREE STANDARD A.C. VOLTAGE SOURCES

THREE SERIES: wide range (0-400 VDC); intermediate range (125 volt band); narrow range (50 volt band, centered at 25 volt intervals from 25 VDC to 375 VDC).

MODELS and PRICES*

MODEL[1]	OUTPUT VOLTAGE RANGE[2] at output currents from 0-1.5 amperes	INPUT POWER[3]	**BASE PRICE (U.S. and Canada) F.O.B. Factory Corona, N.Y.
WIDE RANGE MODELS			
WIDE RANGE MODELS: Voltage range is completely covered in 16 continuously variable bands. Each band range is 50 volts (±25 volts).			
1500	0-400 VDC	1530 W	**$695**
INTERMEDIATE RANGE MODELS			
INTERMEDIATE RANGE MODELS: Voltage range is completely covered in 4 continuously variable bands. Each band range is 50 volts (±25 volts).			
1510	0-125 VDC	980 W	**$595**
1511	25-150 VDC	1030 W	**$595**
1512	50-175 VDC	1080 W	**$615**
1513	75-200 VDC	1130 W	**$615**
1514	100-225 VDC	1180 W	**$635**
1515	125-250 VDC	1230 W	**$635**
1516	150-275 VDC	1280 W	**$655**
1517	175-300 VDC	1330 W	**$655**
1518	200-325 VDC	1380 W	**$655**
1519	225-350 VDC	1430 W	**$665**
1520	250-375 VDC	1480 W	**$665**
1521	275-400 VDC	1530 W	**$665**
NARROW RANGE MODELS			
NARROW RANGE MODELS: Voltage range is continuously variable.			
1530	0-50 VDC	830 W	**$555**
1531	25-75 VDC	880 W	**$555**
1532	50-100 VDC	930 W	**$565**
1533	75-125 VDC	980 W	**$565**
1534	100-150 VDC	1030 W	**$565**
1535	125-175 VDC	1080 W	**$585**
1536	150-200 VDC	1130 W	**$585**
1537	175-225 VDC	1180 W	**$605**
1538	200-250 VDC	1230 W	**$605**
1539	225-275 VDC	1280 W	**$625**
1540	250-300 VDC	1330 W	**$625**
1541	275-325 VDC	1380 W	**$625**
1542	300-350 VDC	1430 W	**$635**
1543	325-375 VDC	1480 W	**$635**
1544	350-400 VDC	1530 W	**$635**

[1]Model No. Suffix	Meters (3½")	Input Power Source Single Phase	Surcharge (see chart above for price**)
None	-	105-125 V, 50-60 CPS	None
M	2	105-125 V, 50-60 CPS	Add $30 to base price
S	-	190-225 V, 50-60 CPS	None
M-S	2	190-225 V, 50-60 CPS	Add $30 to base price
T	-	210-250 V, 50-60 CPS	None
M-T	2	210-250 V, 50-60 CPS	Add $30 to base price

[2]Current rating of 0-1.5 amperes applies over entire voltage range except for voltages below 100 VDC where the rating of 0.1-1.5 amperes applies.

[3]*With all outputs loaded to full ratings and input at 125 VAC, 225 VAC or 250 VAC.*

SPECIFICATIONS*

DC OUTPUT VOLTAGE: (regulated for line and load). At output currents from 0-1.5 amperes (see Note 2.) Refer to chart at left for voltage ranges of designated models.

Regulation (line) ....0.15% or 0.3 volt (whichever is greater). For input variations from 105-125 VAC, 190-225 VAC, or 210-250 VAC.

Regulation (load) ....0.15% or 0.3 volt (whichever is greater). For load variations from 0 to 1.5A except as modified by Note 2.

Internal Impedance ......Less than 0.4 ohms.

Ripple and Noise ....Less than 5 millivolts rms.

Polarity ....Either positive or negative may be grounded.

AC OUTPUTS (unregulated) ....Two outputs, isolated and ungrounded. Each is 6.5 VAC at 15A (at 115 VAC, 208 VAC or 230 VAC input). Allows for drop in connecting leads. May be connected in series for 12.6V (nominal) at 15A, or in parallel for 6.3V (nominal) at 30A.

AC INPUT .... 105-125 VAC, 50-60 CPS, single phase, 190-225 VAC, 50-60 CPS, single phase, or 210-250 VAC, 50-60 CPS, single phase. Refer to table at left for input power of designated models.

AMBIENT TEMPERATURE AND DUTY CYCLE ....Continuous duty at full load up to 50°C (122°F) ambient.

OVERLOAD PROTECTION:

External Overload Protection ....AC and DC magnetic circuit breakers. Trip-Free. Instant manual reset. Front Panel.

Internal Failure Protection ....Fuses, front panel access.

INPUT AND OUTPUT CONNECTIONS .... Heavy duty barrier terminal block, rear of chassis. 8 foot heavy duty rubber covered line cord with integral molded plug, also supplied.

METERS:

Output Voltage ....3½" rectangular voltmeter on metered models.

Output Current ....3½" rectangular ammeter on metered models.

CONTROLS:

DC Output Controls ....Located in recessed compartment with access door on front panel. Calibrated in volts DC.
Wide range units: Two band switches and continuously variable vernier-control.
Intermediate range units: One band switch and continuously variable vernier-control.
Narrow range units: Continuously variable vernier-control.

AC and DC Power ....Magnetic circuit breakers, front panel.

VOLTAGE REFERENCE ....A stable 5651 reference tube is used to obtain superior long-time voltage stability.

TIME DELAY RELAY CIRCUIT ....A 30 second time delay circuit is provided to allow tube heaters to come to proper operating temperature before high voltage can be applied.

PHYSICAL DATA:

Mounting ....Std. 19" rack mounting.

Size ....12¼" H x 19" W x 16½" D behind front panel.

Weight ....175 lb. net, 225 lb. ship. wt.

Panel Finish ....Black ripple enamel (standard). Special finishes available to customer's specifications at moderate surcharge.

*Specifications and prices effective May, 1955, subject to change without notice

A 1.5 AMP POWER SUPPLY FOR EVERY PURPOSE

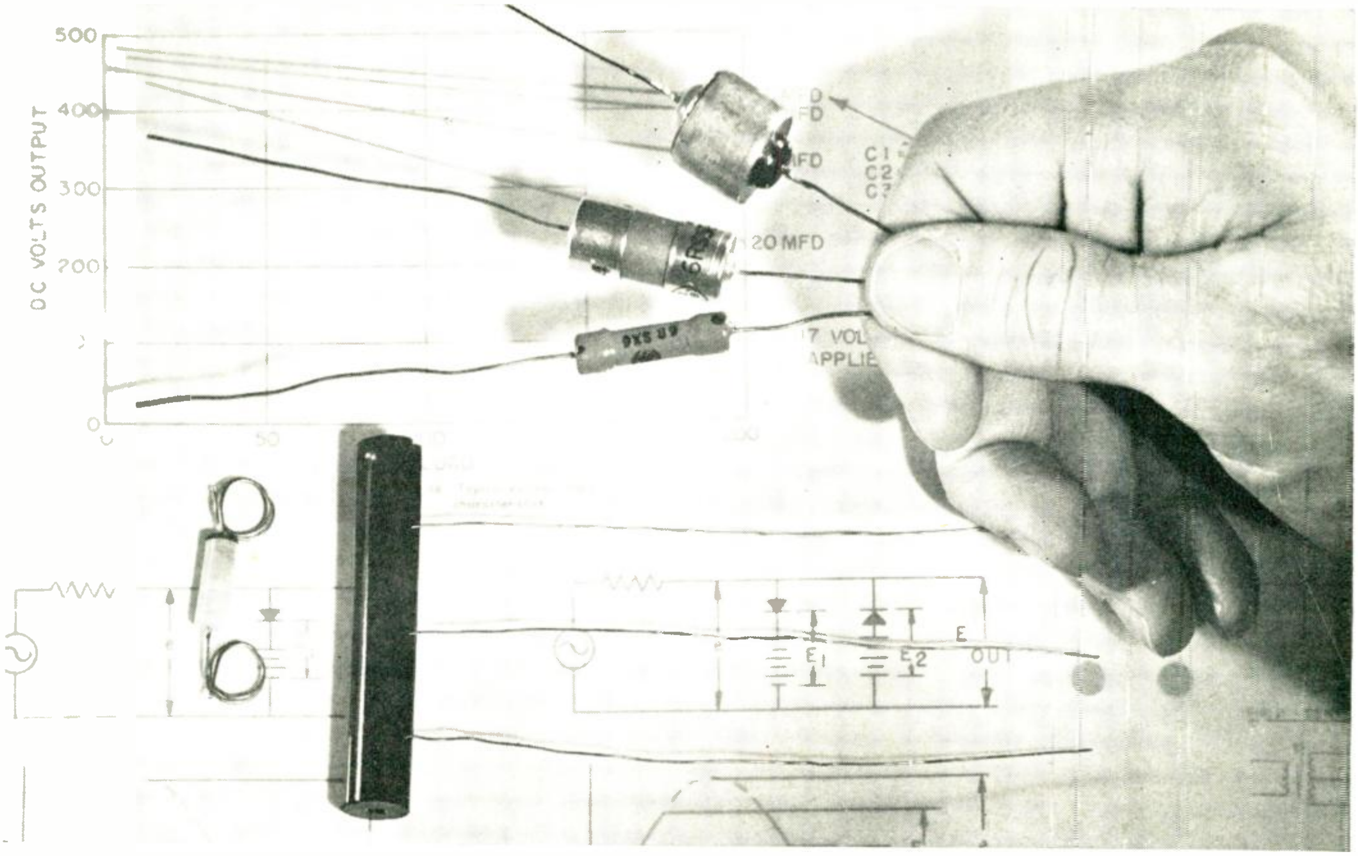
This new Lambda series gives you the economy of standard production models and the adaptability and precision engineering of custom units. Lambda's advanced electrical design provides rugged, stable operation.

Models in the 1.5 amp series are especially recommended as component power supplies for original equipment; in laboratories; for experimental models and prototypes; for computers; in industrial installations; for transmitters; and for use in the television and radio industries.

LAMBDA Electronics Corp.

THE FIRST NAME IN POWER SUPPLIES

103-02 NORTHERN BOULEVARD • CORONA 68, NEW YORK • TWining 8-9400



G-E Miniature *Vac-u-Sel** Rectifiers Provide 60,000 Hours Life; —65 C to 130 C Ambient Range

General Electric miniature Vac-u-Sel rectifier stacks provide outstanding advantages in the areas of:

- Long life expectancy—60,000 hours at 35 C
- Broad ambient temperature range—−65 C to 130 C
- Wide adaptability—variety of stack ratings to 9250 volts peak inverse.

Vac-u-Sel is the G-E trade-mark for a new line of metallic rectifiers with outstanding electrical characteristics.

LONG LIFE EXPECTANCY—Applications requiring 60,000 hours of life and more can be handled with assurance of highly dependable performance with these top-quality rectifier stacks. Long life is an inherent characteristic of these rectifiers. Aging (increase in forward drop) is exceptionally low.

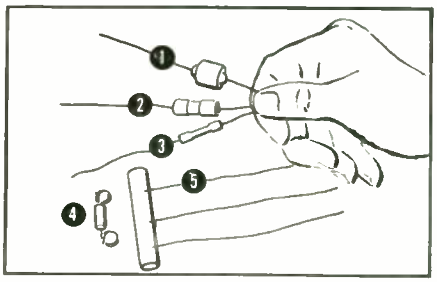
BROAD AMBIENT TEMPERATURE RANGE—All G-E miniature Vac-u-Sel rectifier cells are specially processed to maintain a high stability of characteristics over an ambient temperature range from −65 to 130 C. Full voltage ratings may be used in all high-temperature applications, and current need not be derated in cases where shorter life is acceptable.

WIDE ADAPTABILITY—Miniature Vac-u-Sel rectifiers are available in individual stacks rated up to 9250 volts peak inverse (6500 volts RMS). Higher voltages may be obtained by using two or more stacks in series. Basic cell ratings are 2.5 ma, 8 ma, and 25 ma (half wave).

Vac-u-Sel rectifiers are available in a variety of housings. The ceramic-tube and metal-tube housings are hermetically sealed. Military specifications on protective coatings are met by applying a special finish to the Textolite* tube stacks at additional cost, and by potting (sealing). Special housings can be offered for large-quantity applications.

PROMPT SERVICE—Immediate attention to any proposition can be obtained by contacting your nearest G-E Apparatus Sales Office, or by writing Section 461-37, General Electric Co., Schenectady 5, N.Y.

*Reg. Trade-mark of the General Electric Co.



VARIETY OF HOUSINGS available for Vac-u-Sel rectifiers. 1) Metal-clad casing, 2) Textolite tube, 3) Ceramic tube, 4) Nylon tube, 5) Slotted Textolite tube.

Progress Is Our Most Important Product

GENERAL ELECTRIC

CAPACITY	DIELECTRIC	SIZE	AVAILABLE CAPACITY TOLERANCES
	2-KV		
5-15	N-750	1/4"	5-10-20% GMV
16-47	N-750	5/16"	5-10-20% GMV
48-72	N-750	1/2"	5-10-20% GMV
73-200	N-750	5/8"	5-10-20% GMV
201-250	N-750	3/4"	5-10-20% GMV
251-330	N-750	7/8"	5-10-20% GMV
	3-KV		
5-15	N-750	5/16"	5-10-20% GMV
16-28	N-750	1/2"	5-10-20% GMV
29-56	N-750	5/8"	5-10-20% GMV
57-68	N-750	5/8"	5-10-20% GMV
69-180	N-750	3/4"	5-10-20% GMV
181-240	N-750	7/8"	5-10-20% GMV
	4-KV		
5-56	N-1500	5/8"	5-10-20% GMV
57-180	N-1500	7/8"	5-10-20% GMV
	5-KV		
5-30	N-1500	5/8"	5-10-20% GMV
31-60	N-1500	3/4"	5-10-20% GMV
61-130	N-1500	7/8"	5-10-20% GMV
	6-KV		
5-20	N-1500	3/4"	-10-20% GMV
21-100	N-1500	7/8"	-10-20% GMV

POWER FACTOR: .1% Max. @ 1M C (initial)
INSULATION: Durez phenolic—vacuum waxed

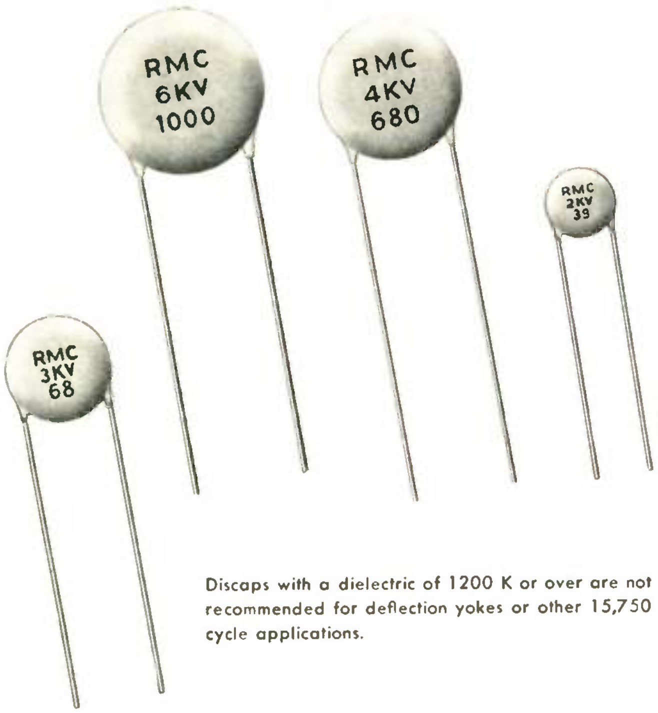
for exacting design...

RMC HIGH VOLTAGE DISCAPS

As a supplier of many types of ceramic capacitors to practically every major television manufacturer RMC can offer you high voltage DISCAPS that will consistently meet and exceed the most exacting design specifications.

Rated at 2000, 3000, 4000, 5000, and 6000 V.D.C., RMC high voltage DISCAPS provide the safety factor necessary in yoke and other critical voltage applications. They are the proved answer to problems encountered in the design of standard or special purpose electronic equipment.

Write today on your company letterhead about your specific requirements.



Discaps with a dielectric of 1200 K or over are not recommended for deflection yokes or other 15,750 cycle applications.

CAPACITY	DIELECTRIC	SIZE	AVAILABLE CAPACITY TOLERANCES
	2-KV		
331-470	1200-K	9/16"	±20% GMV
471-1000	1200-K	5/8"	±20% GMV
1001-2700	HI K	9/16"	GMV
2701-5000	HI K	3/4"	GMV
5001-10000	HI K	3/4'	GMV
	3-KV		
241-500	1200-K	5/8"	±20% GMV
501-1000	1200-K	5/8"	±20% GMV
1001-5000	HI K	3/4"	GMV
	4-KV		
181-680	1200-K	3/4"	±20% GMV
681-1000	HI K	5/8"	GMV
	5-KV		
131-330	1200-K	7/8"	±20% GMV
331-1000	HI K	7/8"	GMV
	6-KV		
101-220	1200-K	3/4"	±20% GMV
221-470	1200-K	7/8"	±20% GMV
221-1000	HI K	7/8"	GMV
471-1000	HI K	7/8"	GMV

POWER FACTOR: 1.5% Max. @ 1 KC (initial)
INSULATION: Durez phenolic—vacuum waxed

DISCAP CERAMIC CAPACITORS

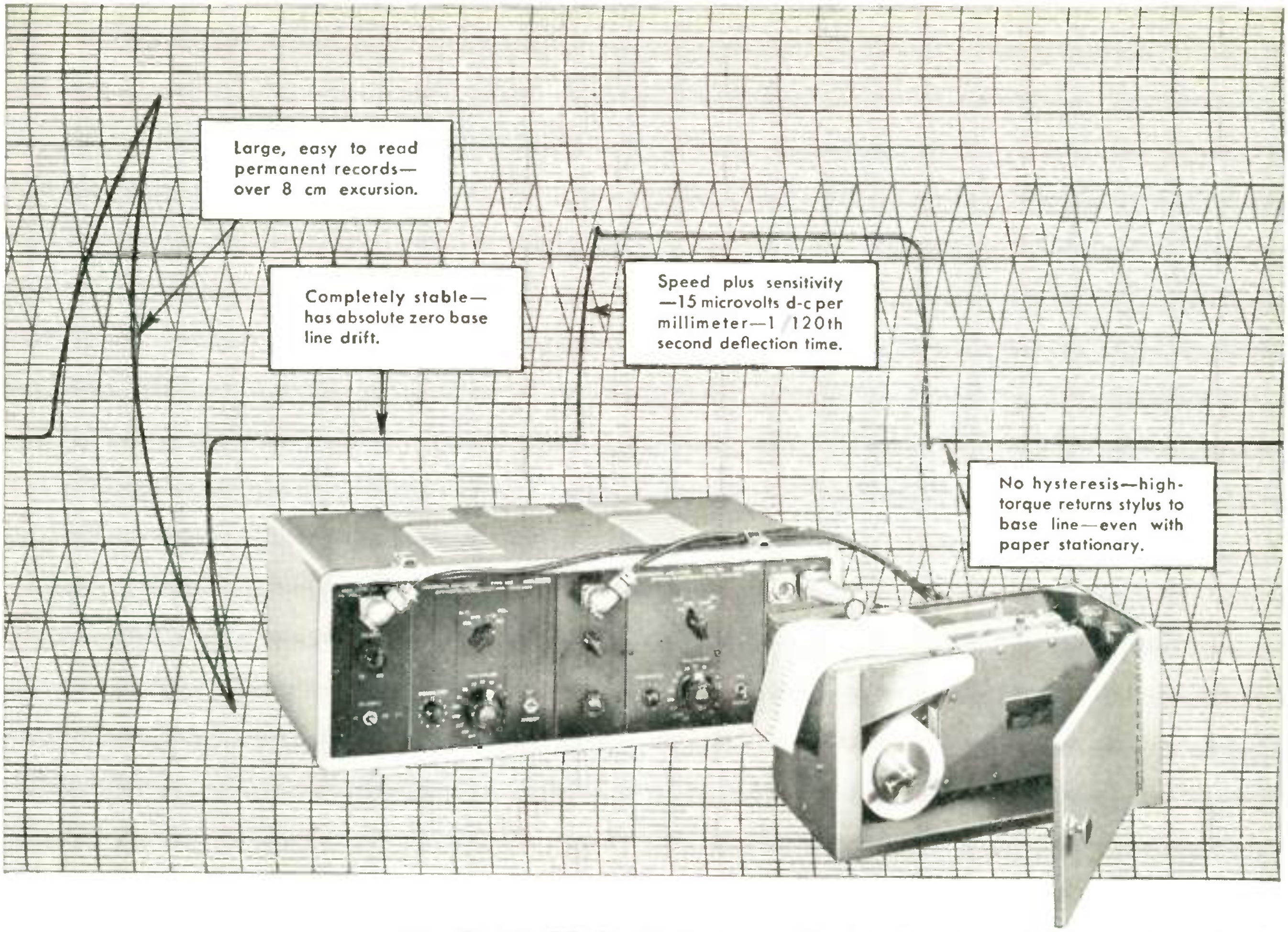
RMC

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



NEW...OFFNER PORTABLE DYNOGRAPH

This high speed, direct writing oscillograph recorder provides exceptionally high, absolutely stable, d-c or a-c amplification. It may be used with reluctance type gauges without auxiliary equipment. The exceptional stability, sensitivity, and versatility of the Dynograph are made possible by the exclusive, patented chopper amplifier. It is used for recording a wide variety of transient variables, such as strain, vibration, temperature, analog computer write-out, etc.

The performance specifications of the Type P Portable Dynograph are identical with those of the Type M Console model, but it is mounted in two convenient carrying cases as illustrated. The Type P is available with one or two channels.

Write for your copy of Bulletin L-742. It gives you complete details and application information on both portable and console models of the Offner Dynograph.

Check these exclusive features:

Speed plus sensitivity. The Dynograph gives you the maximum in speed and sensitivity—15 microvolts d-c per millimeter deflection with a response speed of less than 1/120th second.

Large easy-to-read records. Over 8 cm excursion, permitting the recording of large dynamic variations. Chart speeds 1 to 100 mm per second.

Absolute non-drifting stability. The Dynograph is absolutely stable and non-drifting—it is stable when it starts working and has absolutely zero base line drift.

No Hysteresis. The high-torque movement always returns the stylus to the base line—even with the paper stationary. Linearity is within one percent for four cm.

OFFNER ELECTRONICS INC.

5328 N. Kedzie Avenue, Chicago 25, U. S. A.

ALSIMAG®

SPECIAL PURPOSE CERAMIC MATERIALS

Most of our customers find the exact ceramic for their needs on property chart No. 551. This handy chart (sent free on request) gives mechanical and electrical data on the most frequently used AlSiMag materials. But this is by no means the full selection. AlSiMag offers the widest choice of ceramic materials in the industry. It isn't practical to list them all on one chart.

If you need a ceramic with unusual characteristics, tell us your requirements. We'll be glad to send you specification sheets (like those shown above) on the special purpose AlSiMag material that most nearly fills the bill. If we don't have the right material, perhaps we can develop one. Laboratory records from over 50 years of specialized experience often enable us to produce promptly a new composition "tailor made" for your needs.

Remember, the right ceramic can make all the difference in your product's performance. And, the best source for the right special purpose ceramic is AMERICAN LAVA CORPORATION. Write us today for detailed information on your requirements.

54TH YEAR OF CERAMIC LEADERSHIP

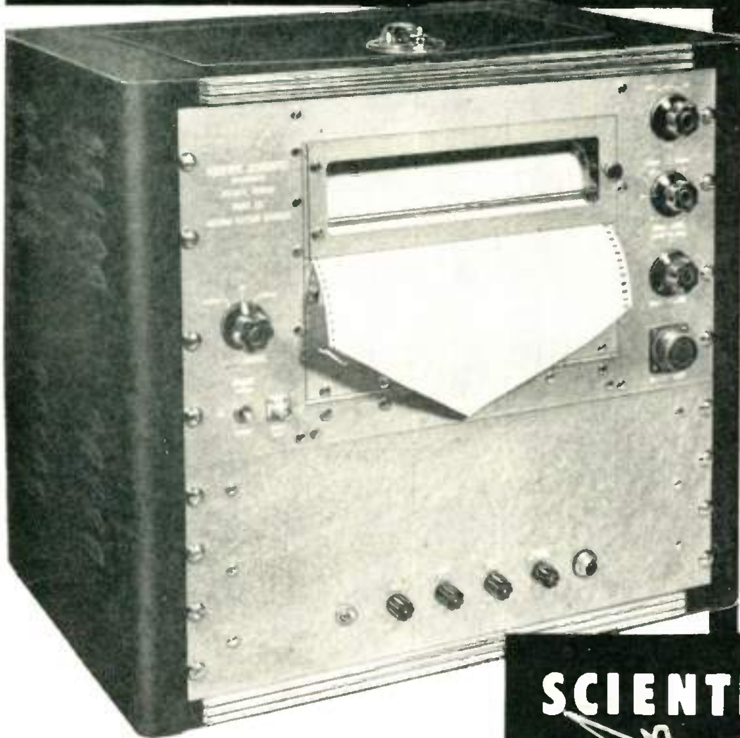
AMERICAN LAVA CORPORATION

CHATTANOOGA 5, TENNESSEE
A Subsidiary of
Minnesota Mining and Manufacturing Company

Branch offices in these cities (see your local telephone directory): Cambridge, Mass. • Chicago, Ill. • Cleveland, Ohio • Dallas-Houston, Texas • Indianapolis, Ind. • Los Angeles, Calif. • Newark N. J. • Philadelphia-Pittsburgh, Pa. • St. Louis, Mo. • South San Francisco, Calif. • Syracuse, N. Y. • Tulsa, Okla.
Canada: Irvington Varnish & Insulator Div. Minnesota Mining & Mfg. of Canada, Ltd., 1390 Burlington Street East, Hamilton Ontario, Phone Liberty 4-5735.

MODEL 121
Logarithmic ANTENNA PATTERN Recorder

An Automatic Precision Instrument for Obtaining Rapid, Accurate and Permanent Antenna Pattern Records



Model 121-A is available complete with bolometer amplifier.

SPECIFICATIONS

1. Chart Systems:

a. Input signals: Standard 1:1 and 36:1 output from Ketay 23CX6 Control Transmitters (or equivalent).

b. Chart speeds of 20, 120 and 720 inches per antenna revolution (360, 60 and 10 degrees per chart cycle).

c. Chart may be moved forward or backward manually or by synchro control; synchro switching is provided to reverse direction of chart transport with respect to direction of antenna rotation.

d. Maximum lag at chart speed of 4 inches per second is less than 0.3% of chart scale. Maximum chart speed is 6 inches per second.

2. Pen Systems:

a. Input signal: Approximately 1000 cycles per second, with an amplitude range of 0.0005 to 5 volts for full scale pen deflection, into an impedance of 1 megohm.

b. Pen position accuracy better than ± 0.25 db.

c. Pen deflection perpendicular to direction of chart motion.

d. Maximum writing speed greater than 40 inches per second.

e. Pen lifting device provided to prevent inking while positioning paper. Pens easily removed and replaced for changing ink color or refilling.

3. Mounting:

Recorder system is housed in a standard desk panel cabinet rack. Overall dimensions are 19 5/6 inches (height), 22 inches (width), and 14¾ inches (depth).

This time-saving instrument is priced considerably lower than any other existing machine of similar performance.

SCIENTIFIC ASSOCIATES, Inc.

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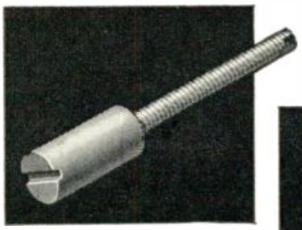
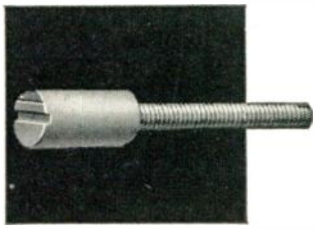
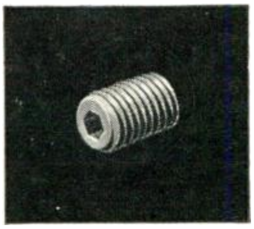


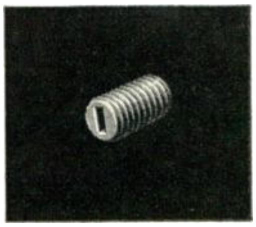


CUT IRON CORE COSTS

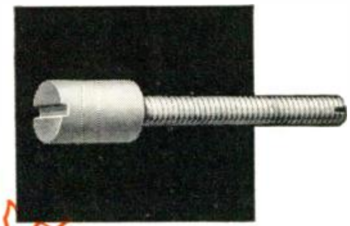
with Stackpole "PREFERRED TYPES"



"EE" SERIES . . . FOR ENGINEERED ECONOMY



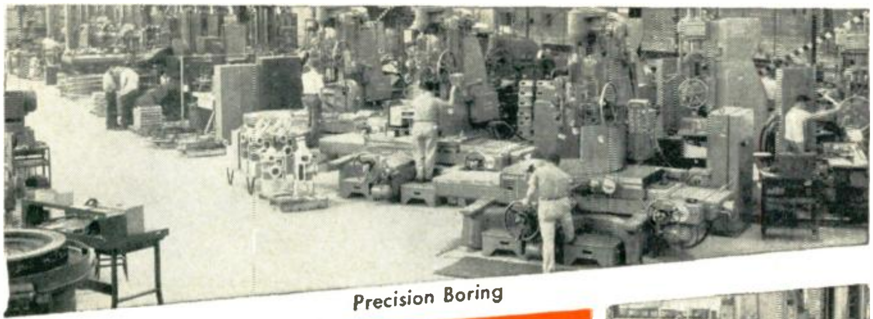
Made to well-known Stackpole quality standards, these new "EE" Cores are available only in commonly needed grades and sizes. They're ready for delivery from stock . . . at low prices . . . and without the usual set-up charge for custom-engineered cores. Mechanical specifications conform to the latest MPA recommendations. Electrical standards fully meet 8 out of 10 requirements of radio, TV, and communications equipment. Write, wire, or 'phone for details.



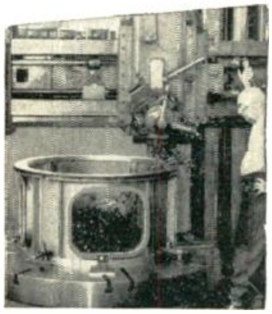
Electronic Components Division

STACKPOLE CARBON COMPANY
St. Marys, Pa.

STACKPOLE

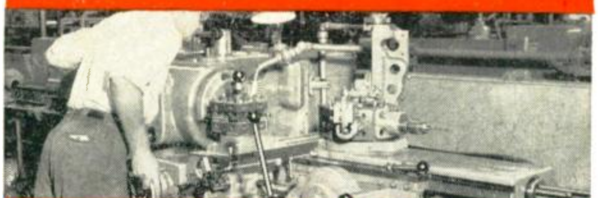


Precision Boring

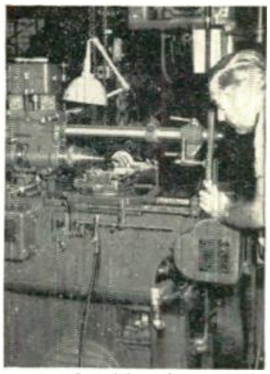


Precision Turning and Boring

Hobbing 56" Ring Gear



Precision Turret Lathe



Precision Gear Cutting

Industrial Engineering Notes

(Continued from page 32A)

have been announced recently by the Department of Defense. They cover the development of pocket-size radio equipment for the study of heart actions, a remote control system for jet aircraft, and the use of an airborne radar and radio wave propagation laboratory. The new flying laboratory was designed to permit study of a wide variety of problems in two fields of general interest to the radio engineer—radar target properties and radio wave propagation. The Naval Research Laboratory, Washington, D. C., instrumented the R5D aircraft, in which four radar sets are mounted in nacelles beneath the plane's wing. The plane's cabin contains four radar consoles, plus necessary control equipment, and high-speed motion picture cameras for recording data obtained during flights. Eight scientists and a crew of three are needed to man the aircraft. The new pocket-size radio equipment is expected to make a significant contribution to present knowledge of the heart and its functions. It was conceived by the Medical Corps, U. S. Navy, and developed under direction by the Aviation Medicine Division of the Naval Medical Research Institute, Bethesda, Md. The patient is supplied with a miniature radio the size of a package of cigarettes and a battery power supply of similar size. The miniature electronic device, weighing less than one pound and carried by the patient, picks up the heart waves, heart sounds and breath sounds and changes them to electrical impulses that can be handled by the transmitter. This in turn sends the audio information to the doctor's radio receiver located in his laboratory. The importance of the development, it was reported, lies in the fact that the patient can go about routine activities or take prescribed exercise while the small radio carried in his pocket sends continuous information to the physician. The remote control system for jet aircraft was developed under an Air Force contract by the Sperry Gyroscope Co. The Air Research and Development Command has reported that it can be adapted with ease to the control of either jet power or piston engine aircraft. The system provides automatic takeoff and landings, and exact split-second control at all times of the aircraft by radio and radar during climbs and dives, cruising, orbiting or other aerial maneuvers. The new drone aircraft control system also protects the aircraft in case all control signals are cut off by ground power failure or bomb damage while the plane is airborne. The announcement stated that in less than five seconds an electronic "brain" takes over control of the plane and forces it to begin a full-power climb of seven degrees, retracts dive flaps if these are extended, and at 200 mph changes to a climbing turn to the left until proper altitude is reached. In the new control layout, the Air Force said, special care was given to accessibility of sub-systems and components for ease of maintenance and to de-

(Continued on page 48A)

UNREGULATED LINE VOLTAGE

TYPICAL RECORDING FROM A LINE VOLTAGE VARIATION SURVEY: The chart at the top is representative of the average line voltage condition found in a comprehensive survey of commercial and industrial establishments. A tracing at the same point made at another time might show entirely different conditions, since periods of stable voltage are often followed by periods of violent transient fluctuations and/or large, gradual swings. The bottom chart, recorded at the same time, shows the output voltage of a Sola Constant Voltage Transformer fed from this line.

SOLA REGULATED LINE VOLTAGE

End Fluctuating Line Voltage Handicap to Reliable Product Performance

Where line voltage fluctuations impair the performance of voltage-sensitive electronic equipment, Sola Constant Voltage Power Transformers often provide a simple, economical solution. Stock or custom designed units are available.

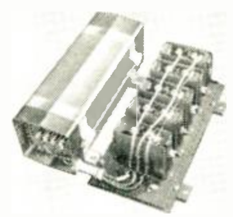
The Sola regulator has no moving parts and requires no manual adjustment or maintenance. Operation is automatic with response time 1.5 cycles or less. Regulates as close as ±1% with line voltage variations as great as 30%.

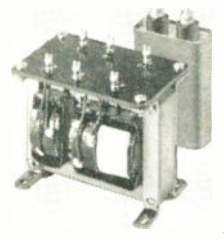
To meet special load requirements or service conditions, units can be custom-designed for production quantity orders. Variations available for special-order work include:

- *Capacity ranges from a fraction of a va to 30kva*
- *Wide variety of voltage inputs and outputs*
- *Frequencies other than 60 cycles*
- *Provisions for wide ambient temperature ranges*
- *Structural features for installation as a component*
- *Premium mechanical features for military service and other special duty*

The cost of Sola voltage regulation may be far less than you anticipate for two reasons: 1) installation of a Sola unit eliminates the need for the conventional non-regulating power supply transformer and any regulating components which you may currently use . . . 2) your requirements may be satisfied by a stock unit or custom design already on file. A Sola sales engineer is always available to discuss your voltage regulation requirements with you.

TYPICAL CUSTOM-DESIGNED SOLA POWER TRANSFORMERS





SOLA *Constant Voltage* TRANSFORMERS

Write for Bulletin 1 F-CVES for facts on the complete line of Sola Constant Voltage Power Transformers.

CONSTANT VOLTAGE TRANSFORMERS for Regulation of Electronic and Electrical Equipment • **LIGHTING TRANSFORMERS** for All Types of Fluorescent and Mercury Vapor Lamps. • **SOLA ELECTRIC CO., 4633 West 16th Street, Chicago 50, Illinois, Bishop 2-1414** • *BOSTON:* 272 Centre Street, Newton 58, Massachusetts • *NEW YORK 35:* 103 East 125th Street • *LOS ANGELES 26:* 2025 Sunset Boulevard • *PHILADELPHIA:* Commercial Trust Building • *CLEVELAND 15:* 1836 Euclid Avenue • *KANSAS CITY 2, MISSOURI:* 406 West 34th Street • *TORONTO 9, ONTARIO:* 617 Runnymede Road

Representatives in Other Principal Cities

Admiral has built COMMUNICATIONS for a sky-ful of planes
Standard communications equipment for nearly all military aircraft is the famed AN/ARC-27. This complex all-channel transmitter-receiver can be tuned to 1,750 VHF and UHF channels. Containing 56 tubes and upwards of 3,000 parts, this unit is being produced in vast quantities to keep pace with America's expanding air power. Approximately one out of every three transceivers completed to date has come out of Admiral plants.
This particular assignment provides ample evidence of Admiral's ability to produce in quantity...and to maintain the strictest quality standards. Production capacity has now been further augmented by means of the new automation equipment, designed and built by Admiral's own engineering staff. Address inquiries to:
Admiral®
CORPORATION
Government Laboratories Division
Chicago 47, Illinois
LOOK TO Admiral FOR
RESEARCH • DEVELOPMENT • PRODUCTION
in the fields of:
COMMUNICATIONS, UHF and VHF, air-borne and ground.
MILITARY TELEVISION, receiving and transmitting, air-borne and ground.
RADAR, air-borne, ship and ground.
RADIAC • MISSILE GUIDANCE • TELEMETERING • DISTANCE MEASURING • TEST EQUIPMENT • CODERS and DECODERS
Send for Brochure
...complete digest of Admiral's experience, equipment and facilities.
ENGINEERS! The wide scope of work in progress at Admiral creates challenging opportunities in the field of your choice. Write Director of Engineering and Research, Admiral Corporation, Chicago 47, Ill.

MICROWAVE SIGNAL GENERATORS

Complete coverage of the range 950-11,500 mcs /sec.

with Polarad single dial operation

Four new Microwave Signal Generators covering the range 950-11,500 mcs/sec. All with famous **Polarad single dial operation.** Each provides the maximum working range possible in one compact signal generator. And, additional Polarad Signal Generators are available to cover 12.4 to 39.7 kmc.

These features on all MSG units assure fast and simple operation: direct reading, single dial frequency control that tracks reflector voltages automatically . . . direct reading attenuator dial . . . conveniently placed controls, in logical sequence . . . high visibility on the face of each instrument.

Polarad Signal Generators are built to the same high standards required for military equipment. They are practical for the factory assembly line—engineered ventilation assures continuous and stable operation of all instrument functions. Components are readily accessible for easy maintenance. And laboratory accuracy is guaranteed under the most rigorous operating conditions.

Write directly to Polarad or your nearest Polarad representative for details.

	MSG-1	MSG-2	MSG-3	MSG-4
Frequency Range	950-2400 MCS/sec.	2150-4600 MCS/sec.	4450-8000 MCS/sec.	MSG 4 6950-10,800 MCS sec. | MSG 4A 6950-11,500 MCS/sec.
	(Frequency set by means of a single directly calibrated control)			
Frequency Accuracy	±1%	±1%	±1%	±1%
Power Output	1 MW	1 MW	.2 MW	.2 MW
Attenuator Range	120 db	120 db	120 db	120 db
Attenuator Accuracy	±2 db	±2 db	±2 db	±2 db
Output Impedance	50 ohms	50 ohms	50 ohms	50 ohms
Input Power	115V±10% 50-60 cps	115V±10% 50-60 cps	115V±10% 50-1000 cps	115V±10% 50-1000 cps
Internal Pulse Modulation: Pulse Width / Delay / Rate / Synchronization	0.5 to 10 microseconds 3 to 300 microseconds 40 to 4000 pulses per second Internal or external, sine wave or pulse		INTERNAL SQUARE WAVE (all models) Rate: 40-4000 cps Synchronization: Internal	
Internal FM: Type / Rate / Synchronization	Linear sawtooth 40 to 4000 cps Internal or external, sine wave or pulse			
Frequency Deviation	±2.5 MCS	±2.5 MCS	±6 MCS	±6 MCS
External Pulse Modulation: Polarity / Rate / Pulse width / Pulse separation	Positive or Negative 40 to 4000 pulses per second 0.5 to 2500 microseconds (For multiple pulses) 1 to 2500 microseconds			
Output Synchronizing Pulses: Polarity / Rate / Voltage / Rise time	Positive, delayed & undelayed 40 to 4000 pps Greater than 25 volts Less than 1 microsecond			
Size | Approx. weight	17" long x 13¼" high x 15½" deep | 60 lbs.		17" long x 15" high x 19½" deep | 100 lbs.	

"THE FINEST SIGNAL GENERATORS OF THEIR KIND" **Polarad** 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 Newton • Philadelphia • San Francisco • Syracuse • Washington, D. C. • Westbury • Winston-Salem • Canada, Arnprior, Toronto—Export: Rocke International Corporation

Industrial Engineering Notes

(Continued from page 44A)

sign simplicity for manufacturing in production quantities. It also was stated that the size and shape of numerous elements have been reduced in order to retain the standard Lockheed F-80C cockpit arrangement, especially for all flight safety and emergency controls. . . . **The National Bureau of Standards has announced the publication of a new volume containing descriptions and evaluations of three computational experiments in the general field of conformal mapping, which is of principal importance in electromagnetic theory, aircraft wing design and hydrodynamics.** The 61-page volume, "Experiments in the Computation of Conformal Maps," is available from the Government Printing Office, Washington 25, D. C., for 40 cents each. . . . **Since the National Bureau of Standards "Directory of Commercial and College Laboratories" has been withdrawn from circulation, an up-to-date revision of this publication now is available in the "Directory of Commercial and College Testing Laboratories," recently issued by the American Society for Testing Materials.** The society will undertake the responsibility for compilation and publication of the directory in the future, it was reported. The directory gives information regarding the location of testing laboratories together with the types of commodities and the nature of the investigations the laboratories are prepared to undertake. It is designed to be of assistance to the large number of purchasers who are not equipped to make their own acceptance tests and who therefore have hesitated to buy on specifications. It also should be of value to small manufacturers and others who are seeking testing laboratory services in the evaluation of raw materials and finished products. The new directory may be obtained from the American Society for Testing Materials, 1916 Race St., Philadelphia 3, Pa., at $1 a copy.

Television

The FCC has issued a notice of proposed rule making in which it requests information on the feasibility of authorizing "booster" UHF television stations. As defined by the FCC, a "booster" is an amplifying transmitter operating on the same channel as the main transmitter and simultaneously transmitting the same programs. As now proposed, a station authorization would cover the main transmitter and one or more boosters to fill "shadow" areas. The FCC already has authorized several booster operations on an experimental basis and said it is of the opinion that proceedings looking towards permanent authorization of UHF boosters now are warranted. For the purpose of determining whether its rules should be amended to permit such operation, the FCC said it seeks views about booster equipment and cost, any effect on color or monochrome

(Continued on page 50A)

Industrial Engineering Notes

(Continued from page 48A)

transmission, minimum separation between boosters and parent and other transmitters, interference safeguards, plans of proponents, hours of operation, whether boosters should operate unattended, minimum power and antenna height, whether their number should be limited as to areas and to main station, station identification and various technical considerations. . . . **The Federal Communications Commission has issued a request for comments on a proposal of the American Telephone & Telegraph Co. whereby the firm would provide a low-cost, off-the-air television program relay service.** The action is part of a rule making proposal instituted last September, wherein FCC invited comments on a proposal looking towards a review of the existing rules and policies on inter-city TV relay stations (Docket 11164). In commenting on that docket, AT&T had suggested that it might furnish program service by an off-the-air pickup arrangement in which it would relay the programs over common carrier facilities of a lower grade than normally employed in the nation-wide network service. The object of this suggestion would be to provide network TV programs in remote areas at lower cost.

Professional Group Meetings

Antennas and Propagation

Chicago Chapter—December 17

"The Application of Ferrite Rods as Used in Receiving and Other Types of Radio Antennas," by Frank Tatro, Henry L. Crowley Company.

Audio

Boston Chapter—March 10

"A Compact Loudspeaker System for Low Frequency Reproduction," by Edgar Villchur and Henry Kloss.

Chicago Chapter—December 17

"Circlotron Amplifier," by Howard Souther and A. M. Wiggins.

Cleveland Chapter—February 24

"The Circlotron Amplifier," by A. M. Wiggins, Electrovoice, Inc.

Automatic Control

Dallas-Fort Worth Chapter—February 24

"Mechanized Intelligence," by William Kiester, Bell Telephone Laboratories.

(Continued on page 56A)

International

Selenium Rectifiers

The WIDEST RANGE in the INDUSTRY

★Power Ratings from Microwatts to thousands of Kilowatts!

★Efficiency to 87%

The most widely used Industrial Power Rectifiers in industry today.

International Rectifier

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New York Office: 501 Madison Avenue, Phone PLaza 3-4942 • Chicago Office: 205 West Wacker Drive, Phone FRanklin 2-3889

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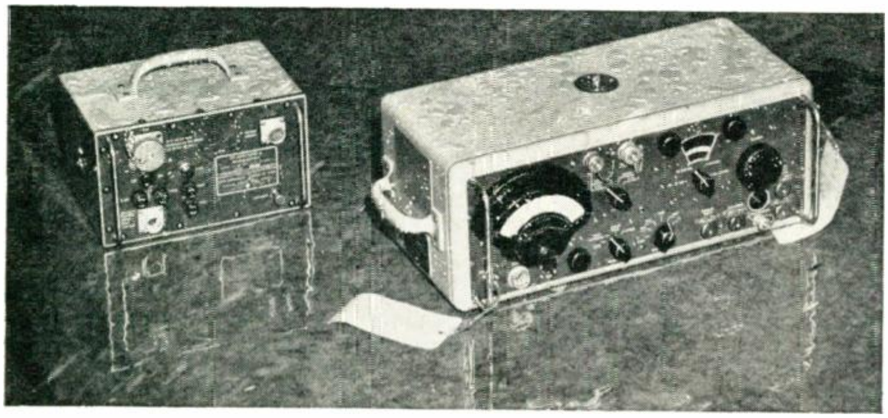
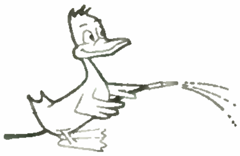
WORLD'S LARGEST SUPPLIER OF QUALITY INDUSTRIAL RECTIFIERS

Wet... but not worried!



Stoddart NM-10A Radio Interference and Field Intensity Meter, covering the frequency range of 14 kc to 250 kc.

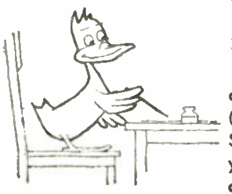
CHANCES ARE YOU'LL NEVER SUBJECT YOUR NM-10A TO THIS KIND OF TREATMENT...



BUT IT'S NICE TO KNOW THAT IT CAN TAKE IT IF IT HAS TO! We turned the hose on the AN/URM-6B — the Navy equivalent of the Stoddart NM-10A — in accordance with Navy specifications. Immediately afterward it was disassembled and found to be dry as a bone inside. And much to our satisfaction, the Navy inspector smiled!

Whether you use this fine, rugged instrument for field intensity measurements of carrier current systems, Navy, maritime or other services . . . or for surveys of conducted or radiated interference, you'll find that the *NM-10A CAN TAKE IT* . . . whether in the lab or in the field.

A complete selection of accessories is available, expanding the utility of the NM-10A to make it one of the most versatile instruments you have ever used.



Write today for further information. Learn about the excellent sensitivity . . . the "hand calibrated" accuracy . . . the sturdy, dripproof construction, enabling use in driving rain or snow . . . the A. C. power supply that permits operation from 105 to 125 volts or 210 to 250 volts A. C., 50 to 1600 cps.

The NM-10A is the identical instrument we supply to the Navy as the AN/URM-6B, a Class One instrument, as shown in MIL-I-16910 (SHIPS). It was designed and is manufactured exclusively by Stoddart Aircraft Radio Co., Inc. When you buy the NM-10A you're getting a quality instrument that meets the rugged requirements of the U. S. Navy!

Stoddart RI-FI Meters Cover the Frequency Range of 14 kc to 1000 mc

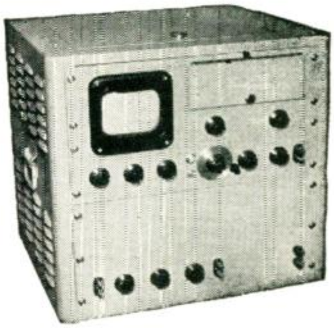
NM-20B — HF
15 kc to 25 mc. Commercial equivalent of AN/PRM-1A. Self-contained batteries. A. C. supply optional includes standard broadcast band, radio range, WWW and communications frequencies. Has BFO.

NM-30A — VHF
20 mc to 400 mc. Commercial equivalent of AN/URM-47. Frequency range includes FM and TV bands.

NM-50A — UHF
375 mc to 1000 mc. Commercial equivalent of AN/URM-17. Frequency range includes Citizens band and UHF TV band.

STODDART AIRCRAFT RADIO Co., Inc.
6644-C Santa Monica Blvd., Hollywood 38, California • Hollywood 4-9294

SPEED UP



AUDIO WAVE FORM ANALYSIS

PANORAMIC SONIC ANALYZER LP-1

Many engineers find that Panoramic's LP-1 expedites their entire measurements program. LP-1 analyzes sound vibrations and electrical waveforms quickly, conveniently, accurately. Designed to eliminate the tedious problems commonly associated with audio waveforms analysis, the Panoramic technique provides valuable visual information *in seconds!*

- Visualizes frequency and amplitude of waveform components between 40 and 20,000 cps; magnifies small portions of spectrum for detailed analysis; displays easily photographed; scans spectrum in 1-second; analyzes changing and static phenomena.

It will pay you to investigate the many unique advantages of LP-1.

- **SPECIAL APPLICATIONS**
- Investigations of closely spaced sound and vibration frequencies. Harmonic analysis of waveforms having low frequency fundamentals. Spectrum analysis requiring constant band width.
- Panoramic's LP-1 offers scores of unique advantages; it will pay you to check their application to your problems.

WRITE TODAY for Complete Specifications

PANORAMIC RADIO PRODUCTS, INC.

Made by the makers of Panadaptor, Panalyzor, Panoramic Sonic Analyzer, and Panoramic Ultrasonic Analyzer

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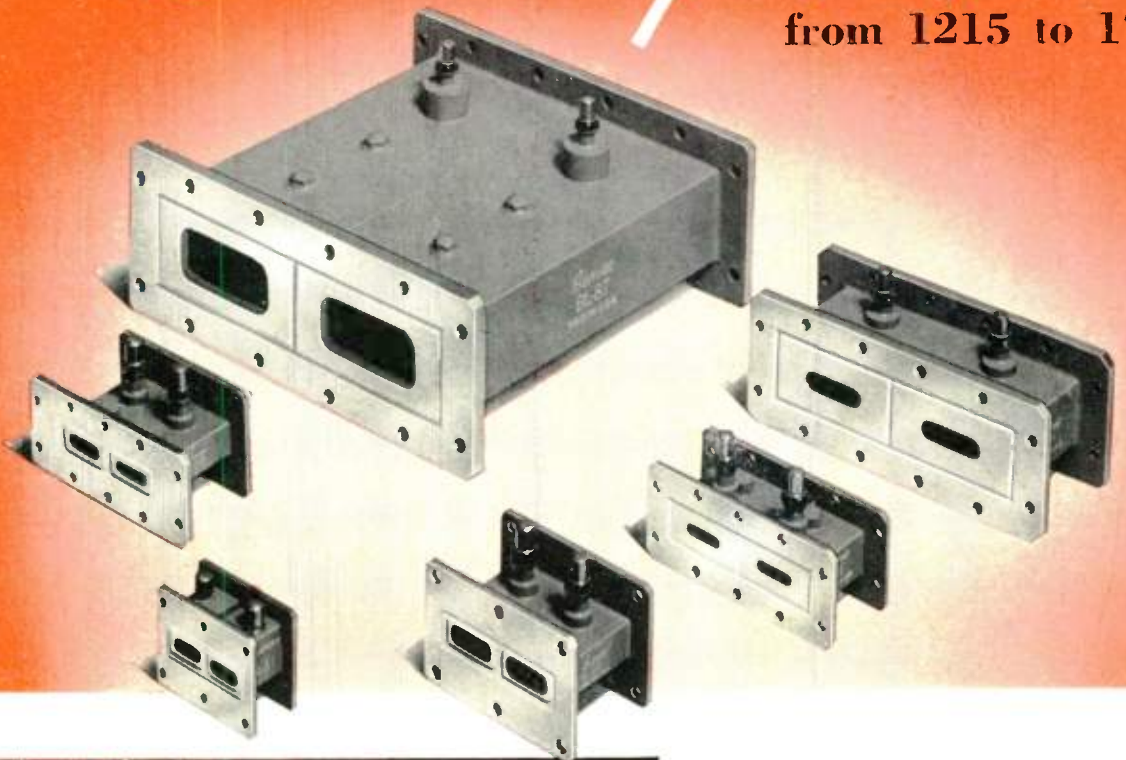
NOW... *Bomac* / dual TR **MICROWAVE** tubes covering the frequencies from 1215 to 17,000 mc.



PAT. APPLIED FOR

DUAL TR AND ATR TUBES

Band	Frequency	Center Frequency	Power Level (KW Max)	Type	Description	Tube Designation
L	1215-1355	1285	2000	TR	Dual, Band Pass	6634/BL90
S	2600-3000	2800	750	TR	Dual, Band Pass	6636/BL87
S	2650-2950	2800	750	ATR	Dual	BL92
C	5400-5900	5650	700	TR	Dual 5865	6640/BL60
C	5400-5900	5650	3000	TR	Dual 6568, Band Pass	BL613
C	5400-5900	5640	300	ATR	Dual, Fixed-Tuned, Contact Mount	BL63
X	8490-9578	9000	200	TR	Dual 1B63A	6334/BL27
X	8490-9578	9000	200	TR	6334 tapped Flanges Both Ends	BL78
X	8490-9578	9000	200	TR	6334 plus Separate Channel 1B63A	BL81
X	8490-9578	9000	200	TR	6334 with Recovery Time 1 μsec at 1 db	6643/BL84
X	8490-9578	9000	200	TR	BL84 with Heater	BL84H
X	8490-9578	9000	200	TR	6334 with Large X Flange Input and Small X Flange Output	6642/BL600
X	8490-9578	9000	200	TR	Dual 1B63A, 2 μsec Recovery Time	6646/BL604
X	8490-9578	9000	200	TR	BL604 with Heater	6647/BL604H
X	8490-9578	9000	200	TR	6334 with Special Hole Dimensions for Aluminum Flanges	BL607
X	8490-9578	9000	200	TR	6334 with Special Saddle Type Flange	6648/BL615
X	8500-9600	9050	250	TR	Large X Guide	6501
X	8500-9600	9050	250	TR	Four-Element Tube for Large X Guide	6564/BL71
X	8490-9578	9000	250	—	Integral 6334 Hybrid Duplexer	BL507
Ku	15000-17000	16000	100	TR	Dual, Band Pass	6560/BL35

Unlimited design possibilities are now afforded the systems engineers with Bomac's complete line of dual TR tubes.

Bomac dual tubes are designed for use with suitable short-slot hybrid junctions to provide balanced duplexers of utmost simplicity.

In operation, the balanced duplexer is similar to magic-T or rat race duplexers. The very low coupling between the transmitter and antenna eliminates the necessity of employing one or more ATR tubes and reduces the losses introduced in radar systems by these tubes. The features of compactness, weight and excellent performance with respect to both transmission and reception characteristics offered by these dual tubes hold many advantages for microwave systems designers and engineers.

Bomac offers a complete line of hybrids to be used in conjunction with these dual tubes — the tubes can be supplied with integral shutters, offering continuous crystal protection.

We invite your inquiries regarding

- ENGINEERING
- DEVELOPMENT
- PRODUCTION

Bomac Laboratories, Inc.

BEVERLY, MASSACHUSETTS

GAS SWITCHING TUBES, TR, ATR and Pre-TR · DUAL TR and ATR TUBES · SILICON DIODES · WAVEGUIDE SWITCHES
REFERENCE CAVITIES · MAGNETRONS · PRESSURIZING WINDOWS · SHUTTER TUBES · HYDROGEN THYRATRONS
REFLEX KLYSTRONS · TRAVELING WAVE AMPLIFIER TUBES · SYSTEMS

Catalog on request. Write (on your company letterhead) Dept. P-6 BOMAC Laboratories, Inc. Beverly, Mass., or phone Beverly 6000.

A new source of precision instrumentation

...members of a family

Wave Analyzer • Model 20
Fourier amplitude analysis of complex waveforms. Direct percentage readings for components from 50 cps to 50 kcs. Extremely portable, minimum size and weight.

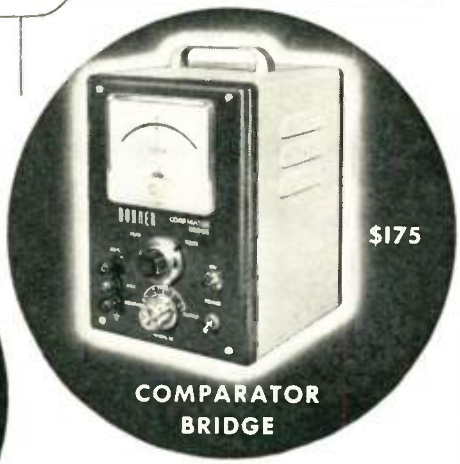
Video Probe Amplifier • Model 50
Zero loss accessory for all oscilloscopes, permitting circuit examination with only 2 mmf loading. Frequency range 5 cps to 12 mcs. Serves also as 40 X video amplifier.

Laboratory Oscillator • Model 10
Six decade RC oscillator, 2 cps to 2 mcs. Stable operation, low distortion, 20 volts output.

Comparator Bridge • Model 60
Extremely precise potentiometric voltmeter, internal standard accurate to 0.05%. Digital direct reading at null balance. Double utility, serving also as a precision voltage source.

DONNER SCIENTIFIC COMPANY

2829 Seventh St., Berkeley, California
All prices f.o.b. Berkeley, California
Write for technical bulletins —specify model number!



simplify custom FILTER installation

The 4200 Variable Filter and 4201 Program Equalizer are now available in component form, as illustrated, for the custom builder.

In addition to the flexibility of installation, all the features and characteristics of the standard models are retained.

The high and low sections of either model may be obtained separately. Complete wiring instructions included.

Send for Bulletin TB-4

Model 4200 Variable Filter
(Send for Bulletin S)

Model 4201, Program Equalizer
(Send for Bulletin E)

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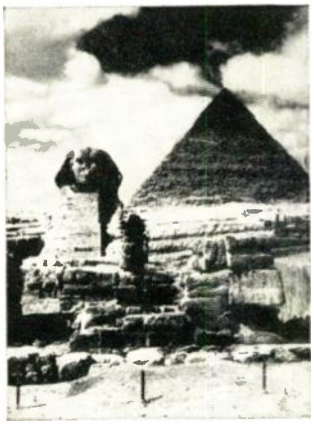
PYRAMID SOLID DIELECTRIC GLASSEAL® CAPACITORS FOR
6 POINT PREFERENCE



1 Especially sturdy capacitors capable of withstanding vibrational stresses of high acceleration and frequency, and severe shock conditions encountered in guided missiles and airborne equipment.

2 Utilize new, rugged compression-seal type, glass-to-metal solder-seal terminals. Terminals will *not* work loose or rotate under *any* operating condition.

3 Functional operating range from —55°C to +125°C.

4 Operates normally under severe humidity conditions.

5 Production tests for voltage breakdown, capacitance, power factor, insulation resistance and seal are performed on a 100% basis.

6 Capacitance range: .001 mfd. to 1.0 mfd.; voltage range: 100 to 600 V.D.C. operating; can be provided to standard tolerance of ±20% or to closer tolerances, if desired.

Burton Browne/New York

+ FACTORS

1. Hermetically sealed in metallic cases.

2. Power factor less than 1%.

3. Subminiature in size.

4. Available in both inserted tab and extended foil constructions.

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CAPACITOR DIVISION
PYRAMID ELECTRIC COMPANY
1445 Hudson Blvd. • North Bergen, N. J.

Professional Group Meetings

(Continued from page 50A)

Broadcast and Television Receivers

Chicago Chapter—February 18
"What is a Good Tube?" by Carl Atkins.

Chicago Chapter—January 21
"A Case History of the Application of a TV Booster Station for Improved VHF Reception," by Wendell C. Morrison, RCA Laboratories.

Circuit Theory

Los Angeles Chapter—March 10
"Feedback System Design," by R. A. Burt, Electronic Systems, Inc., and J. A. Aseltine, Lear, Inc.

Component Parts

New York Chapter—March 3
"A Forward Look in Tantalytic Capacitors," by D. F. Warner, General Electric.
"Recent Advances in Pulse Forming Networks," by A. Graydon and L. Swett, both of General Electric.

Electron Devices

Washington Chapter—February 28
"The Magnetic Amplifier," by Donald G. Scorgie, Naval Research Lab.

Electronic Computers

Washington Chapter—March 2
"Magnetic Recording Devices for Business Machines," by S. J. Begun, Brush-Clevite Development Company.

Washington Chapter—February 5
"The Lattice Videotron," by F. B. Maynard, National Union Electric Corporation.
"The Anodyne Tube-A Versatile Beam Deflection Device," by H. J. Wolkstein, National Union Electric Corporation.

Washington Chapter—December 1
"Diode Amplifiers," by A. Holt, NBS.

Engineering Management

Boston Chapter—March 10
"Railroading Today and Tomorrow," by Patrick B. McGinnis, New York, New Haven, Hartford Railroad.

Chicago Chapter—January 21
"Management from the Viewpoint of the Armed Services," by Lt. Col. B. L. Mathews.

San Francisco Chapter—March 8
"The Intangible Factors in Engineering," by Russell H. Varian, Varian Associates.

(Continued on page 60A)

hitch your missile to a star...

Navigation and Control Devices PRODUCED for Missiles and Aircraft

Kollsman has designed, developed and produced the following navigation and control systems and components:

FOR NAVIGATION OR GUIDANCE

Photoelectric Sextants for remote semi-automatic celestial navigation. Flight tests in jet aircraft gave an overall geographical positional accuracy of two miles.

CLASSIFIED **Automatic Astrocompasses** for precise automatic celestial directional reference and navigation.

Photoelectric Tracking Systems For many years Kollsman has specialized in high precision tracking systems.

Periscopic Sextants for manual celestial observations.

CLASSIFIED **Computing Systems** to provide precise data for automatic navigation and guidance, operated by optical, electromechanical, and pressure sensing components.

Photoelectric Sextant

FOR CONTROL

proven components now in production

Pressure Pickups and Synchrotel Transmitters

to measure and electrically transmit • true airspeed • indicated airspeed • absolute pressure • log absolute pressure • differential pressure • log differential pressure • altitude • Mach number • airspeed and Mach number.

Pressure Monitors — to provide control signals for altitude, absolute and differential pressure, vertical speed, etc.

Acceleration Monitors — for many applications now served by gyros.

Pressure Switches — actuated by static pressure, differential pressure, rate of change of static pressure, rate of climb or descent, etc.

Motors — miniature, special purpose, including new designs with integral gear heads.

SPECIAL TEST EQUIPMENT

optical and electromechanical for flight test observations.

Please write us concerning your specific requirements in the field of missile or aircraft control and guidance.

Technical bulletins are available on most of the devices mentioned.

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Engineering

THE RECORD-SETTING G-12A

This is the JK Glasline unit providing stability of 1 part in 10^9 over a 24 hour period, stability that requires the most accurate methods of measurement available.

TO COMPLETE THE ENVIRONMENTAL PICTURE: THE NEW JK09 OVEN

To the protection JK Glasline design provides against moisture, contamination, vibration and barometric pressure, the JKO9 oven adds control of temperature. It is production tooled for economy and uniformity, is small and light (1.28" x 1.70" ; 1.5 oz.), and is capable of maintaining a temperature constant to within ±1°C *over a range of* −55° *to* +100°C. Here is an oven that matches the performance of many, massive multi-stage heaters — an example of JK's ultra-stable miniaturization program.

JK-G9: Precision "Glasline" quartz crystals, sealed in evacuated glass for cleanliness and protection, over a complete range of 800 cycles to 5 mc.

PRODUCTS

THE JAMES KNIGHTS COMPANY

Crystals for the Critical - Sandwich, Illinois

Facilities That Measure to 1 part in 10^9 Back JK Research, Production and Certification.

Just a short time ago a crystal offering the extreme stability of the JK G-12A was viewed as a scientific curio. Even its measurement was a scientific effort; its first use limited to laboratory equipment.

But today JK Glasline crystal units in these exacting stability ranges are meeting the demand for ultra stable frequency control in many field applications. They are being production engineered, manufactured, tested, and performance certified in commercial production quantities for use as precision time base units, as long range navigational aids, and for spectrum conserving communications systems. They provide unprecedented stability and reliability, with the compactness and environmental protection that today's field equipment requires.

These advanced research, production and testing facilities are important to everyone concerned with frequency management problems because they are the keys that can unlock tomorrow's door for you. The control of the quartz crystal, so long an art, has become a science.

JK-G4: "Glasline" Crystal Filter Resonator. For broad filter applications such as power line carrier communications and telemetering. Frequency range 50 to 200 kc.

(ACTUAL SIZE)

JK-G3: 10 mc to 150 mc. Miniature size, minimum aging drift, high Q for maximum performance. JK miniaturized "Glasline" crystals meet the growing need for minimum size with maximum stability.

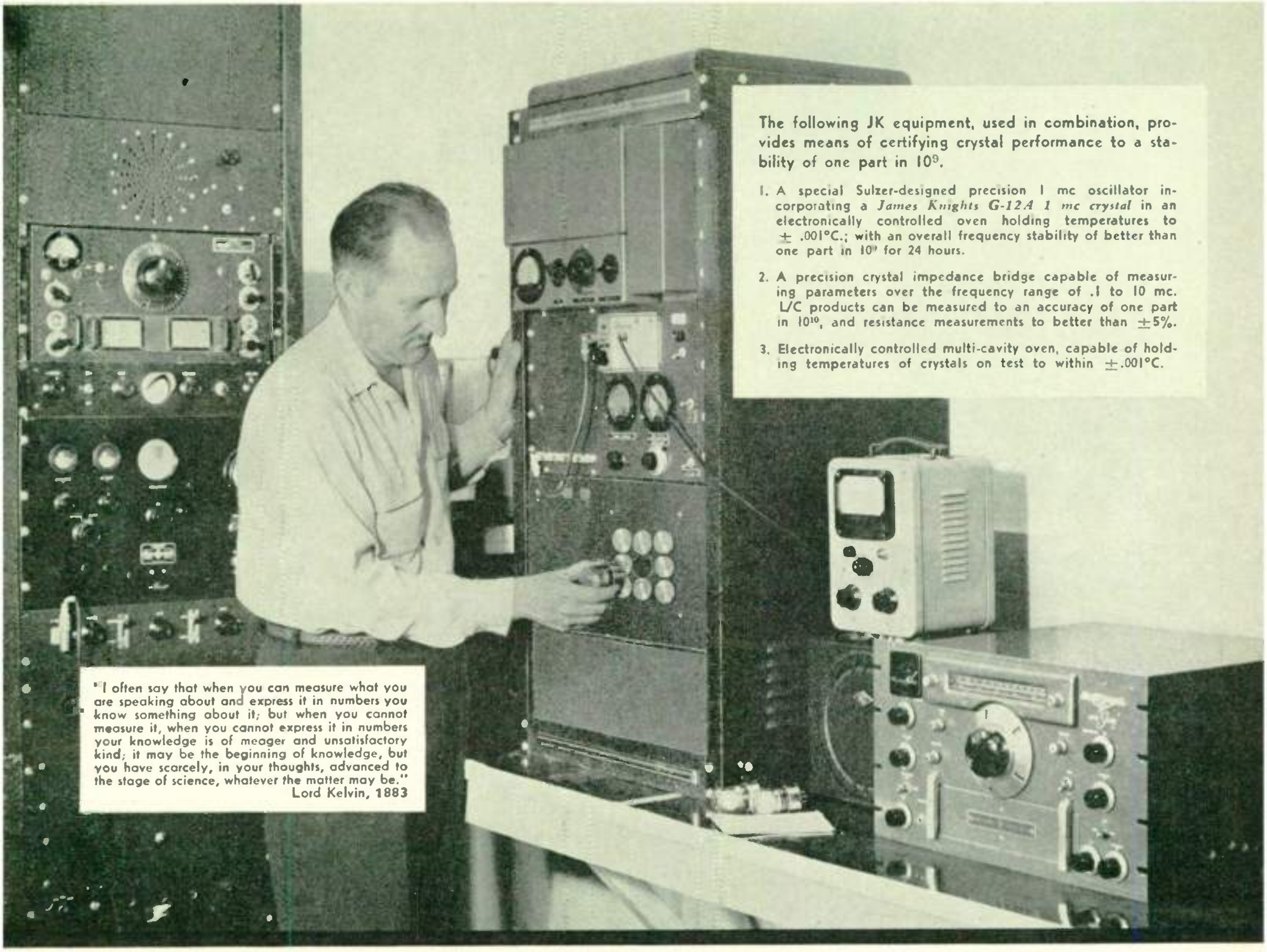
Ultra-Precision frequency control requires that crystal oven, power supply and oscillator circuitry be compatible in design and construction for optimum precision and reliability. The James Knights Company will combine their precision crystals and ovens with tried and proven circuitry in packages that will meet your mechanical layout requirements, at a saving of valuable customer engineering time. Precision frequency signal sources covering the wide frequency range of 60 cycles to 150 mcs can be made available to meet your specific application.

JK "THERMYSTAL"

AN ADVANCED FREQUENCY CONTROL DESIGN

1. **Higher Merit Factor:** Vacuum enclosure increases Q of crystal.
2. **Calibration Accuracy:** ± (1 cycle + .0001%)
3. **Temperature Stability:** 30 to 900 kc ± .0001%, 1000 kc to 150 mc ± .00005%. Oven temperature varies less than ± 1°C over ambient range of -55°C to + 85°C.
4. **Secular Stability:** Less than .001% per year. Crystal is specially processed and sealed in glass enclosed vacuum.
5. **Low Oven Power:** 6.3 v @ 1.5 amp maximum. Thermostat cycles less than 3 times per minute at room temperature.

that unlocks tomorrow's door



The following JK equipment, used in combination, provides means of certifying crystal performance to a stability of one part in 10^9.

1. A special Sulzer-designed precision 1 mc oscillator incorporating a *James Knights G-12A 1 mc crystal* in an electronically controlled oven holding temperatures to ± .001°C.; with an overall frequency stability of better than one part in 10^9 for 24 hours.
2. A precision crystal impedance bridge capable of measuring parameters over the frequency range of .1 to 10 mc. L/C products can be measured to an accuracy of one part in 10^{10}, and resistance measurements to better than ±5%.
3. Electronically controlled multi-cavity oven, capable of holding temperatures of crystals on test to within ±.001°C.

"I often say that when you can measure what you are speaking about and express it in numbers you know something about it; but when you cannot measure it, when you cannot express it in numbers your knowledge is of meager and unsatisfactory kind; it may be the beginning of knowledge, but you have scarcely, in your thoughts, advanced to the stage of science, whatever the matter may be."
Lord Kelvin, 1883

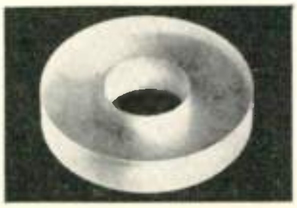
JK STANDARD MILITARY AND COMMERCIAL TYPE CRYSTALS

Quality and service are James Knights traditions that apply to the simplest JK crystals as well as to the most complex, and apply to our smallest customer as well as our largest. So whatever your requirements — look to James Knights as a dependable, cooperative source for quality, price and delivery.

TEMPERATURE CONTROL OVENS: Available for a wide range of applications.

PRESSURE MOUNTED: A complete line of commercial and military types.

MILITARY TYPES: Hermetic sealed, metal cased, in frequency ranges from 16 kc to 100 mc.



ULTRA-SONIC TRANSDUCERS: Carefully oriented and processed to your specifications, in a variety of shapes with holes, dimples, soldered-on leads, and backing plates. Can be plated with a variety of metals.

Professional Group Meetings

(Continued from page 56A)

Microwave Theory and Techniques

Northern New Jersey Chapter—February 16

"The Nature of Ferro-Magnetism in Ferrites," by J. Rowen, Bell Television Laboratories.

Philadelphia Chapter—February 16

"Antenna Systems for UHF TV Boosters," by Bruce Rankin, David Sarnoff Laboratories, RCA.

Nuclear Science

Oak Ridge Chapter—February 16

"Storage Devices for Digital Computers," by R. J. Klein, ORNL.

Boston Chapter—February 10

"Scanning of Radioactive Isotopes Within the Human Body," by Gordon L. Brownell, Massachusetts General Hospital.

Telemetry and Remote Control

Los Angeles—March 15

"The APOTA Automatic Tracking Antenna System," by James E. Palmer, Sandia Corporation.

"Modifications to the RDB Telemetering Standards Under Consideration by the Inter-Range Panel," by R. E. Perry, US NAMTC.

Vehicular Communications

Detroit Chapter—February 16

Each of the following speakers discussed maintenance and performance problems and techniques common to their systems: T. P. Rykala, Michigan Consolidated Gas Company; F. M. Hartz, The Detroit Edison Company; O. L. Santi, Michigan Bell Telephone Company; J. E. McFatridge, Detroit Police Radio Bureau; F. L. Kahle, Mobile Communication Services.

Section Meetings

Akron

"The Dawn of the Age of Space Travel," by D. C. Romick, Goodyear Aircraft Corp.; March 15, 1955.

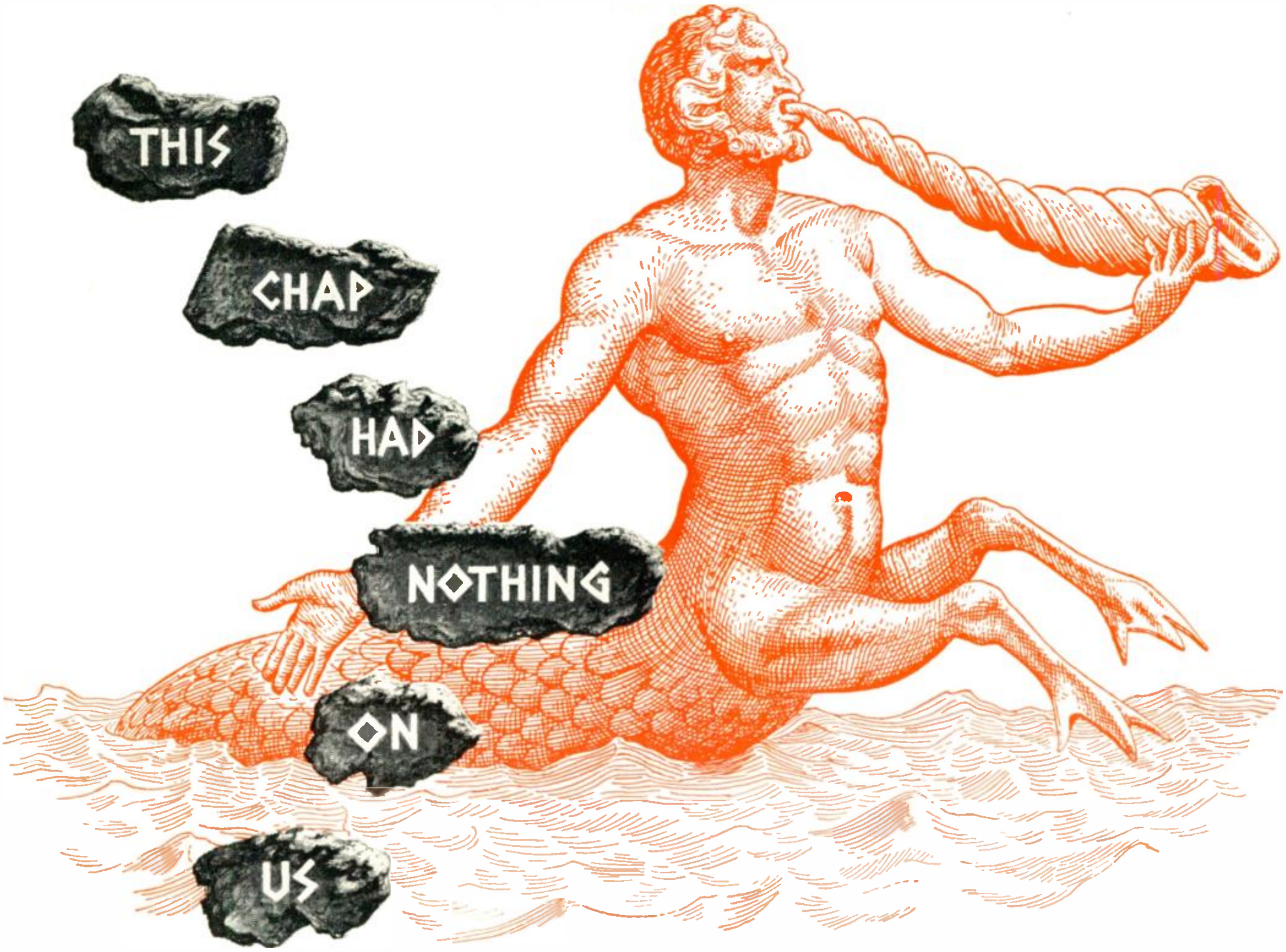
Albuquerque-Los Alamos

"Digitel—A Telemetering Data Digitizer," by G. R. Bussey, Sandia Corp.; April 15, 1955.

Baltimore

"Ferro-Electrics as Information Storage Media," by Prof. C. F. Pulvari, Catholic University of America; April 13, 1955.

(Continued on page 62A)



He's Proteus, Neptune's son, a highly versatile character—hence the adjective *protean.* You never knew whether next you'd see him as a lion, a raging fire, a reptile, or an angry bull.

Here at Driver-Harris, we do protean marvels too . . . with *metals.* For instance, Nichrome*, the unique heat-resistant, corrosion-resistant, electrical-resistance alloy known the world over, has long been the engineer's yardstick of comparison not in one, but *in at least 3 widely different applications.*

TO GENERATE HEAT: In all applications of producing heat by electricity, particularly to temperatures above 1700°F., Nichrome and Nichrome V set the quality standard. From simple electrical appliances such as ranges, broilers, toasters, etc. to giant industrial furnaces, no other alloys enjoy such widespread recognition and use.

TO RESIST HEAT: Because of its unsurpassed resistance to heat and corrosion, Nichrome is used for making massive furnace muffles and retorts often weighing tons, and work-loading fixtures of all shapes. The outstanding property of Nichrome here is its extremely long life, which results in low heat-hour costs.

TO RESIST ELECTRICITY: The greatest contribution to outstanding stability and miniaturization of resistors is made by Nichrome wire. Drawn to sizes as small as .0005 and finished in a variety of insulations, Nichrome is a boon to electronics in the manufacture of high tolerance resistance units of all types.

Added to the nickel and chrome of Nichrome and Nichrome V is always one exclusive ingredient—the supreme mastery of the Driver-Harris specialists, gained in their 57 years of melting and drawing experience.

In recognition of its unique properties, the United States Patent Office in August, 1908, granted solely and exclusively to us the trademark NICHROME. There is *only* one Nichrome, and it is made only by Driver-Harris.

*T. M. Reg. U. S. Pat. Off.

SPECIAL PURPOSE ALLOYS SINCE 1899 — D H — TRADE MARK REG. U. S. PAT. OFF.

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BRANCHES: Chicago, Detroit, Cleveland, Louisville, Los Angeles, San Francisco

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MAKERS OF THE MOST COMPLETE LINE OF ELECTRIC HEATING, RESISTANCE, AND ELECTRONIC ALLOYS IN THE WORLD

Section Meetings

(Continued from page 60A)

BEAUMONT-PORT ARTHUR

"Transistor Application in Broadcast Radio," by Roger Webster, Texas Instruments, Inc.; March 30, 1955.

BINGHAMTON

"Application of Glass to Electronic Components," by W. H. McKnight, Corning Glass Works; April 11, 1955.

BUFFALO-NIAGARA

"The TimeAlyzer," by K. R. Wendt, Wendt-Squires, Inc.; March 16, 1955.

CEDAR RAPIDS

"Predicted Wave Signaling Systems," by Donald Martin, Collins Radio Company; March 16, 1955.

"Scattering of Radio Waves," by Dr. R. L. McCreary and "Aircraft Flight Simulator," by W. T. Hedgcock, both of Collins Radio Company; April 13, 1955.

CENTRAL FLORIDA

"Horizontal and Vertical Organization in Engineering Management," by Charles Marshall, WPAFB; February 18, 1955.

CHICAGO

"The Bell Solar Battery," by Dr. G. Raisbeck Bell Telephone Labs.; March 17, 1955.

CLEVELAND

"Class B Operation of Audio Frequency Junction Transistors," by K. E. Loofbourrow; RCA Tube Division; March 31, 1955.

CONNECTICUT VALLEY

"Hearing Processes," by Dr. W. E. Kock, Bell Telephone Lab.; March 17, 1955.

DAYTON

"Radio Astronomy," by Donn Van Stoutenberg, Ohio State University; April 10, 1955.

DETROIT

"Recent Developments in Ferroelectric and Ferromagnetic Ceramics and Their Application as Nonlinear Circuit Elements," by Dr. H. W. Welch, Jr., University of Michigan; March 18, 1955.

ELMIRA-CORNING

"Color Television vs. Color Movies," by D. G. Fink, Philco Corp.; February 21, 1955.

"Synthetic Mica," by Dr. W. G. Lawrence, Alfred University; March 14, 1955.

EMPORIUM

"Radio Astronomy and the Sun," by O. C. Coates, Naval Research Lab.; March 22, 1955.

"Television Pick-up Tubes for Black-and-White and Color," by Dr. P. K. Weimer, RCA; April 19, 1955.

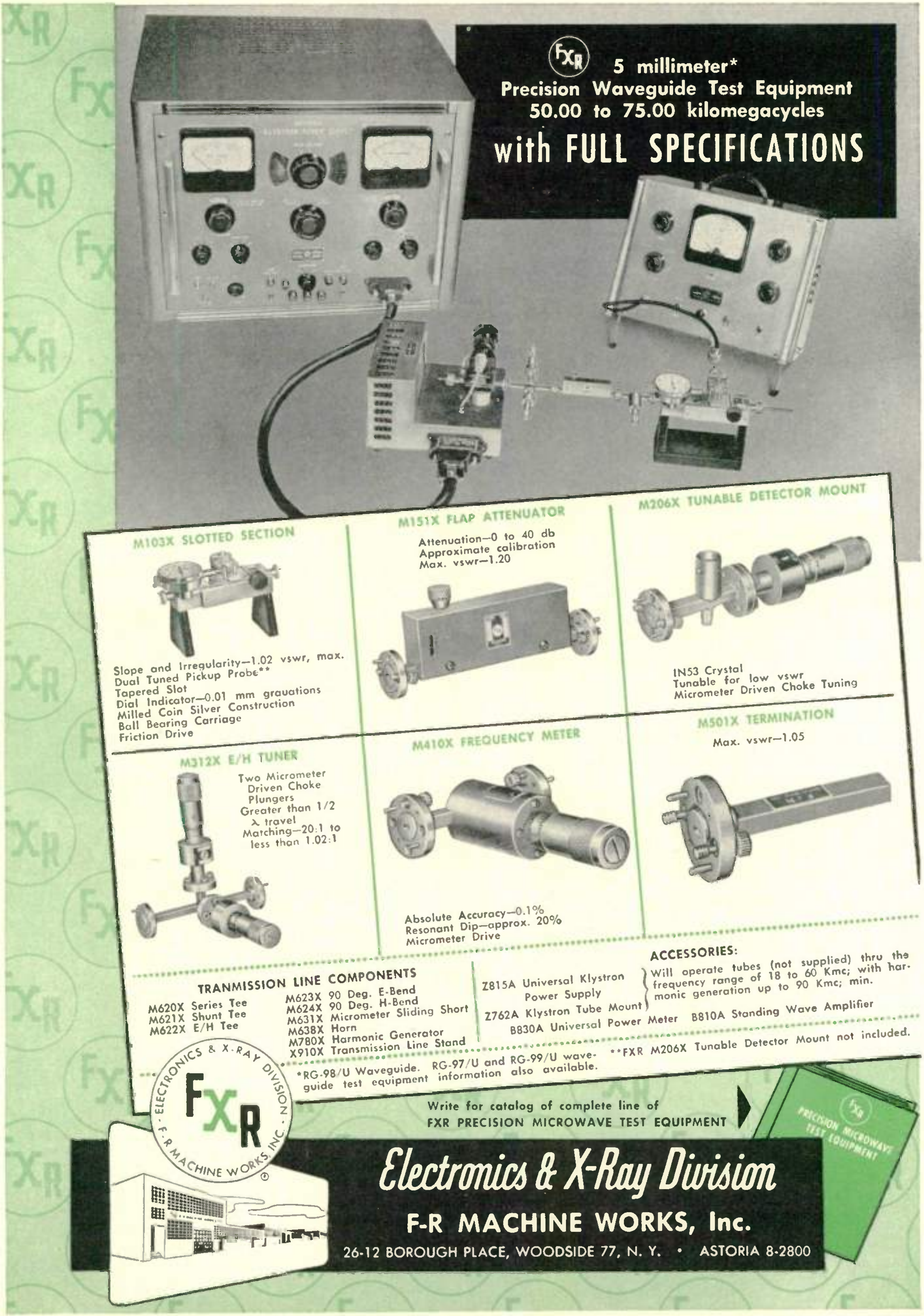
EVANSVILLE-OWENSBORO

"High Fidelity Sound Reproduction—Past and Future," by Marvin Camras, Armour Research Foundation and Hobby Talk by W. T. Millis, General Electric Company; April 13, 1955.

FORT WAYNE

"Sampled-Data Control Systems," by Dr. J. M. Salzer, Magnavox Labs.; March 10, 1955.

(Continued on page 64A)


FXR 5 millimeter*
Precision Waveguide Test Equipment
50.00 to 75.00 kilomegacycles
with FULL SPECIFICATIONS
M103X SLOTTED SECTION
Slope and Irregularity—1.02 vswr, max.
Dual Tuned Pickup Probe**
Tapered Slot
Dial Indicator—0.01 mm grauations
Milled Coin Silver Construction
Ball Bearing Carriage
Friction Drive
M151X FLAP ATTENUATOR
Attenuation—0 to 40 db
Approximate calibration
Max. vswr—1.20
M206X TUNABLE DETECTOR MOUNT
IN53 Crystal
Tunable for low vswr
Micrometer Driven Choke Tuning
M312X E/H TUNER
Two Micrometer Driven Choke Plungers
Greater than 1/2 λ travel
Matching—20:1 to less than 1.02:1
M410X FREQUENCY METER
Absolute Accuracy—0.1%
Resonant Dip—approx. 20%
Micrometer Drive
M501X TERMINATION
Max. vswr—1.05
TRANMISSION LINE COMPONENTS
M620X Series Tee
M621X Shunt Tee
M622X E/H Tee
M623X 90 Deg. E-Bend
M624X 90 Deg. H-Bend
M631X Micrometer Sliding Short
M638X Horn
M780X Harmonic Generator
X910X Transmission Line Stand
Z815A Universal Klystron Power Supply
Z762A Klystron Tube Mount
ACCESSORIES:
Will operate tubes (not supplied) thru the frequency range of 18 to 60 Kmc; with harmonic generation up to 90 Kmc; min.
B830A Universal Power Meter
B810A Standing Wave Amplifier
*RG-98/U Waveguide. RG-97/U and RG-99/U waveguide test equipment information also available.
**FXR M206X Tunable Detector Mount not included.
ELECTRONICS & X-RAY DIVISION
FXR
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Write for catalog of complete line of
FXR PRECISION MICROWAVE TEST EQUIPMENT
PRECISION MICROWAVE TEST EQUIPMENT
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Section Meetings

(Continued from page 62A)

HAMILTON

"Glass as an Engineering and Industrial Material," by J. L. Webb, Jr., Corning Glass Works, February 14, 1955.

"Practical Demonstration of Developments in Hi-Fi Sound," by F. H. Slaymaker, Stromberg Carlson Company; March 14, 1955.

"Nuclear Electric Power Stations," by Dr. G. C. Laurence, Atomic Energy of Canada; April 18, 1955.

HOUSTON

"A Proposed Analog to Binary Digital Converter," by C. E. Stevens, Rice Institute Student, and "The Hall Converter," by K. L. Scott, Student, A & M College of Texas; April 19, 1955.

INDIANAPOLIS

"Application of IBM Equipment to Engineering and Statistical Problems," by P. H. Sterbenz, IBM Corp.; February 25, 1955.

ITHACA

"A System for Recording and Reproducing Television Signals by Means of Magnetic Tape," by Dr. H. F. Olson, RCA Labs.; April 7, 1955.

LITTLE ROCK

"The Bell Solar Battery," by H. J. McMains, S. W. Bell Telephone Co.; April 15, 1955.

LONDON

General Business Meeting; March 22, 1955.

"Recent Component Developments in Europe and the North American Continent," by A. Ainlay, Rogers Majestic Electronics Ltd.; April 5, 1955.

LONG ISLAND

"Light Amplifiers," by C. G. Fick, General Electric Company; April 12, 1955.

LOS ANGELES

"Future Problems in Military Communications," by Col. Taylor, "The Role of Telecommunication in Air Transportation," by John Anderson, Aeronautical Radio, Inc., and "Improving Air-Ground Communications," by E. W. Pappenfus, Collins Radio Co.; April 5, 1955.

MILWAUKEE

"Sound Waves and Their Similarity to Microwaves," by F. K. Harvey, Bell Telephone Labs.; April 6, 1955.

PITTSBURGH

"Ferromagnetism," by Dr. L. J. Dijkstra, Westinghouse Electric Corp.; April 11, 1955.

PRINCETON

"Moment Detection," by Prof. J. J. Slade, Jr., Rutgers University; April 14, 1955.

SACRAMENTO

"N M R; What It Is—What It's For," by Dr. Emery Rogers, Varian Associates; April 15, 1955.

SEATTLE

"An Introduction to the IBM 701 Computer," by Randy Porter, Boeing Airplane Co., and "Calibration of a Fuel Gauge on an IBM 701 Computer," by Georges Brigham, Boeing Airplane Co.; January 25, 1955.

"Frequency Measurement Involving Digital Counting Techniques," by Irwin Compton; Hawthorne Electronics, March 3, 1955.

SYRACUSE

"Operations Research and Its Impact on Engineering" by Dr. Andrew Schultz, Cornell University; April 5, 1955

(Continued on page 66A)

Power Meter Model P-2
DC to 11,000 mc

MICRO POWER METER DC to 11,000 mc

measures microwave power with only one probe

Over the entire frequency range DC to 11,000 MC, Polarad's new Micro Power Meter utilizes only **one** power probe, supplied as an integral part of the instrument. This unique power probe will sustain severe overloads without burnout since it does not contain hot wire barreters or other delicate components.

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e. Dynamic test—provision is made for external modulation so that klystron tubes may be dynamically tested with external r-f measuring equipment.

- Special adapter mount for all commercial types of klystrons.
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Section Meetings

(Continued from page 64A)

Toledo

"Television Transmission," by Messrs. Clark and Hornhorst, American Telephone and Telegraph Co.; January 13, 1955.

"Commercial Aircraft Control Radio," by Fred Hjorgsberg, Toledo C.A.A.; February 10, 1955.

"Electronic Counters and General Instrumentation," by Seymour Sterling, S. Sterling Co.; March 10, 1955.

Toronto

"Some Applications of Electronic Techniques to Measurements in Fluid Mechanics Research," by Dr. J. G. Hall, University of Toronto; March 28, 1955.

Tulsa

"Push-Pull Single Ended Audio Amplifier," (by tape) by Arnold Peterson and D. B. Sinclair, General Radio Company, and "The Physics of Music and Hearing," (by tape) by Dr. W. E. Kock, Bell Labs.; March 17, 1955.

Twin Cities

"Analog Between Microwaves and Sound Waves," by F. K. Harvey, Bell Telephone Labs.; April 5, 1955.

Vancouver

"Seismic Prospecting," by W. Dieteker and "Transistors," by Messrs. C. Goodman and A. Csepe, all of U.B.C.; February 21, 1955.

"Some Practical Aspects of Antenna Design," by Prof. Kersey, University of B. C.; March 21, 1955.

"Communications System for the Trans Mountain Oil Pipe Line," by J. S. Gray, Trans Mountain Oil Pipe Line Company; April 4, 1955.

Washington, D. C.

Capacitor Symposium: "Rolled Paper and Film Impregnated Capacitors," by W. A. Allison, Sprague Electric Co.; "Capacitors Employing Metallized Organic Dielectrics," by D. A. McLean, Bell Telephone Labs.; "Aluminum and Tantalum Electrolytic Capacitors," by F. R. Flood, General Electric Co.; and "Titanate Ceramics for Capacitors," by J. G. Breedlove, American Lava Corp.; April 11, 1955.

SUBSECTIONS

Berkshire County

"Guided Missiles with Emphasis on Guidance Problems," By A. W. Robinson, General Electric Company; March 7, 1955.

"Applications of Engineering in Medicine," by J. H. Heller, M.D., N. E. Institute for Medical Research; March 16, 1955.

Buenaventura

"Engineers, Patents and Color Television," by H. R. Lubcke, Consulting Engineer; March 10, 1955.

Centre County

"Printed Circuits," (with Demonstration) by William Leiss, Pennsylvania State University; February 17, 1955.

"The Penn State Computer Laboratory—Past, Present and Future," by Dr. John Warfield, Pennsylvania State University; March 22, 1955.

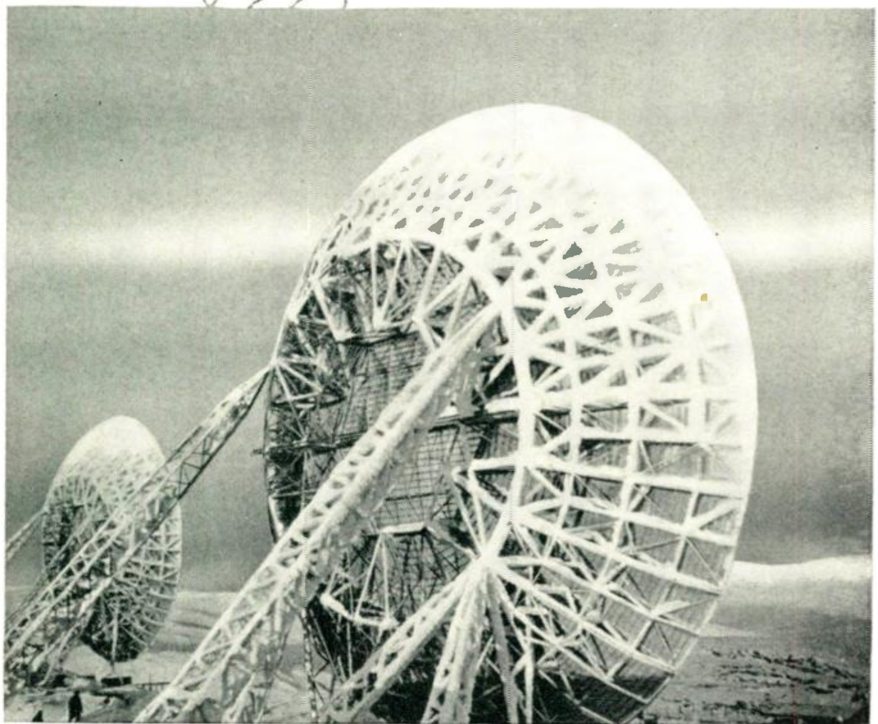
Charleston

"Paper Manufacture," by M. R. Frierson, Public Relations Coordinator; February 8, 1955.

Lancaster

"Electronic X-Ray Screen Intensifications," by Dr. R. M. Morgan and R. E. Sturm, Johns Hopkins Hospital; April 13, 1955.

(Continued on page 68A)



Section Meetings

(Continued from page 66A)

MONMOUTH

"Human Engineering," by Dr. Jesse Orlansky, Dunlap and Associates, Inc.; April 20, 1955.

TUCSON

"Transistor Circuits and Applications," by Dr. Virgil Bottom and Andrew Jacobsen, Motorola Research Labs.; January 28, 1955.

"Methods of Network Synthesis," by Dr. Louis Weinberg, Hughes Aircraft Co.; February 18, 1955.

"Developments in Magnetic Recording Heads," by Dr. Otto Kornei, Clevite Research Lab.; March 11, 1955.

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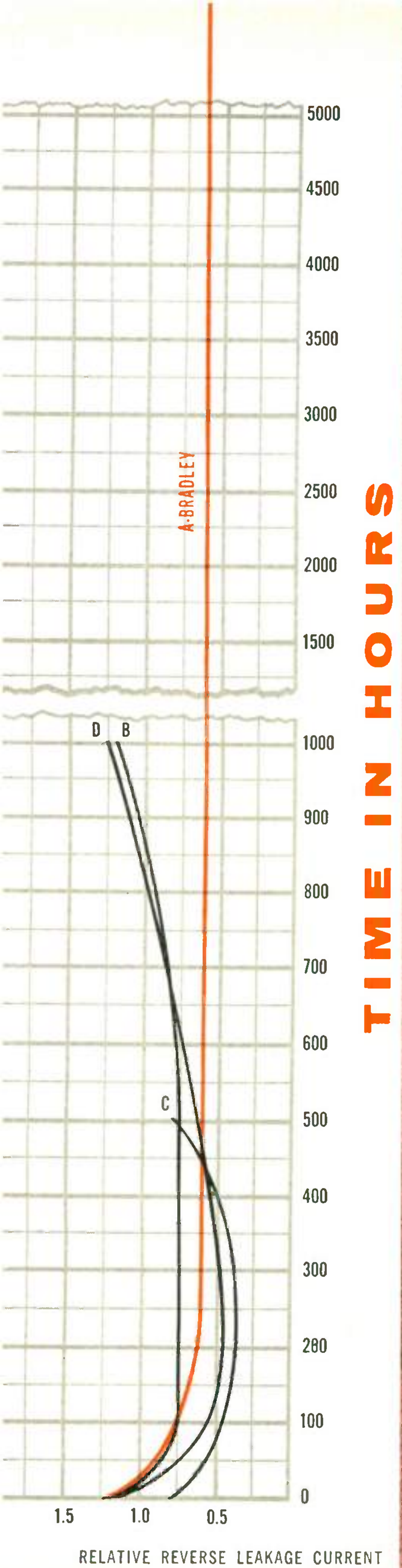
La Rosa, R., 317 Oak St., S. Hempstead, L.I., N. Y.

(Continued on page 70A)



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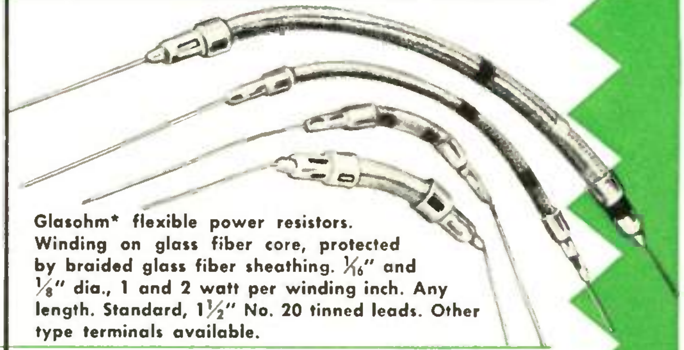

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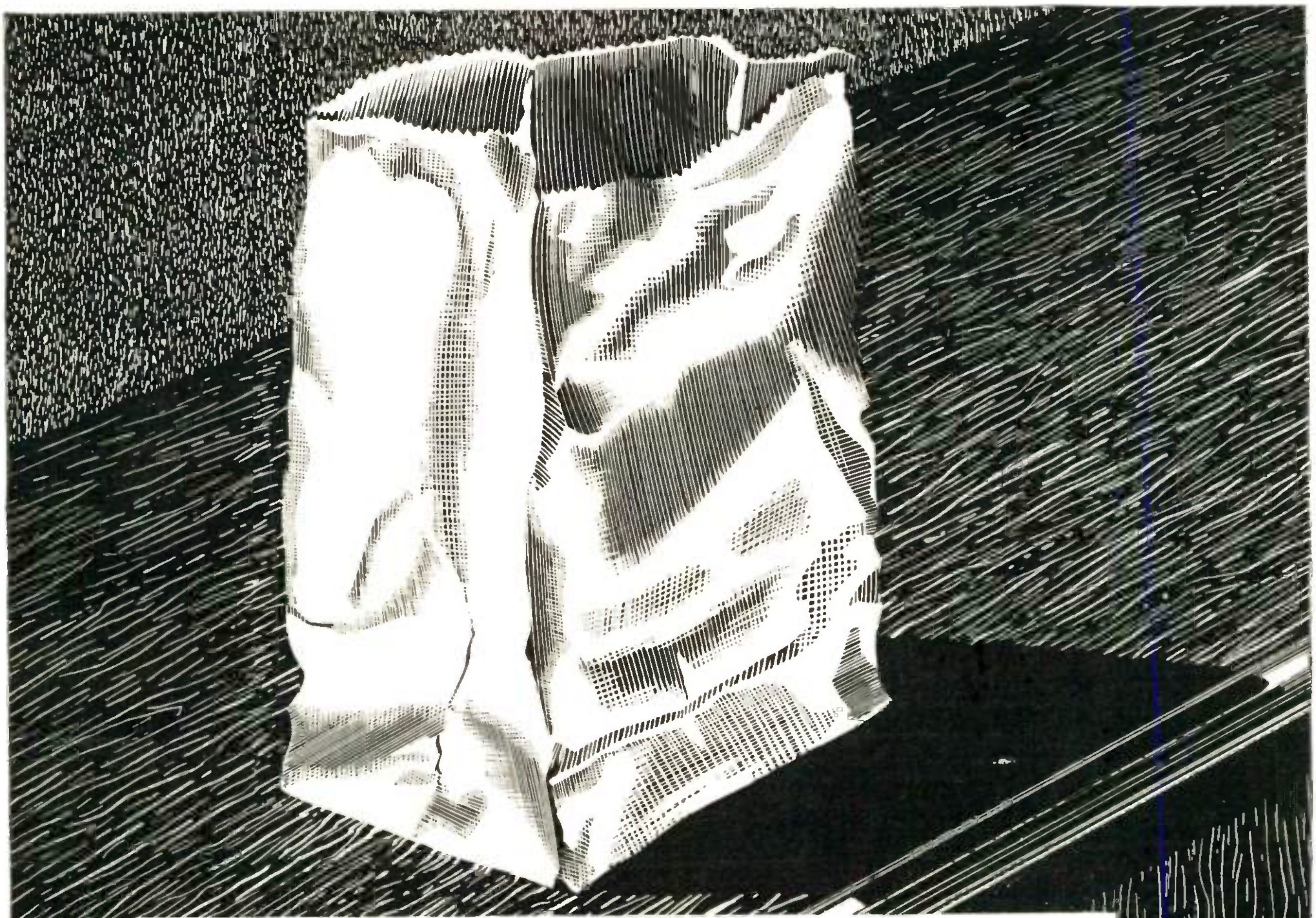
Lieberman, F., Jerrold Electronics Corp., 2222 Chestnut St., Philadelphia 3, Pa.

(Continued on page 76A)



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(Continued on page 78A)

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VARIABLE RESISTORS FOR PRINTED CIRCUITS

Type UPM-45
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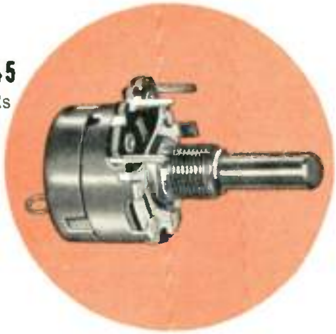
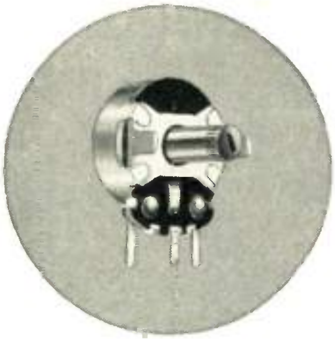


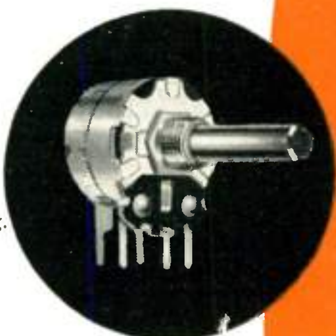
Type GC-U45
Threaded bushing mounting. Terminals extend perpendicularly 7/32″ from control's mounting surface. Available with or without associated switches.

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.

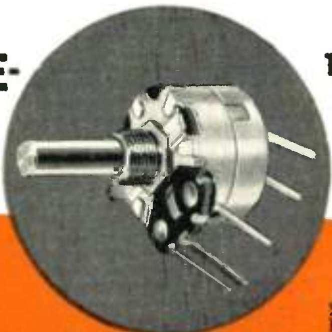


Type XP-45
For TV preset control applications. Control mounts on chassis or supporting bracket by twisting two ears. Available in numerous shaft lengths and types.



Type XGC-45
For applications using a mounting chassis to support printed circuit panel. Threaded bushing mounting.

VARIABLE RESISTORS FOR SOLDERLESS "WIRE-WRAP" CONNECTIONS



Type WGC-45
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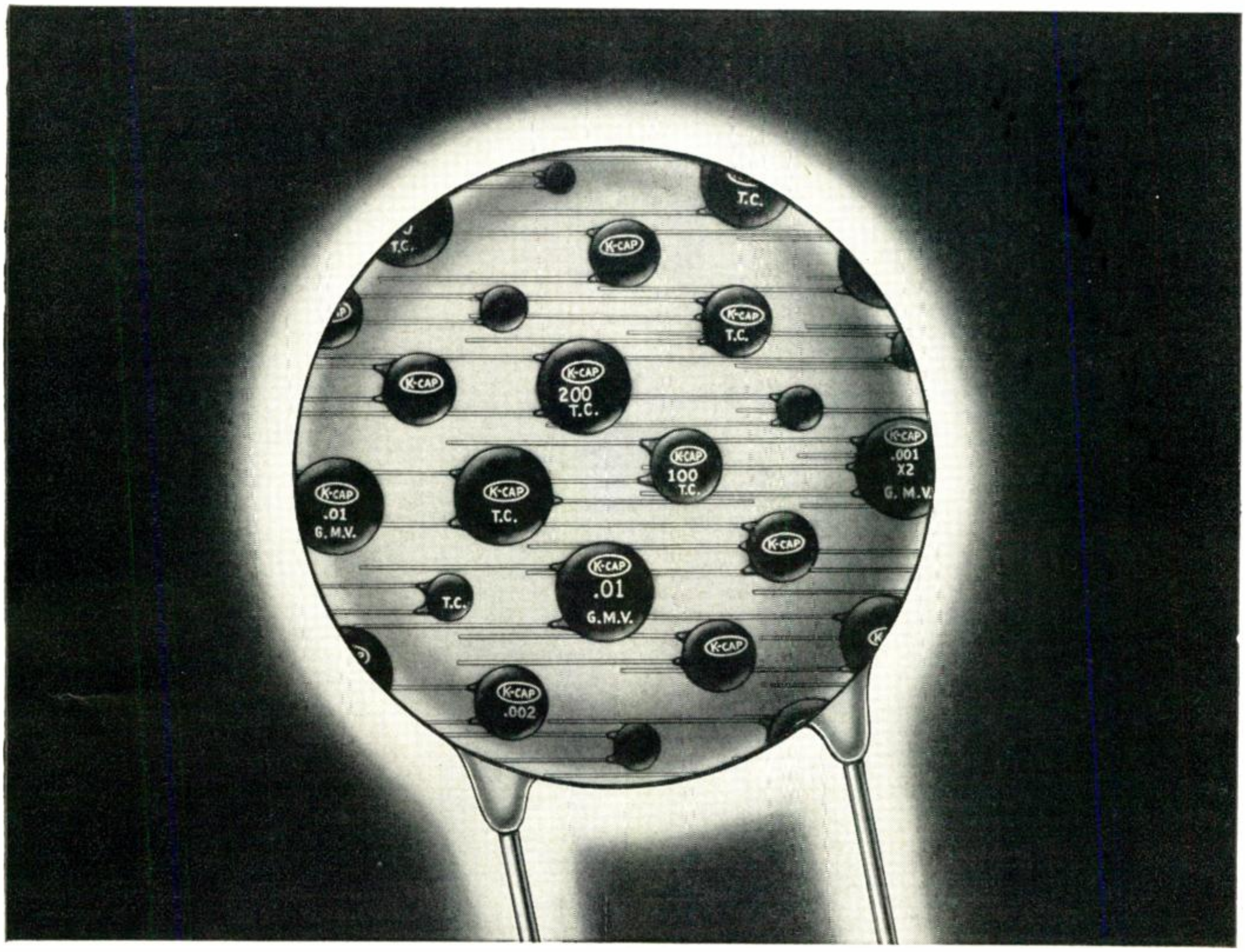

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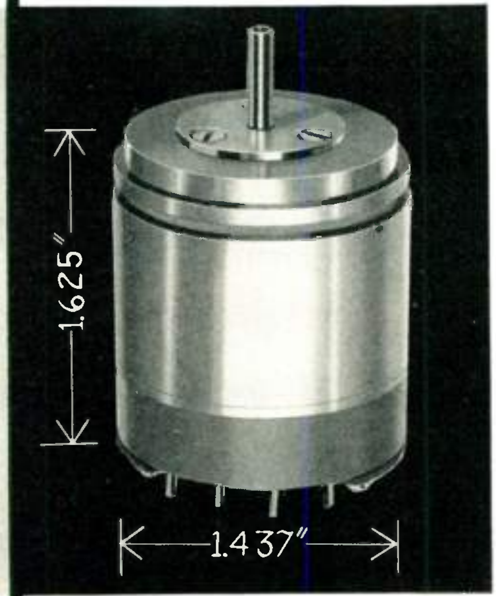

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(Continued on page 88A)

IRE

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(Continued from page 86A)

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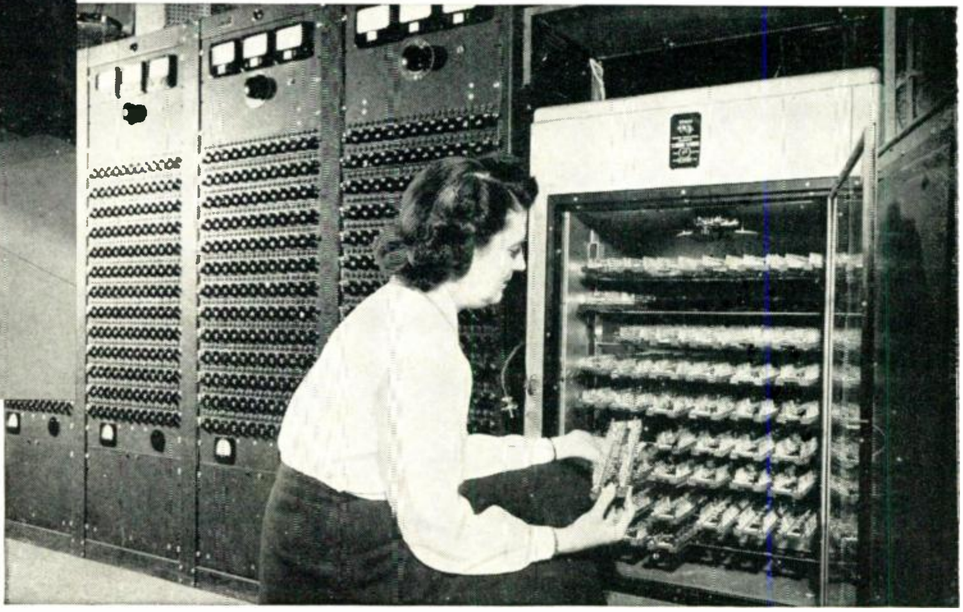

How G.E. assures 85°C operating reliability of the new 1N315 Germanium Rectifier

This array of testing equipment mass-checks individual germanium rectifiers to 10,000-hour standards under *maximum voltage and current* conditions and at *85°C*. Each rectifier is electrically connected to a testing circuit and any unit which falls below guaranteed limits is automatically rejected.

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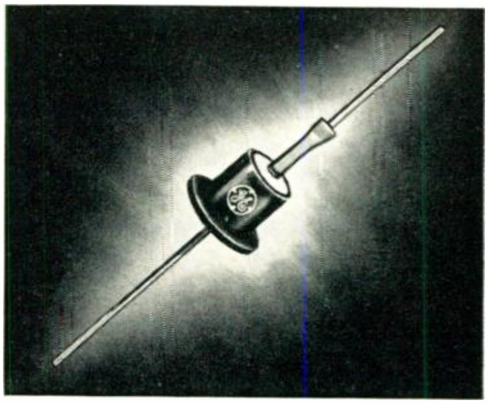
G.E.'s new 1N315 rectifier operates efficiently and with long life at 85°C because of new germanium crystal growing techniques developed by the General Electric Research Laboratory and improved junction forming methods. Controlled characteristics are assured by accurate machines which reproduce assemblies to extremely close tolerances. The high reverse resistance and low forward resistance characteristics make it specially adaptable for use in magnetic amplifiers or other circuits where high temperature and extremely low leakage current are required.

Full details are available on request. Write or wire today to: *General Electric Company, Section X5265, Semiconductor Products, Electronics Park, Syracuse, New York.*



SPECIFICATIONS

1N315 and 1N315A

(Resistive or Inductive Load)

	55°C	71°C	85°C	
Maximum allowable peak inverse voltage	200	200	100	V
Maximum allowable D-C output current	100	100	100	ma
Maximum full load forward voltage drop	.48	.46	.44	V
Continuous reverse working voltage	150	100	50	V
1N315 minimum forward to reverse current ratio (average forward/average reverse at full load)	700	300	200	
1N315A minimum forward to reverse current ratio (average forward/average reverse at full load)	1600	750	400	
Maximum operating frequency (70% rectification efficiency)	50	50	50	KC
Storage temperature	95	95	95	°C

Progress Is Our Most Important Product

GENERAL ELECTRIC

News-New Products

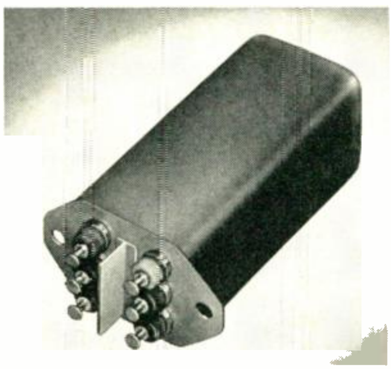
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 20A)

particular installation. The microwave absorbing material can be of any type, including plastic foam, hairflex, or thin flexible material. To prevent disturbance from outside rf sources, the absorber material is backed with a copper screen. An ordinary room can be easily assembled by a few men in one day, with the use of a minimum of standard tools. McMillan also offers a complete design and engineering service for special-type constructions, when needed.

Electronic Chopper

Avion Instrument Corp., 299 State Highway No. 17, Paramus, N. J. has developed a new electronic chopper with no moving parts. The new chopper, identified as Model 307, is able to modulate dc to frequencies up to 400 cps, for a minimum life period of 3,000 hours. Modulation is accomplished by illumination of a photoconductive element in a typical voltage divider.



The new features which have been provided, include: temperature insensitive operation over a range of -50 to $+100°C$; dc to ac conversion ratio over 0.5; noise pick-up of less than 200 μv rms; 115-V, 3-ma, ac excitation.

The Model 307 chopper measures $\frac{7}{8}$ high $\times\frac{7}{8}$ wide $\times 2$ inches long and weighs 1.6 ounces. Further information, including price and delivery, may be secured from the manufacturer.

(Continued on page 92A)

News-New Products

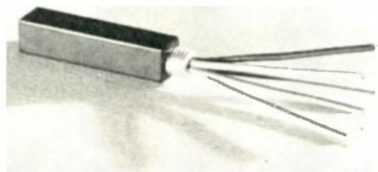
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 90A)

Missile Relay

Smaller than a cigarette, a new subminiature sensitive relay especially designed for guided missile and other applications requiring an extremely small, compact and sensitive relay has been designed by the **Price Electric Corp.**, Frederick, Md.

The new model, the Husky 503, is provided with one set of single pole double throw contacts rated at 0.5 amperes, 120 volts dc non-inductive.



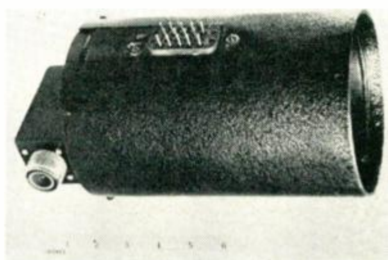
When adjusted to a sensitivity of 0.05 watts, the 503 will resist 30G vibration up to 2,000 cps. It will withstand operating shock of 50G for 0.011 second duration, mechanical shock of 2,000 foot pounds and 1,000G rotary acceleration.

The relay contains a new internal mechanism and is less than 2 inches in length and 13/32 of an inch square.

Full data is available on request.

Daylight Viewing CRT

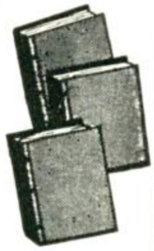
National Union Electric Corp., 350 Scotland Rd., Orange, N. J., announces the type NU-DVI-3, a high brightness, daylight viewing, five-inch oscillograph presentation unit that uses both electrostatic focusing and magnetic deflection. It presents a small brilliant spot and is capable of providing a sharply focused television raster at high brightness levels.

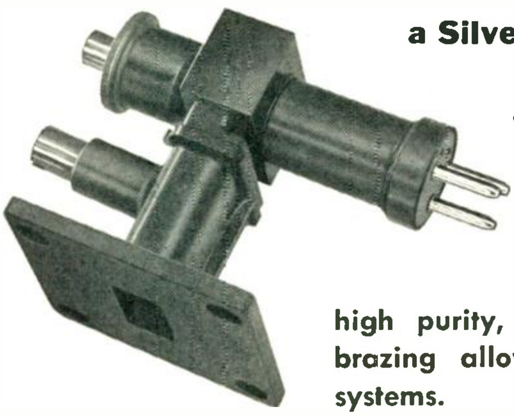


The unit is designed to withstand MIL shock and vibration tests, and to facilitate mounting on aircraft instrument panels and for similar applications. The cathode-ray tube and other components within the aluminum cylindrical container are treated to prevent corona at high altitudes.

The NU-DVI-3 weighs 4 pounds 8 ounces, has a diameter of 5¼ inches and an over-all length of 10 inches.

(Continued on page 94A)





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 92A)

AM Signal Generator

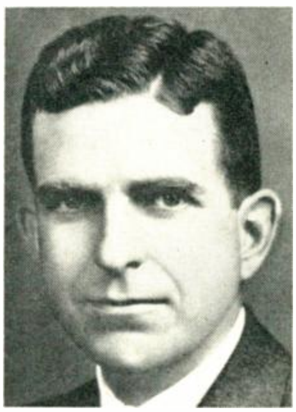
New London Instrument Company, Inc., 82 Union St., New London, Conn., announces the introduction of a new AM Signal Generator to its line of electronics precision measuring instruments.

All components, with the exception of the power supply, are in the RF cavity. This minimizes leakage and decreases the number of filters needed.

The Model 162 features negligible FM, and low drift with temperature and power line changes. Bands are changed by a rotating turret arrangement. Internal modulation is 400 and 1,000 cps from 0 to 50 per cent. External modulation is 0 to 80 per cent with 100–1,000 cps and carrier frequency up to 300 kc and 100–15,000 cps with carrier frequency of 300 kc to 50 mc. The complete unit weighs 35 lbs.

Lowell Appointed by Sorensen

Henry T. Lowell, Jr., has been appointed sales manager of **Sorensen & Co.**, Stamford, Connecticut, manufacturers of electronic power regulating equipment.



Mr. Lowell comes to Sorensen after serving with the Westinghouse Electric Corp. for 17 years. He is a graduate of the University of Maine with a B.S. degree in electrical engineering.

(Continued on page 98A)

Professional Group on Communications Systems

The field of communications systems is as old as radio itself. For three decades after Marconi first spanned the Atlantic by radio in 1901 radio researchers devoted practically all their efforts to developing and perfecting new systems for utilizing electromagnetic waves for communication purposes.

Gradually, however, this newfound knowledge opened the doors to other applications in industrial, military, scientific, and medical fields. Under the stimulus of wartime research these newer applications multiplied and expanded at a tremendous rate so that at the end of World War II the complexion of the radio engineering field had altered drastically.

What had once been a fairly small, homogeneous group with common technical interests had now become a large body of engineers and scientists working in at least a score of separate and distinct specialized fields. Communications engineers found it increasingly difficult to find the specialized technical information they urgently needed since the sources for such information now had to serve many other fields as well.

On April 8, 1952, the IRE formed the Professional Group on Communications Systems. The response was almost instantaneous as communications engineers all over the U.S., and in other countries as well, realized that for the first time an organization had been formed just to serve their particular technical needs.

As membership increased, the Group began publishing its own TRANSACTIONS, which is now issued free to all members three times a year. The Group also has become very active in sponsoring meetings in various parts of the country throughout the year. The value of these activities is attested to by the fact that some 1100 engineers have joined the Group and have found this affiliation indispensable to their professional well being.

W. R. G. Baker

Chairman, Professional Groups Committee

At least one of your interests is now served by one of IRE's 23 Professional Groups

Each group publishes its own specialized papers in its *Transactions*, some annually, and some bi-monthly. The larger groups have organized local Chapters, and they also sponsor technical sessions at IRE Conventions.

Aeronautical and Navigational Electronics (G 11)	Fee $2
Antennas and Propagation (G 3)	Fee $4
Audio (G 1)	Fee $2
Automatic Control (G 23)	Fee $2
Broadcast & Television Receivers (G 8)	Fee $2
Broadcast Transmission Systems (G 2)	Fee $2
Circuit Theory (G 4)	Fee $2
Communication Systems (G 19)	Fee $2
Component Parts (G 21)	Fee $2
Electron Devices (G 15)	Fee $2
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Instrumentation (G 9)	Fee $1
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Reliability and Quality Control (G 7)	Fee $2
Ultrasonics Engineering (G 20)	Fee $2
Vehicular Communications (G 6)	Fee $2

IRE Professional Groups are only open to those who are already members of the IRE. Copies of Professional Group Transactions are available to non-members at three times the cost-price to group members.

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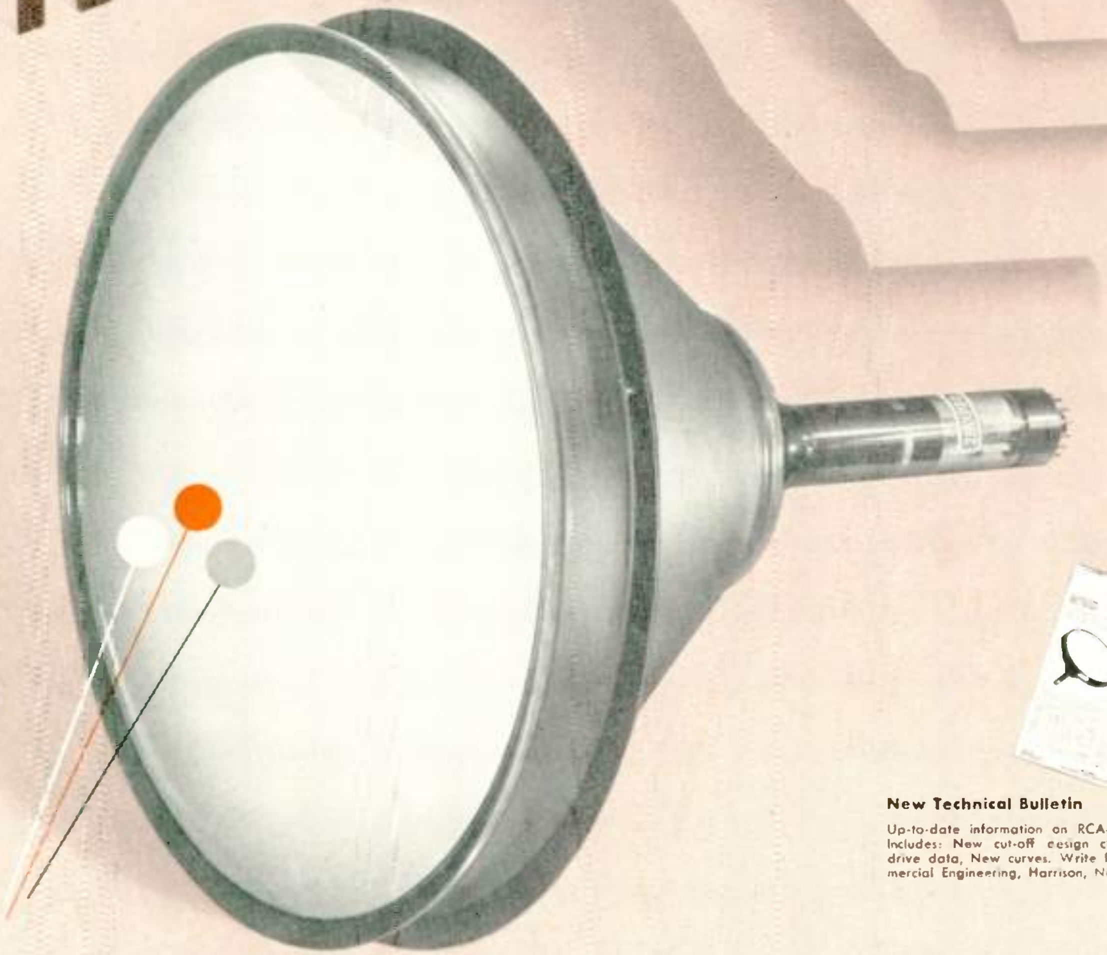

New Technical Bulletin

Up-to-date information on RCA-21AXP22. Includes: New cut-off design chart, New drive data, New curves. Write RCA, Commercial Engineering, Harrison, New Jersey.

RCA 21" COLOR PICTURE TUBES

To meet your production-line requirements on color-TV sets, RCA's extensive manufacturing facilities are now turning out 21-inch color picture tubes—*in quantity*. And, these facilities are being expanded to keep pace with your needs. Now you can plan "21-inch color" with confidence! Now, you can capitalize on the advantages of the RCA-21AXP22 in your color TV line. Consider just a few features of this remarkable tube:

- *Brilliant, faithful colors*—Through unique shadow mask.
- *Big, life-size pictures*—255 sq. in. area.
- *Bright, sharp pictures in color or black-and-white*—Precise registration—uniform color field.
- *Compact overall size for attractive cabinet styling*—70° deflection, plus short electron gun.
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- *Light, strong metal-bulb construction*—Weight 28 lbs.—no external magnetic shielding required.
- *Economical*—All these features—and more—at moderate cost.

For complete technical data on the continuously produced RCA-21AXP22 color tube—*and the other RCA tubes needed to round out your design and production program*—call your RCA Field Representative.

RADIO CORPORATION of AMERICA

ELECTRON TUBES **HARRISON, N.J.**

PROCEEDINGS OF THE IRE

Published Monthly by

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CONTENTS

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Haraden Pratt

SECRETARY, 1955

Haraden Pratt, Secretary of the IRE, was born in San Francisco, California on July 18, 1891. He commenced his career in radio as an amateur in 1906, and became a wireless telegraph operator and installer of equipment for the United Wireless Telegraph Company and Marconi Wireless Telegraph Company of America during the years 1910–1914.

Mr. Pratt, who received the B.S. degree in electrical engineering from the University of California in 1914, became a construction and operating engineer for the Marconi Company's 300-kilowatt spark-type Trans-Pacific radio stations in California.

From 1915 to 1920 he was an Expert Radio Aide for the Navy Department and was primarily occupied with the construction and maintenance of its high-powered radio stations. Commencing in 1920, he established the public service radiotelegraph system of the Federal Telegraph Company on the Pacific Coast. In 1925 he constructed and operated a radiotelegraph system between Salt Lake City and Los Angeles for the Western Air Express; this was followed in 1927 by development work on radio aids for air navigation of which he was in charge at the Bureau of Standards in Washington, D. C. In 1928 he became Chief Engineer, and later Vice-President, of Mackay Radio and Telegraph Company. He constructed its world-wide communication plant.

In 1945 he became Vice-President and Chief Engineer of the Commercial Cable Company, All America Cables and Radio, Inc., and the American Cable and Radio Corporation. For many years he held offices in other companies of the International Telephone and Telegraph Corporation. He retired from these activities in 1951.

Mr. Pratt has attended most international radio and telegraph conferences since 1926 as either a technical or industry adviser. He was a director of the American Standards Association from 1939 to 1942, chairman of the Radio Technical Planning Board from 1945 to 1948, and has been a member of the Joint Technical Advisory Committee.

During World War II he became chief of the National Defense Research Committee's Division 13 on Communications, and in 1948 was awarded a Presidential Certificate of Merit. In October, 1951 Mr. Pratt was appointed by President Truman to the newly-created post of Telecommunications Advisor to the President. He has since retired from government service.

Mr. Pratt is a member of Sigma Xi, Fellow of the American Institute of Electrical Engineers, Fellow of the Radio Club of America, Associate Fellow of the Institute of the Aeronautical Sciences, and a member of the Veteran Wireless Operators Association.

He joined the IRE as an Associate in 1914, became a Member in 1917, and a Fellow in 1929. He has been a Director since 1935, Treasurer in 1941–1942, Secretary since 1943, and President in 1938. Mr. Pratt received the Institute's Medal of Honor in 1944.

Keeping IRE Section Historical Records

MERRILL S. SMITH
FORMER CHAIRMAN OF THE CEDAR RAPIDS SECTION

Records are history. This is true whether you are trying to keep account of your wife's shoe budget or IRE Section records. Why keep historical records anyway? In the first place, you figure it is necessary to prove to the little woman just how much she really spends on shoes each year. The reason for this is obvious. Why keep IRE Section Records? That is the subject of this discussion.

Seriously now, how many IRE Sections look ahead and actually consider the need for historical records at some time in the future?

The Cedar Rapids Section has just completed its tenth year of existence. We are still in our youth. Yet our section has lost four of the last ten chairmen. Soon we will not be able to sit around a table, depending upon our memories, to bring up the past. It seems necessary to provide assurance that those records we now have will be preserved and that proper records will be collected each year in the future. How might we best accomplish this?

The Cedar Rapids Section created the office of Section Historian in 1953. It became obvious to the historian, from a review of the accumulation of records, that a uniform system of preparation, collecting and preserving records was necessary.

Records should be kept permanently only if they will be used for some purpose at a later date. The question is, when and to whom would such section records prove valuable?

The section nominating committee would do well to review the past participation of section members in section activities. What could be more helpful than a complete list of officers, committee chairmen and committee members for three or four years back. If the nominating committee were considering a particular individual, would it not be helpful if it could refer to a permanent record card containing data on professional background and experience as well as full details of participation in local section activities? We can conceive of a "traveling" membership record card which follows a member from section to section as he changes his residence from one area to another. The newly elected section chairman and his executive committee could make good use of such records in selecting committee chairmen.

The incoming program chairman would find a chronological record of past programs very interesting in avoiding duplication of program material, selecting popular subjects and outstanding speakers. If appropriate comments were added, other sections would find this record most useful.

It would seem essential to preserve a financial history of the section. This might consist of the annual financial report as a minimum and, of course, would include the financial results of any special activities such as technical conferences. There is no need to preserve bank statements, cancelled checks, bills, etc. indefinitely.

The Cedar Rapids Section has found Conference Reports very valuable. These reports consist of recommendations of all Technical Conference Committee chairmen and may be used in planning conferences for the future. In effect, a collection of these reports forms a "How to Run a Conference" instruction book.

The minutes of section and executive committee meetings are a must for the permanent record. These minutes contain reference to all business transactions of the section. It is important, however, to verify the quality of the minutes prepared by the secretary. Normally this is covered by the chairman, as the minutes are read during the regular business meeting. A standard form of minutes can be established to assure complete coverage. All minutes should be typed on a good grade, rag content paper for permanence.

It is customary for retiring section chairmen to prepare a final report to the section for delivery at the annual meeting. With careful planning, this report can be made into a condensed historical review of the accomplishments of the section during this chairman's term of office. A file of similar reports would present a continuous story on section history. Using such material, the Section Historian can readily complete his work with a minimum searching for information.

Only a few examples of the use of permanent records can be given here. The purpose of this editorial is to start a discussion on the need for planning in the preservation of vital IRE records. Standardization of the form and content of records is needed to assure retention of proper information. Possibly a committee should be established on a regional or national level to study this problem and make recommendations for inclusion in the Sections Manual. Such a committee might answer the questions of whether a Section Historian is necessary, and whether the office of Section Historian should be established officially with firm responsibilities. What do you think on this subject?

A Developmental Pocket-Size Broadcast Receiver Employing Transistors*

D. D. HOLMES†, ASSOCIATE, IRE, T. O. STANLEY†, AND L. A. FREEDMAN†, ASSOCIATE, IRE

Summary—**This paper describes a pocket-size developmental AM broadcast receiver which utilizes eight junction transistors. Its performance is comparable to that of conventional personal receivers. Emphasis has been given to developments which contribute to stability with respect to temperature, battery voltage, and variations among transistors. The superheterodyne circuit employed uses a single-transistor frequency converter to perform the functions of both mixer and oscillator. Refined detector and automatic-gain-control circuits and an audio amplifier embodying further development of the principle of complementary symmetry are incorporated. Reduction in physical size and battery requirement, as compared to conventional receivers, is substantial.**

The circuits are described in detail and certain aspects of components and of physical arrangement, which contribute to the small size, are discussed. Detailed performance data are also included.

General Description

The receiver is shown in Fig. 1. Operation of tuning and volume-on-off controls is accomplished by means of rim-drive wheels which protrude through slots at the end of the receiver case. The tuning indication is marked on the tuning wheel and is viewed through the window at the lower right. A 2×3 inch speaker is located behind the lower portion of the decorative grille, and the back of the case is vented to improve acoustical performance. The 4-cell battery is contained in a compartment which is located at the rounded end of the receiver and provided with a snap-on cover.

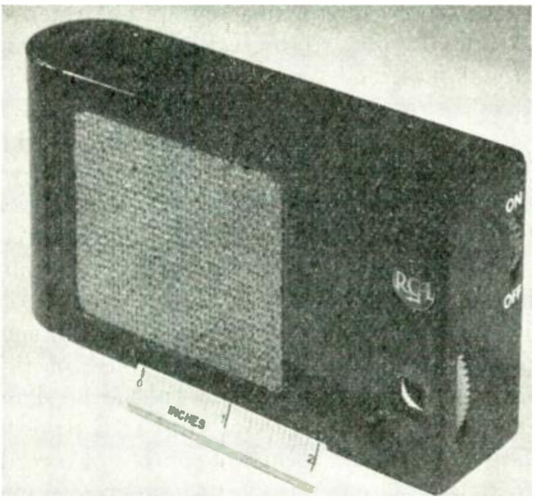


Fig. 1—Pocket-size broadcast receiver.

The over-all dimensions of the receiver are: height $2\frac{3}{4}$ inches, length $5\frac{1}{8}$ inches, thickness $1\frac{1}{4}$ inches. The total weight of this unit is 17 ounces. The dimensions of the receiver are determined principally by the speaker, tuning condenser, antenna core and battery, i.e., the transistors and small components occupy only a fraction of the total volume of the receiver. For pocket use, the thickness is an important dimension. In this receiver the desired minimum thickness was achieved by employing shallow (approximately 1 inch) versions of a conventional speaker and a conventional tuning condenser. The magnet structure of the speaker was rearranged to minimize the over-all depth. Approximately the same volumes of magnet and iron are employed in both the original and modified units; no degradation in performance is introduced by the modification. The tuning condenser was modified by removing some of the plates and shortening the shaft and frame.

The salient performance characteristics of the receiver are as follows:[1]

Sensitivity	100 μv/m
Noise performance	
ENSI	15 μv/m
Input for 20 db S/N	1,300 μv/m
Power output	125 mw
Selectivity (ACA)	28 db
AGC (figure of merit)	37 db

The battery life of the receiver shown in Fig. 1, employing RM-1 cells, is approximately 50 hours. Alternative battery cases housing battery complements representing different design compromises among battery life, size, weight, cost, etc., may readily be substituted. For example, the battery case shown in Fig. 2 (above)

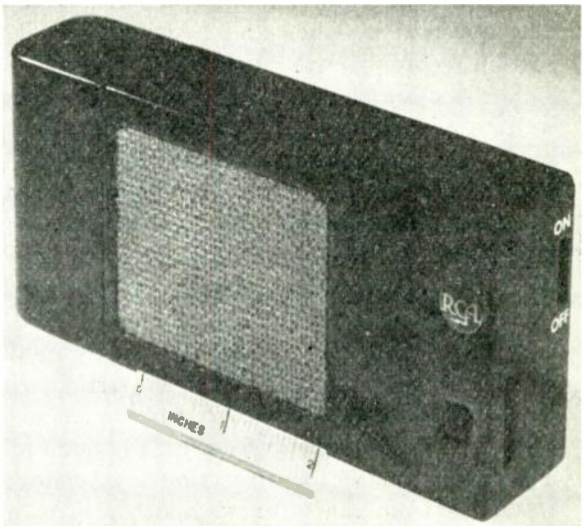


Fig. 2—Receiver case with alternative battery case attached.

* Original manuscript received by the IRE, January 26, 1955.
† RCA Laboratories Div., Princeton, N. J.

[1] These data reflect the characteristics of the experimental transistors used; units of exceptionally high and exceptionally low performance were avoided.

accommodates either RM-502 cells, providing a battery life of 120 hours, or conventional "pen-lite" cells, providing a 36-hour life. This battery case increases the receiver length by ½ inch.

Circuit Description

A schematic diagram of the complete receiver is shown in Fig. 3 on the following page.

Transistors V1 through V4 are experimental radio-frequency units,[2] and serve as converter, first and second IF and second detector stages respectively. The intermediate frequency is 455 kc. Transistors V5 through V8 are the audio-amplifier complement. Transistors V5 and V6 are identified by the type numbers 2N34 and 2N35 respectively. Transistors V7 and V8 are experimental high-current *p-n-p* and *n-p-n* junction units, respectively, connected as a class B complementary symmetry output pair working directly into voice coil of speaker.

The 5-volt battery consists of four RM-1 Mallory Mercury cells. The positive side of the battery is set at +1 volt with respect to ground by the bleeder combination of R7 and R8. The battery is centertapped to provide the symmetrical supply required by output stage.

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

Antenna

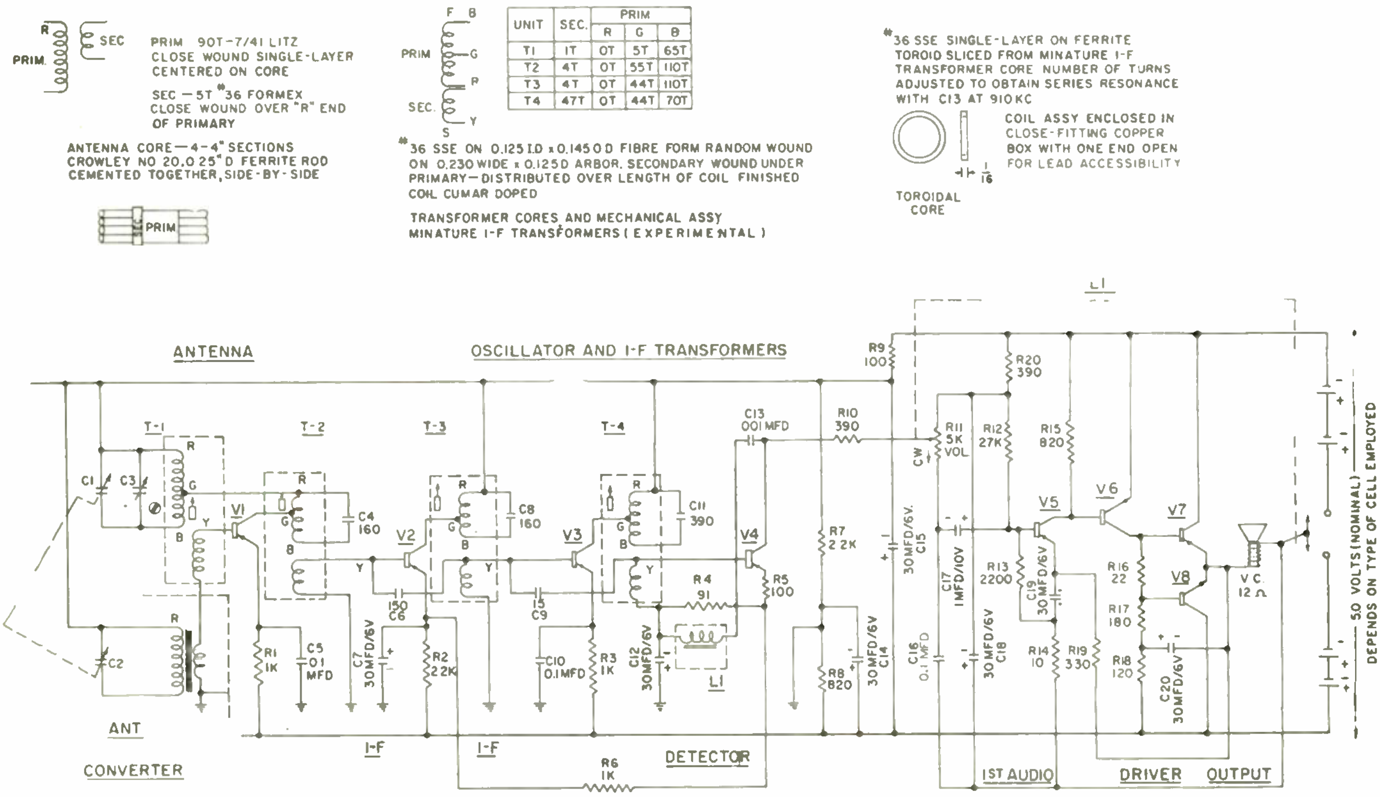
The receiver antenna consists of a ferrite-cored loop. The core is made up of four 4-inch sections of one-quarter inch diameter ferrite rod placed side-by-side to form a flat structure. This arrangement provides a large antenna volume in a shape which is compatible with the remainder of the receiver. As large a volume as was considered practicable was utilized since the power extracted from a given radiated field by such an antenna is proportional to the volume of the ferrite core.

The antenna loop and secondary winding serve as an impedance transformer; the turns ratio is designed to match the high tuned impedance of the antenna to the low input impedance of the converter at approximately midband. This entails no appreciable sacrifice in performance at the extremes of the band.

A grounded copper shield is located beneath the antenna between the antenna and the remainder of the receiver to reduce feedback of IF and its harmonics from the detector circuit to the antenna.

Converter

Oscillator and mixer functions are performed by V1, the converter transistor. The oscillator transformer, T1, provides tickler feedback from collector to base. The IF



ALL CAPACITANCE VALUES MMFD UNLESS OTHERWISE SPECIFIED
ALL RESISTORS ⅓ WATT—10% TOLERANCE

TRANSISTORS
V1–V4 (p-n-p) EXPERIMENTAL R-F ALLOY JUNCTION TRIODE
V5-(p-n-p) — TYPE 2N34
V6-(n-p-n) — TYPE 2N35
V7-(p-n-p), V8 (n-p-n) EXPERIMENTAL HIGH CURRENT ALLOY JUNCTION TRIODE

Fig. 3—Schematic of pocket-size broadcast receiver.

take-off transformer, T2, is located in the collector circuit in series with the oscillator transformer primary. Signal is applied to the converter base via the antenna secondary winding in series with the oscillator feedback winding. The input circuit is returned to chassis ground; the collector voltage is -4 volts. The converter is constant-emitter-current biased at approximately 1 ma by means of the 1K emitter resistor, R1, which returns to the $+1$ volt bus. The 1-volt bias supply is several times the emitter-to-base operating potential of the transistor. Thus, variations in emitter-to-base operating potential among transistors, or with temperature for a given transistor, have a negligible effect upon emitter current. The bleeder which fixes ground potential is relatively stiff (2 ma) so that base current variation with temperature and among transistors does not shift the operating point. Constant-emitter-current bias provides stability of operating point over a wide temperature range and affords transistor interchangeability.

The converter emitter is bypassed to ground by C5. Bypassing to ground in this manner (rather than to the $+1$-volt bus) provides effective rf isolation of the input circuit from the positive side of the battery.

The turns ratios of the oscillator transformer have been determined experimentally to provide high operating Q and near-optimum oscillator injection over the band. A representative value of optimum injection is 0.1 volt rms at the converter base. The oscillator frequency is substantially unaffected by a 50 per cent reduction in battery voltage or a variation in ambient temperature from 0 degrees C. to 50 degrees C.

The input impedance of the converter varies approximately 2 to 1 over the broadcast band. An average value is 300 ohms. The converter output impedance at IF is approximately twice that of the same transistor as an IF amplifier, or about 40,000 ohms. A typical value for conversion gain is 22 db.

IF Amplifier

The three IF transformers, T2, T3 and T4 comprise three single-tuned circuits which serve as interstage coupling networks, having essentially unity coupling between primary and secondary windings. These transformers embody an experimental arrangement and are ferrite cored. They are approximately $\frac{3}{8}$ inch in diameter by $\frac{1}{2}$ inch high. For any given operating Q, minimum insertion loss is incurred by choosing the transformer turns ratios so that the reflected load impedance is equal to the transistor driving impedance. In this receiver, each transformer is designed for an operating Q of 35; the resulting insertion loss is 2.5 db per transformer. The different operating conditions of the various stages give rise to differences in input and output impedances among stages; the various transformer turns ratios are chosen accordingly.

The first IF transformer, T2, feeds the base of the first IF transistor, V2, the input impedance of which is 150 ohms. A neutralized common-emitter connection is employed. The operation of the neutralizing circuit may be visualized by referring to the single-generator common-emitter π-equivalent circuit of the transistor, shown in Fig. 4. Feedback from the output to point b'

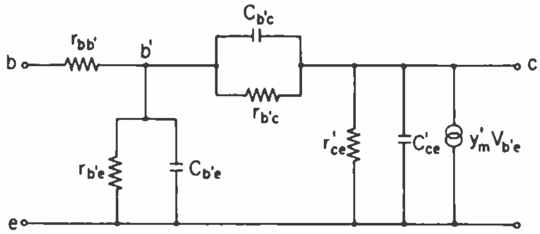


Fig. 4—Single-generator common-emitter π-equivalent circuit of experimental radio-frequency alloy junction transistor.

occurs via the parallel combination of $C_{b'c}$ and $r_{b'c}$. The transistor feedback capacitance and resistance are in the order of 15 $\mu\mu$f and 0.1 megohms respectively. This feedback may be neutralized by deriving a reverse-phase voltage from the output which is then fed back through a suitable parallel rc circuit to the base connection, point b. The presence of $r_{bb'}$ renders the neutralization slightly dependent upon driving impedance and frequency, and influences the requisite parameters in the neutralizing circuit. This effect is second order at 455 kc, however, and can be neglected.

The first IF transistor is provided with an initial constant-emitter-current bias of 0.5 ma; choice of bias for this stage is dictated in part by AGC considerations. The operation of the AGC circuit is described in conjunction with the second detector.

Neutralization is accomplished by utilizing feedback from the secondary of the second IF transformer, T3, through the 150-$\mu\mu$f neutralizing capacitor, C6, to the base. The proper primary-to-secondary polarity is as indicated on the schematic. Because the resistive component of transistor feedback is relatively small, resistor neutralization need not be employed. Since the primary-to-secondary turns ratio of T3 is approximately 10 to 1, a neutralizing capacitor which is ten times the effective transistor feedback capacitance is used. T3 provides interstage coupling between the 20,000-ohm output impedance of the first IF stage and the 150-ohm input impedance of the second IF stage. An IF stage gain of approximately 28 db is realized.[3]

The second IF transistor, V3, is constant-emitter-current biased at 1.0 ma. The third IF transformer, T4, operates between the 20,000-ohm output impedance of the second IF stage and the 20,000-ohm input impedance of the second detector. This value of detector input impedance applies at receiver-sensitivity level; the detector input impedance decreases with increasing signal level. The primary-to-secondary turns ratio of this transformer is approximately 1 to 1 so that a 15-$\mu\mu$f neutralizing capacitor, C9, is required.

[3] By way of example, IF transistors having $r_{bb'}=75$ ohms, $\alpha_{cb}=20$, $C_{b'c}=11$ $\mu\mu$f, and $C_{b'e}=0.001$ μf ("alpha cutoff" of 6 mc) would give typical performance.

Automatic gain control of the first IF stage is accomplished by varying its dc emitter current as a function of the signal level at the detector. The manner in which 455-kc gain, input impedance, and output impedance vary with emitter current is shown in Fig. 5. The transistor gain decreases rapidly as the dc emitter current is reduced below 0.5 ma. The input and output impedances increase with decreasing emitter current so that the stage gain is further decreased by input and output circuit mismatches. A single-stage control range of approximately 45 db is obtained. It is evident that automatic gain control will result in a change in operating Q; the effect of this change on over-all bandwidth is discussed below.

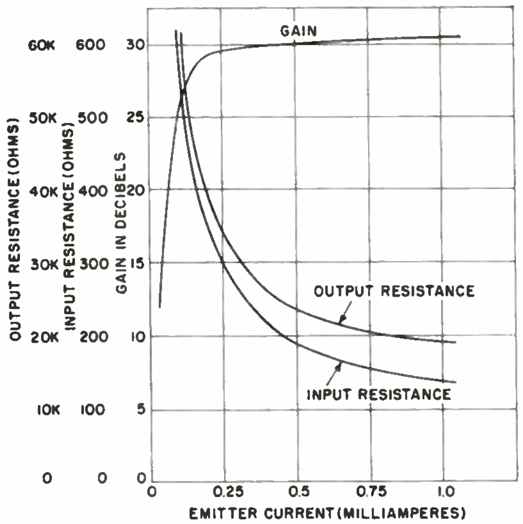


Fig. 5—Typical variation of rf transistor characteristics with variation of emitter current.

Second Detector and AGC

The second detector transistor, V4, is driven by the third IF transformer secondary. This stage operates with the base and emitter at the same dc potential for the zero-signal condition. As signal is applied to the detector, substantially three components of collector current are developed; these components increase with signal. Included are a high-frequency component (IF and its harmonics), an audio-frequency component, and a dc component. The return path for the high-frequency component is from the collector through the 910-kc series resonant circuit formed by C13 and L1, and thence through the series combination of R4 and R5 to the emitter. The 910-kc series-resonant circuit confines the IF second harmonic current to one return path: reduction of IF second harmonic feedback to the antenna is simplified since the inductive field of the resulting single loop may then be readily shielded from the antenna. Resistor R10 blocks high-frequency components of collector current from the volume control and associated wiring. Audio-frequency components of collector current return via C12 through the series combination of R4 and R5 to the emitter. The degeneration provided by R4 and R5 serves to reduce detector distortion.

The detector audio output is developed across the upper portion of the 5K volume control, R11. The collector is returned to the volume control slider to provide a detector-circuit output impedance which is substantially independent of volume control setting; the collector impedance is high compared to the volume control resistance. The desirability of a constant audio driving impedance for the particular audio amplifier utilized is discussed later.

Curves of detector distortion and output voltage vs detector input are shown in Fig. 6. Distortion at approximately receiver-sensitivity level is 11 per cent at 80 per cent modulation, and falls to less than 2 per cent at a signal level corresponding to the knee of the AGC characteristic.

The dc component of detector emitter current is employed for AGC purposes. The dc emitter current return path is from the positive side of the battery through the resistor, R2, in the emitter circuit of the first IF stage, thence through R6 and R5 to the detector emitter. Negative feedback for dc in the detector circuit is provided by R5, introducing an effective delay in the AGC action of the detector. The 30-μf emitter bypassing capacitor of the first IF stage, C7, in combination with R6, serves to filter out audio-frequency components of detector emitter current.

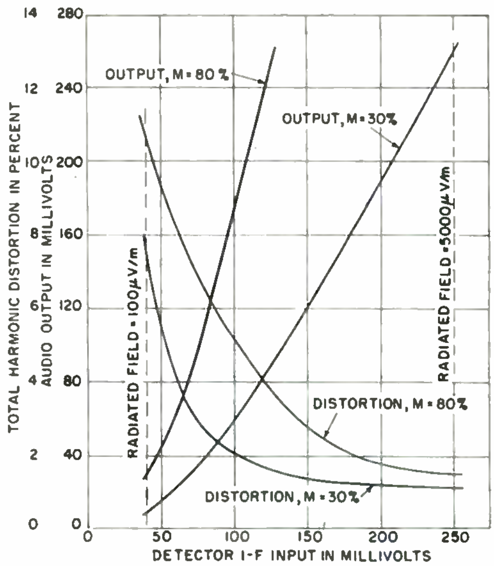


Fig. 6—Detector characteristics.

The operation of the AGC circuit is as follows. For the zero-signal condition, the first IF stage is constant-emitter-current biased at 0.5 ma, and substantially no dc emitter current flows in the detector circuit. As signal is increased, detector dc emitter current is developed, and flows through the resistor R2, in the emitter circuit of the first IF stage. Since the bias arrangement of the first IF stage holds the current in R2 essentially constant, the detector may be considered to "rob" the IF stage of emitter current, i.e., the dc emitter current shifts from the IF stage to the detector. As the detector emitter current approaches 0.5 ma, the IF stage emitter current (and stage gain) approaches zero. Thus, the detector dc emitter current at the flat portion of the AGC characteristic is equal to the first IF stage dc emitter current at zero signal. The value of 0.5 ma is chosen as being high enough to insure operation of the first IF stage near maximum gain for zero-signal condition and low enough to prevent detector overload at maximum volume control setting under strong signal conditions.

The detector circuit input impedance varies with signal level, being about 20,000 ohms at receiver sensitivity level, and dropping to about 5,000 ohms at the knee of the AGC characteristic. This produces a lowering of the operating Q of the last IF transformer with increasing signal level. At the same time, however, AGC action increases the input and output impedances of the first IF stage; the operating Q's of the first and second IF transformers are increased so that a negligible change in over-all bandwidth occurs.

The first four stages of the receiver are decoupled from the battery by R9 and C15 which prevent audio-frequency components developed across the battery impedance from affecting the oscillator and AGC circuits. The +1-volt bus is bypassed to ground for audio and radio frequencies by C14.

Audio

The detector is followed by a three-stage audio amplifier employing cascade complementary symmetry[4] in the first two stages, which operate class A, and push-pull complementary symmetry in the output stage, which operates class B. The voice coil is driven directly from the emitters of the output stage. Signal feedback from the voice coil to the first audio transistor reduces distortion, relaxes the degree of match required in the output transistors, and enhances interchangeability of transistors. Stabilization of operating biases over a wide temperature range is achieved by over-all dc feedback.

The signal feedback from the voice coil to the emitter of the first stage, V5, attenuated by the R19 to C19-R14 voltage divider, increases the input impedance of the audio section to about 10,000 ohms. For this type of feedback, the degree of degeneration is a function of the relative magnitudes of this input-impedance and the signal source-impedance. The particular arrangement of detector and volume control used here was chosen to provide constant source-impedance so that the degeneration (10 db) would be independent of volume control setting. The increasing impedance of C19 at decreasing frequencies determines the low frequency roll-off of the amplifier. High frequency roll-off is controlled at input to audio section by the shunt 0.1-μf capacitor, C16.

The collector of the first audio transistor feeds the base of the second, or driver transistor, V6. Effectively, the collector of the driver works directly into the bases of both output transistors. Actually these bases are separated only by a 22-ohm resistor, R16, which develops a small initial bias. The major portion of the signal current of the driver is made to flow into the output stage bases by returning the coupling resistor, R17, to the common output stage emitters, via C20. Relatively little signal current flows through R17, since only the output stage base-to-emitter voltage need be developed across this resistor, rather than the full output voltage. The dc return path for the driver collector current is R17 and R18 (no appreciable direct current flows into the output bases). The quiescent driver current is made sufficiently large to permit the output bases to be driven to a peak-to-peak voltage equal to the battery supply voltage. The total resistance of resistors R16, R17, and R18 is chosen so that the quiescent voltage developed at the driver collector by this current is approximately battery centertap voltage.

The common emitters of the output stage directly drive the 12-ohm voice coil, which returns to the battery centertap. When the bases of the output stage are driven negative with respect to the battery centertap, the *p-n-p* transistor, V7, conducts, current being fed from the upper half of the battery to the voice coil. The *n-p-n* transistor, V8, and the lower half of the battery function similarly for positive excursions. The peak-to-peak voltage available across the voice coil is less than the battery voltage by the required peak base-to-emitter voltages in the output transistors and by signal voltage developed in the dynamic battery impedance. The experimental output transistors employed are electrically similar to those described by Armstrong and Jenny,[5] requiring approximately 0.5 volt between base and emitter for 145 ma collector current. Thus, with a nominal supply voltage of 5 volts , a maximum peak-to-peak voltage across the voice coil of about 3.5 volts is realized, corresponding to a power output of about 125 mw.

The three audio stages are dc coupled to permit the use of over-all dc feedback for stabilization purposes. Bias for the base of the first transistor is developed by a bleeder, R12 and R13, between the negative side of the battery and the battery centertap. The emitter of this transistor is returned to the battery centertap through R19 and the voice coil. Thus, any dc voltage developed across the voice coil by unbalanced currents in the output transistors is subtracted from this bias, constituting dc feedback. The collector current of the *p-n-p* first stage develops a voltage from base to emitter of the *n-p-n* driver stage, controlling the driver collector current. The magnitude of the driver collector current determines the voltages applied to the output stage bases. When these voltages bracket the battery centertrap voltage, balanced currents flow in the output transistors. Any departure from this balance, due to variations in characteristics among transistors or ambient tempera-

[4] G. C. Sziklai, "Symmetrical properties of transistors and their applications," Proc. IRE, vol. 41, pp. 717–724; June, 1953.

[5] L. D. Armstrong and D. A. Jenny, "Behavior of germanium junction transistors at elevated temperatures and power transistor design," Proc. IRE, vol. 42, pp. 527–530; March, 1954.

ture variation, is of the proper polarity to be self-correcting, through the previously mentioned dc feedback to the first stage. The unbalance is held within ±10 ma from 0 degrees C. to 50 degrees C. For example, the replacement of a normal driver transistor by one having three times the current gain results in an unbalance of only 3 ma. The maintenance of this balance avoids excessive quiescent current drain on either half of the battery supply, and prevents loss of dynamic range due to asymmetrical overload.

The collector current of the first stage is shared by the base of the driver and the shunting resistor, R15. The dc feedback functions to adjust this current to that value which will bias the driver stage to 8 ma collector current. The component of current flowing into the base of the driver stage (for 8 ma collector current) diminishes with increasing temperature, ultimately reversing. The component flowing into R15 remains relatively unchanged. This latter component constitutes essentially the total operating collector current of the first stage near the upper temperature limit of the amplifier. The value of R15 is low enough to insure adequate operating current for the first stage to at least 50 degrees C.

Threshold bias for the output transistors, which operate class B, is developed across R16. The choice of the magnitude of this bias represents a compromise between (a) the appearance of "cross-over" distortion (insufficient bias), and (b) excessive quiescent battery drain (more than sufficient bias). The compromise chosen, good at room temperature, tends toward (a) at temperatures below 10 degrees C., and toward (b) at temperatures above 40 degrees C.

The volume control and the bleeder for the first audio stage are returned to the junction of R20 and C18, decoupling them from the negative supply with respect to the battery centertap, to which the emitter of the input stage ultimately returns. The base and emitter of the driver are both returned to the negative supply, and so require no decoupling. Voltage developed across the internal impedance of the lower half of the battery, appearing at the positive battery terminal and at ground with respect to the centertap, is not directly applied to the audio section. Paths for current do exist, however, through the 910-kc filter in the detector, and through the dynamic detector collector resistance. The audio impedance of these paths is sufficiently high (more than 50,000 ohms) that negligible distortion is introduced.

Power Supply

The battery consists of four separate cells, grouped into two pairs of two cells each to provide a centertapped supply for the output stage. The three-point power switch is located at the battery centertap, where it simultaneously connects the two halves of the battery to each other and to the centertap lead. The use of a three-point switch, instead of a double-pole switch, facilitated the modification of a conventional hearing-aid volume-control-and-*single-pole-switch* combination.

The average battery drain on program material at typical listening level is about 20 ma. The first four stages require 5 ma; the first audio and driver require 8 ma; the output stage draws about 7 ma. The zero-signal current of the output stage is a few ma; the current increases with signal to as much as 50 ma (average) for maximum output on a continuous tone. By the use of the class B output stage rather than a conventional class A output stage the over-all battery requirement has been reduced by a factor of more than three.

As the battery voltage decreases and battery internal impedance increases with operating life, the maximum power output capability of the receiver decreases, ultimately determining the useful battery life. Appreciable degradation of receiver performance in other respects does not occur within this life. The battery life for different types of cells and various maximum undistorted output-power end points is shown in Table I.

TABLE I

BATTERY LIFE CHARACTERISTICS

Cell Type	*RM-1*	*RM-502*	*"Pen-Lite"*
Initial maximum undistorted output power	125 mw	135 mw	150 mw
Life to 75 mw maximum undistorted output power	50 hours	120 hours	20 hours
Life to 50 mw maximum undistorted output power	50 hours	120 hours	36 hours

PERFORMANCE

The major receiver performance characteristics are shown in Fig. 7. These characteristics are generally comparable to those of conventional personal receivers employing vacuum tubes.

The image rejection and sensitivity characteristics are shown in Fig. 7(a). The operation of the antenna circuit at approximately one-half its unloaded Q (matched condition) results in a steepening of the slope of the image rejection characteristic. Approximately 30 db rejection is provided at the upper extreme of the band; rejection increases to 60 db at the lower extreme of the band. The IF rejection (not shown) is 30 db at 600 kc.

The AGC characteristic, and the signal-to-noise ratio as a function of signal level, are shown in Fig. 7(b). The AGC figure of merit is 37 db. The receiver noise performance, depending upon many factors (antenna effectiveness, receiver bandwidth, converter noise, etc.), is not readily compared with a conventional receiver on the basis of any one contributing factor. The signal-to-noise ratio for a given signal level is in the order of 4 db lower than that of a conventional personal receiver.

The IF selectivity, shown in Fig. 7(c), represents a design compromise between selectivity and gain for the three single-tuned circuits employed. The adjacent channel attenuation is 20 db. A requirement of more selectivity might justify the substitution of one or more double-tuned circuits.

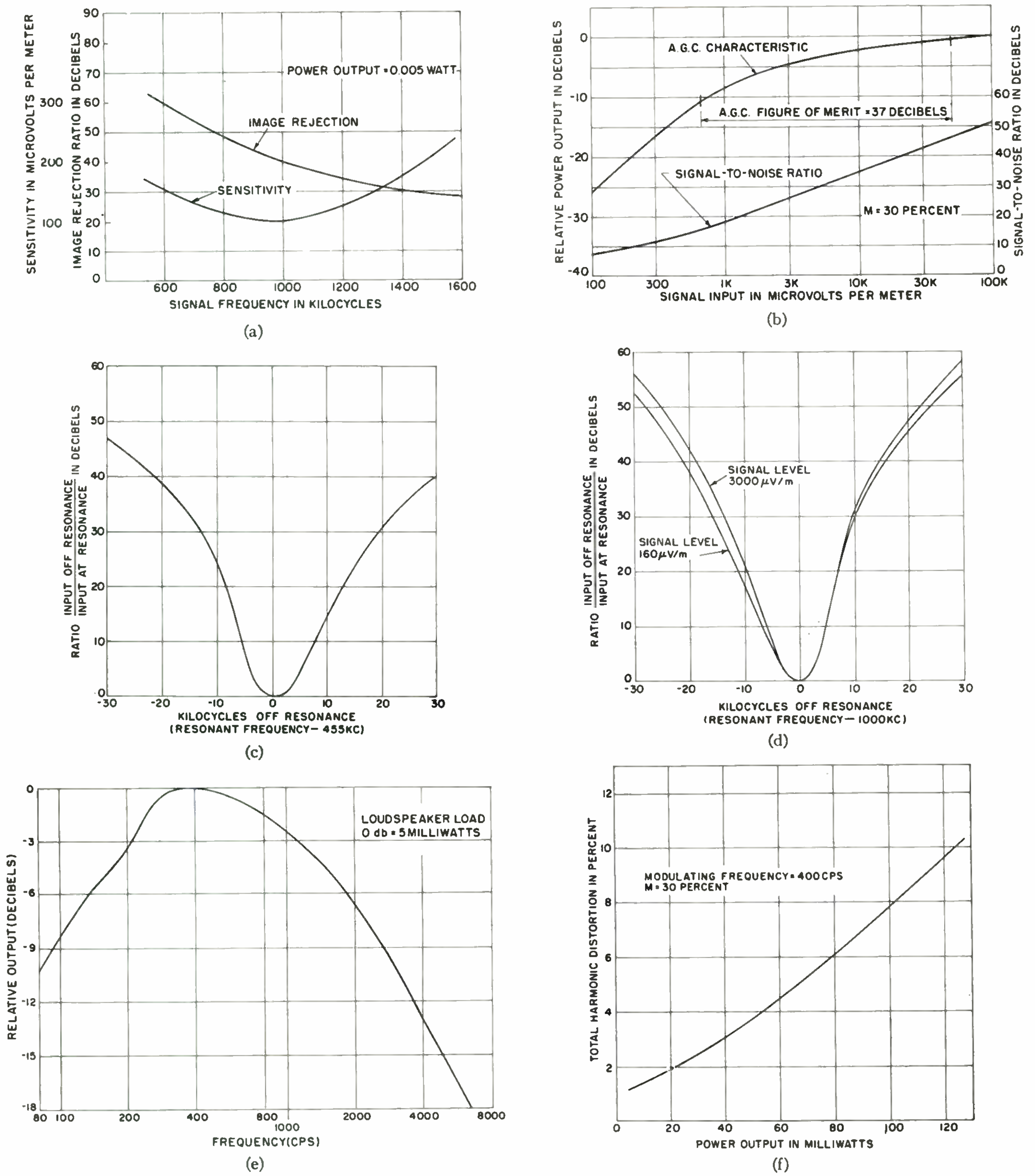


Fig. 7—Major receiver performance characteristics. (a) Image rejection and sensitivity. (b) Signal-to-noise ratio and AGC characteristic. (c) IF selectivity. (d) Over-all selectivity. (e) Electrical fidelity. (f) Distortion characteristic.

The over-all selectivity characteristic is displayed in Fig. 7(d). Over-all selectivity curves for two typical signal levels are included, demonstrating that the selectivity change due to detector and AGC action is small.

The electrical fidelity of the receiver is shown in Fig. 7(e). The high frequency roll-off has been adjusted in listening tests to provide a pleasing tonal balance.

The distortion-vs-power-output characteristic is shown in Fig. 7(f). This characteristic is influenced by the battery complement employed; the maximum power output obtained increases approximately as the square of the battery voltage, and decreases with increased battery internal impedance. For a battery complement, 4 RM-1 cells, (5.0 terminal volts under load) as shown, the maximum power output (10 per cent distortion) is approximately 125 mw. Maximum power output achieved with 4 penlite cells (5.6 terminal volts under load) for example, is approximately 160 mw.

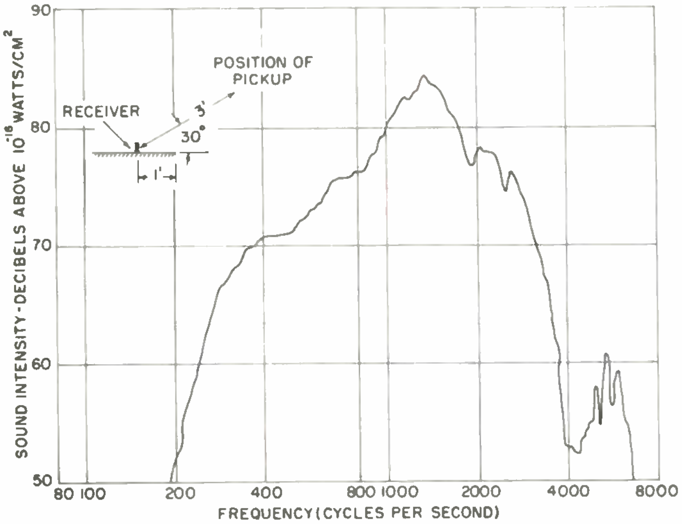


Fig. 8—Acoustical performance.

The over-all acoustical performance of the receiver appears in Fig. 8. The curve shown was obtained in a free field sound room with the receiver placed on a table, as indicated in the sketch.

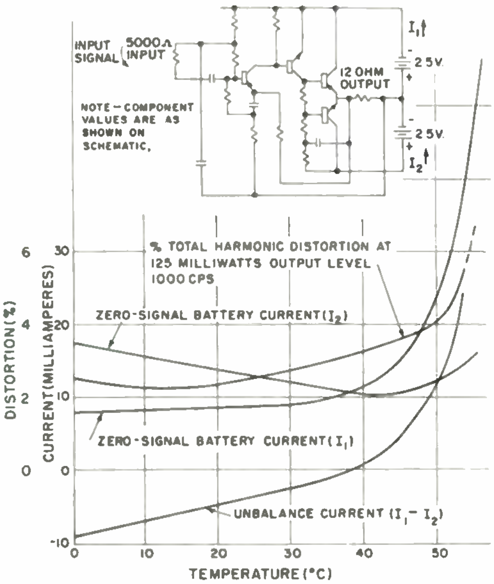


Fig. 9—Audio amplifier performance vs ambient temperature.

The performance of the converter, IF amplifier and AGC circuits is substantially unaffected by a variation in ambient temperature over the range from 0 degrees C. to 50 degrees C. The detector (and thus the receiver) sensitivity varies approximately 2 to 1 over this range, being higher at elevated temperatures. The performance of the audio amplifier as a function of ambient temperature is shown in Fig. 9. These data were taken on an experimental "breadboard" while the receiver was under development. The unbalance current remains within limits of approximately ±10 ma over the temperature range from 0 to degrees C. 50 degrees C. The distortion, for a constant output of 125 mw, is substantially unaffected over this temperature range.

Physical Details

Front and rear views of the receiver with the covers removed are shown in Fig. 10. The ferrite-cored antenna may be seen at the top of a "chassis" which is essentially vertical and to which are attached, by means of brackets, the tuning condenser, speaker, and battery compartment. The eight transistors, mounted in subminiature sockets, occupy the space immediately below the antenna. The miniature oscillator and IF transformers are mounted on the chassis, which serves as an electrical ground for the rf and IF portions of the receiver. The top of the chassis is bent over to form a horizontal flange. This flange, the subminiature sockets, a copper shield, fibre spacers, and the antenna are cemented together to form an integral sandwich.

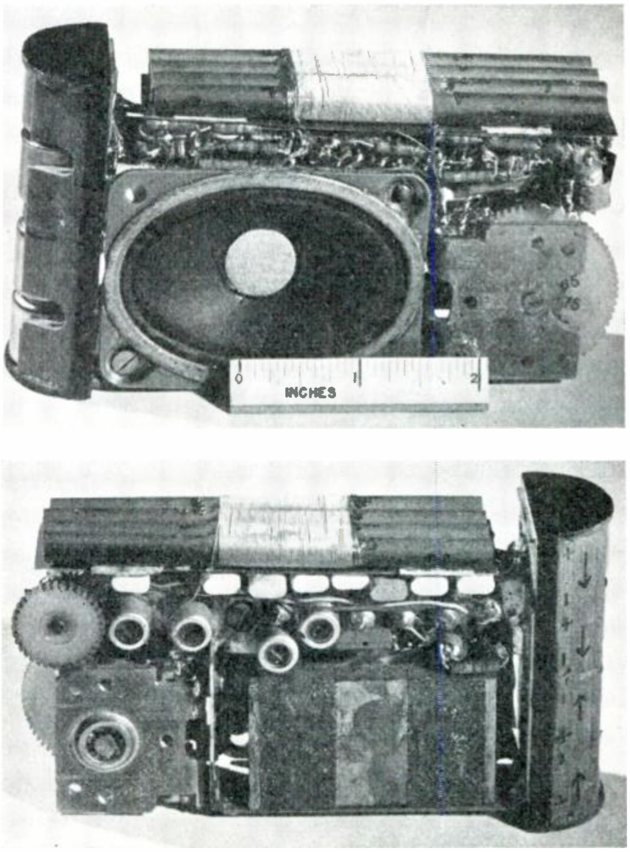


Fig. 10—Front and rear views of the receiver with covers removed.

The tuning capacitor, supplied by Radio Condenser Corporation, Camden, N. J., is a modification of a capacitor employed in a conventional personal-portable receiver design. The oscillator and rf sections employ four stator plates each instead of five and seven respectively, providing the requisite reduction in capacitor depth. The resultant reduction in the ratio of maximum to minimum capacitance can be tolerated because of the relatively low shunt capacitance reflected across the antenna and oscillator-tuned circuits by the converter transistor. The tuning wheel is mounted on the rotor shaft, within the frame of the condenser.

The speaker is a modified version of the RCA 214S1 unit. Standard parts were used except for the field structure, arranged as shown in Fig. 11 at right. The structure was magnetized, after assembly, by the impulse method, using a ten-turn coil on each magnet (series-opposing) and a 4,000-ampere discharge.

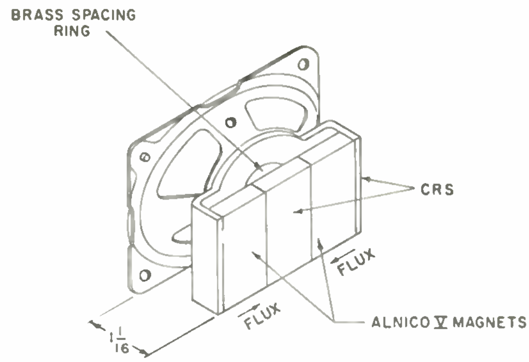


Fig. 11—Speaker, showing modified field structure.

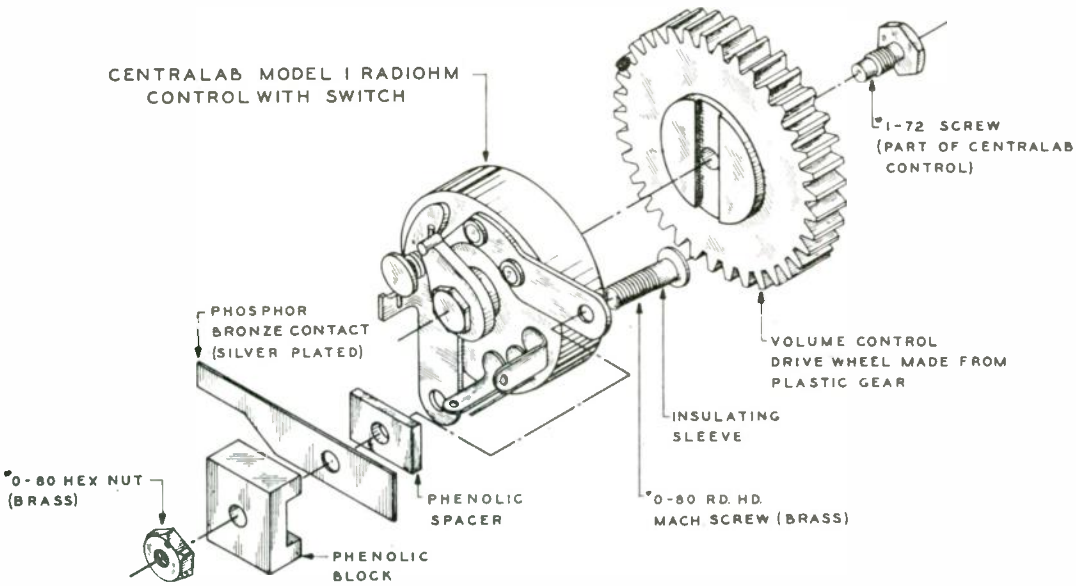


Fig. 12—Exploded view of modified volume-control-and-switch.

The volume control as acquired (Centralab Model 1 Radiohm, with switch) is equipped with a SPST switch, modified as shown in Fig. 12 to provide desired 3-point switch. Switch action is such that the movable contact is forced between the two stationary contacts.

Three small 0.1-μf ceramic capacitors of an experimental type are employed. The small size ($0.2\times0.50\times0.03$ inch) is achieved through the use of partially-reduced barium-strontium titanate as the dielectric.

The 30-μf electrolytic capacitors are of the sintered-tantalum type which is notable for its high capacitance-to-volume ratio. These units were mounted by pressing them into holes in the chassis plate. In those instances where it was necessary that the outer case of the capacitor be insulated from ground, the unit was first covered with a plastic sleeve.

The battery compartment and receiver case shown here were milled from linen-base phenolic stock. For convenience, the receiver case was milled in two sections which were then cemented together to form a slide-on unit which fits flush with the battery case and is screw-fastened to a bracket on the bottom of the receiver chassis. Other experimental receiver cases have been constructed using resin-impregnated fibre-glass cloth.

An Experimental Automobile Receiver Employing Transistors*

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Summary—**This paper describes an experimental automobile broadcast receiver utilizing nine experimental junction transistors in a superheterodyne circuit. The receiver operates directly from the six-volt storage battery without vibrator, power transformer, or rectifier. The average current drain, including that for two pilot lights, is approximately one-tenth that of a conventional automobile receiver.**

The performance of this receiver is comparable to that of conventional automobile receivers. Particular emphasis has been placed on maintaining performance over a wide range of ambient temperature, both to accommodate the severe requirements specified for automobile service, and to establish the operability over such a temperature range of apparatus employing *germanium* transistors.

The receiver circuits and performance characteristics, including performance data for the ambient temperature range −40 degrees C. to +80 degrees C., are described in detail. Techniques which render circuit operation insensitive to variation of ambient temperature and which permit interchangeability of transistors are discussed.

General Description

THE AUTOMOBILE receiver, shown in Fig. 1, uses nine experimental *p-n-p* alloy junction transistors in a superheterodyne circuit employing a 455-kc intermediate frequency. A permeability-tuned rf stage and a class-B transformer-coupled output stage are incorporated. The current drain of the radio is 250 ma for the zero signal condition; an additional 300 ma are required by the two pilot lights. On a sustained tone, with maximum power output, total drain rises to one amp.

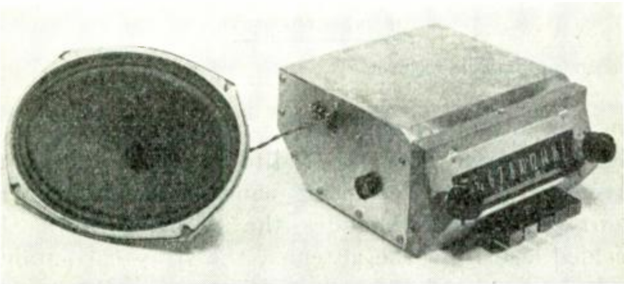


Fig. 1—Experimental transistor automobile receiver.

The performance characteristics of the receiver at 20 degrees C. are summarized below:[1]

Sensitivity	2	μvolts
Noise Performance		
Input for 20 db S/N	12	μvolts
ENSI	0.4	μvolts
Power Output	2	watts
Selectivity (ACA)	41	db
AGC (Figure of Merit)	63	db

The operating temperature range of the receiver is −40 degrees C. to +80 degrees C. Circuit techniques provide stabilization for the most part, and thermistors are used for temperature compensation in the audio amplifier.

The receiver is constructed in three main sections as shown in Fig. 2. These sections are the tuner assembly, in the foreground; the audio amplifier, mounted over the tuner; and the high-frequency part of the receiver, in the background.

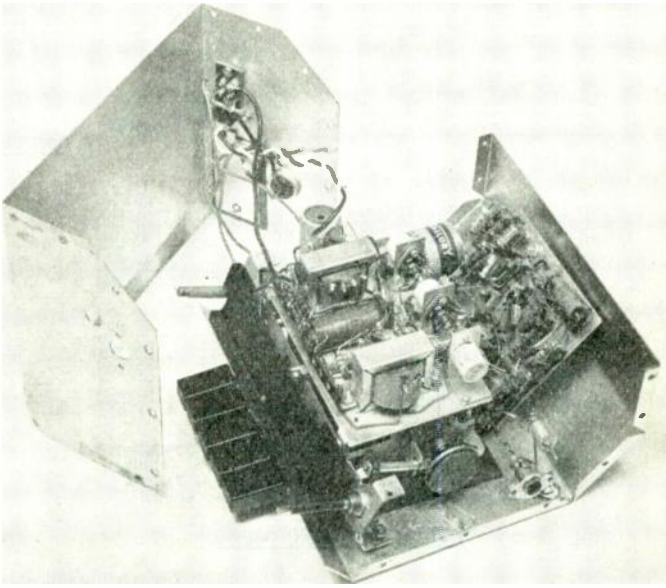


Fig. 2—Receiver with cover removed.

Circuit Description

A schematic diagram of the receiver is shown in Fig. 3 (on the next pages). Transistors V1 through V6 are experimental radio-frequency units[2] and serve as rf stage, mixer, oscillator, two IF stages, and second detector. The audio complement, V7 through V9, are experimental *p-n-p* transistors electrically similar to those described by Armstrong and Jenny[3] but incorporate a mechanically and thermally improved mounting arrangement.

Tuner Transformers

The electrical elements of a push-button permeability tuner manufactured by Radio Condenser Corporation, Camden, N. J., were revised for operation in the transistor receiver. Three tightly-coupled transformers are em-

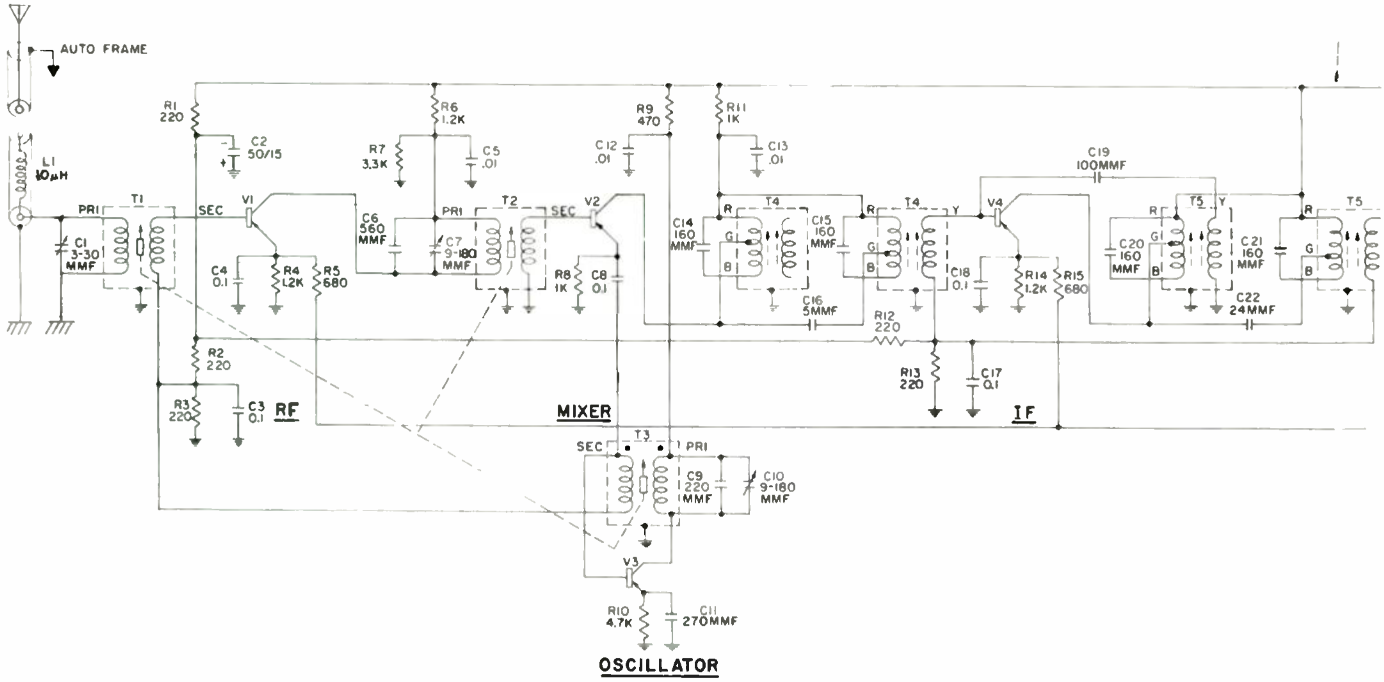
* Original manuscript received by the IRE, January 26, 1955.

† RCA Laboratories Div., Princeton, N. J.

[1] These data reflect the characteristics of the experimental transistors used; units of exceptionally high and exceptionally low performance were avoided. Long-term life performance of the transistors was not evaluated.

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

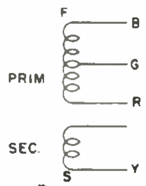
[3] L. D. Armstrong and D. A. Jenny, "Behavior of germanium junction transistors at elevated temperatures and power transistor design," Proc. IRE, vol. 42, pp. 527–530; March, 1954.



UNIT	PRIMARY T	PRIMARY LITZ	SECONDARY T	SECONDARY SSE	FORM I.D.	FORM O.D.	GEAR SETTINGS FRONT	GEAR SETTINGS CAM	GEAR SETTINGS REAR
T1	322	4-43	5	#35	.187	.205	43-40	.125	28-100
T2	107	5-42	28	#35	.187	.205	50-40	.125	88-100
T3	102*	7-41	14	#35	.187	.250	43-40	.125	
	20								52-60
	25								46-60
	20								72-60
	20								65-60
	17								71-60

WOUND ON MEISSNER MACHINE - PRIMARY PROGRESSIVE UNIVERSAL ON BAKELITE FORM SECONDARY DISTRIBUTED HAND WOUND OVER PRIMARY. *THE PRIMARY OF T3 IS A STEPPED-PITCH WINDING, THE STEPS ARE AS SHOWN

TUNER TRANSFORMERS



UNIT	SEC	PRIM R	PRIM G	PRIM B
T4	4T	OT	55T	110T
T5	4T	OT	32T	110T
T6	14T	OT	40T	110T

*38 DSE ON 0.125 I.D. x 0.145 O.D. FIBRE FORM, RANDOM WOUND ON 0.230 WIDE x 0.125 D. ARBOR. SECONDARY WOUND UNDER PRIMARY-DISTRIBUTED OVER LENGTH OF COIL FINISHED COIL CUMAR DOPED.

Transformer cores and mechanical assembly miniature IF transformers described (experimental).

IF TRANSFORMERS

Fig. 3—(*left half*) Schematic of transistor automobile receiver.

ployed in the antenna-to-rf input, rf-to-mixer interstage, and oscillator circuits. Maximum unloaded Q of these transformers is obtained by making the effective diameter of the windings as large as the tuner dimensions will permit. Tracking of the signal circuits is insured by maintaining equal effective diameters of the respective coils. Oscillator tracking is accomplished by approximating a variable-pitch oscillator-transformer winding, the desired variable pitch being obtained by changing the coil-winder gear ratio at intervals along the coil. The pertinent data for the antenna, interstage, and oscillator transformers are shown on the schematic.

Individual coil assemblies of tuner are enclosed in shield cans. Magnetite sleeves of $\frac{3}{8}$-inch inner diameter are provided for antenna and interstage transformers. Powdered-iron tuning slugs are 1.2 inches long by 0.18 inch diameter; slug travel is about 1 inch.

Tuner Circuitry

A conventional automobile rod antenna is employed; this antenna feeds the tuned primary of the antenna transformer, T1, as shown on the schematic, Fig. 3. Since the antenna is effectively a voltage source in series with a capacitor, the power fed to T1 will increase with increasing Q and decreasing shunt capacitance. The shunt capacitance includes the capacitance of the shielded lead from the antenna, the stray capacitance of the wiring, and the reflected rf stage input capacitance. The coil Q is limited by the winding dimensions to 50–70 across the band. The primary is designed to tune with a total of 75 $\mu\mu$f, allowing for the above shunt capacitance as well as the antenna capacitance and a trimmer capacitor, C1, of 3 to 30 $\mu\mu$f.

The choice of operating Q for the antenna and interstage transformers represents a compromise between the rejection of image and intermediate frequencies, which increases with operating Q, and insertion loss, which also depends on operating Q. Mimimum insertion loss between the antenna and the rf stage input obtains for the "matched" condition, i.e., when the turns ratio of T1 is adjusted so that the transformer tuned impedance, referred to the secondary winding, is equal to the input impedance of the rf stage. For this condition, the operating Q is one-half the unloaded coil Q. For the compromise

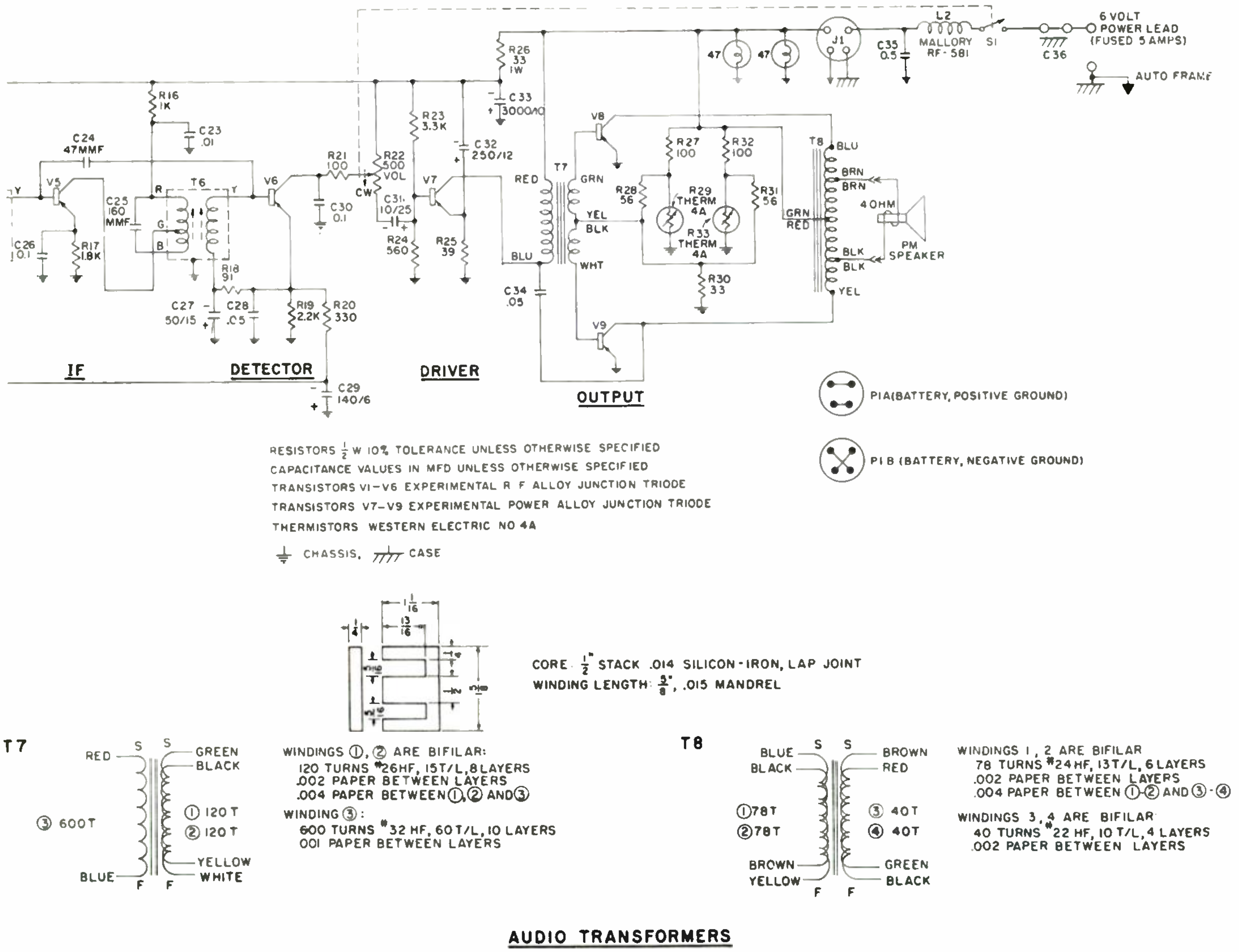


Fig. 3—(*right half*) Schematic of transistor automobile receiver.

used, the operating Q is eight-tenths of the unloaded Q, with an associated increase in insertion loss of 2 db above that applying for the matched condition. This compromise is obtained by adjusting the turns ratio so that the secondary tuned-impedance is one-fourth the input impedance of V1. A typical value for the input impedance of V1 at midband is 75 ohms.

The rf stage bias arrangement renders the stage relatively insensitive to changes in ambient temperature. The base is returned to a low resistance bias source of -1.5 volts at the junction of R2 and R3. Below AGC threshold, constant-emitter-current bias of 1.3 ma obtains, via the emitter resistor, R4, in conjunction with R5, R19 and R20. Emitter is returned to ground for rf by C4. Bias conditions with respect to AGC action are discussed in connection with the second detector.

The rf interstage transformer, T2, provides coupling from the collector of V1 to the base of the mixer, V2. In the mid-frequency range of the broadcast band the output impedance of the rf transistor is 10,000 to 15,000 ohms, and the mixer input impedance is typically about 500 ohms. The operating Q of T2 is 15 to 20 and the transformer insertion loss is 3.7 db. This compromise between operating Q and insertion loss is obtained by adjusting the turns ratio of T2 so that the tuned primary impedance of T2 and the reflected input impedance of the mixer each equal the output impedance of V1. Tuning of the primary of T2 is provided by C6 and C7. The gain of the rf stage is about 20 db at midband.

The collector circuit of V1 returns to the tap on the bleeder formed by R6 and R7. The bleeder tap is bypassed to ground by C5. This arrangement decouples the collector circuit from the common supply, and serves as a voltage divider to reduce the collector voltage of the rf stage. An improved signal-to-noise ratio is obtained by operation at reduced collector voltage.[4]

The tuned primary of the oscillator transformer, T3, in the collector circuit of the oscillator transistor, V3, affords an unloaded Q of from 40 to 60 over the oscillation frequency range. A relatively high tank capacitance is employed for stability. The number of secondary turns was determined experimentally for adequate

[4] Mueller and Pankove, *loc. cit.*

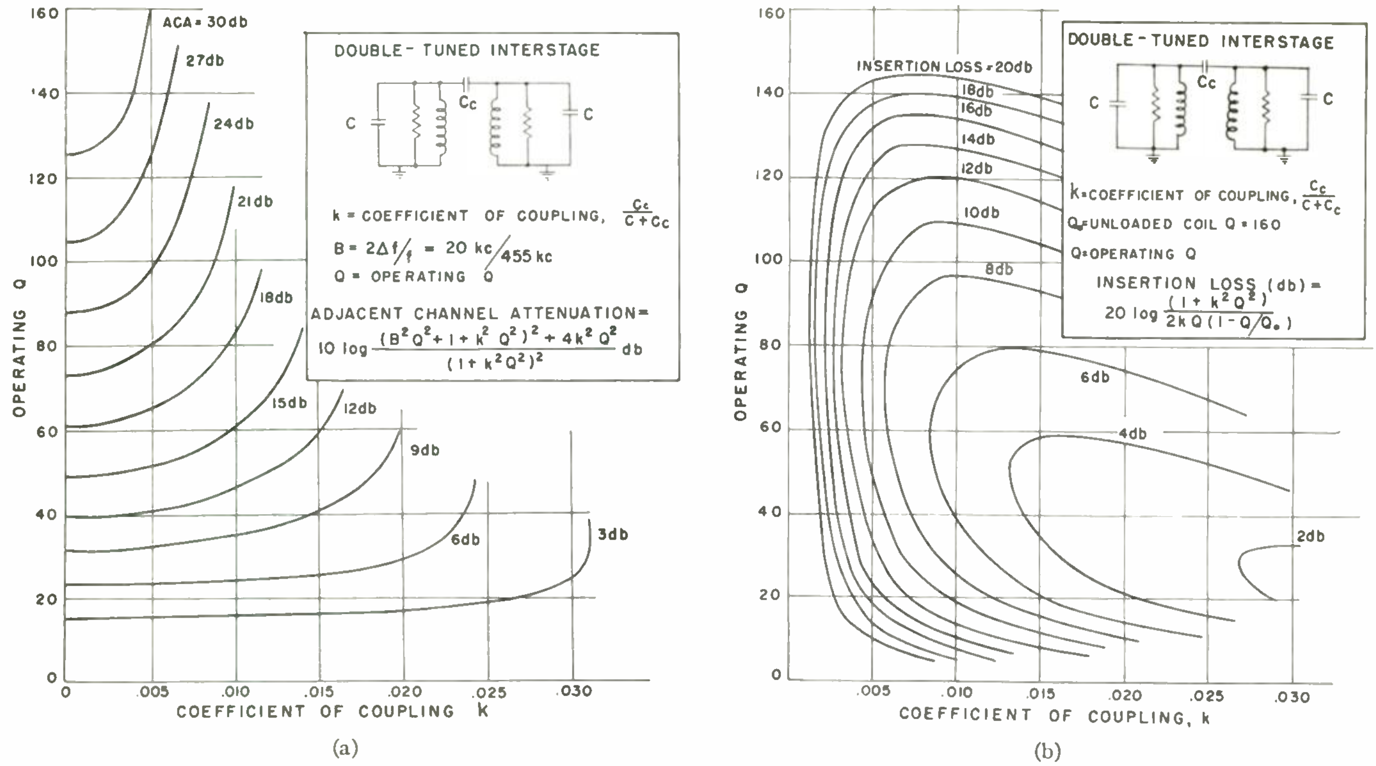


Fig. 4—Design curves for double-tuned IF coupling transformers. (a) Variation of adjacent channel attenuation with operating Q and coefficient of coupling, for IF = 455 kc, channel separation = 10 kc. (b) Variation of insertion loss with operating Q and coefficient of coupling, for unloaded $Q = 160$.

mixer injection. The secondary applies feedback to the base of V3 and is returned to the −1.5 volt bias source at the junction of R2 and R3. The R10-C11 network in the emitter circuit of V3 introduces degeneration which reduces the net positive feedback in the oscillator circuit. The loading of the oscillator tuned circuit by the oscillator transistor input circuit is thus reduced so that oscillator tuning becomes relatively independent of the oscillator transistor input impedance. The reactance of the effective transistor base-to-emitter capacitance is in series with the relatively high reactance of C11. Thus, variation of the effective transistor base-to-emitter capacitance with frequency does not deteriorate oscillator tracking. (A decrease of the effective transistor base-to-emitter capacitance with increasing frequency arises from the presence of transistor base-lead resistance, which is in series with the emitter-junction capacitance. In the 1 to 2 mc range, the base-lead resistance and the reactance of the emitter-junction capacitance are comparable in magnitude.) The resistor R10, in conjunction with the base bias, provides sufficient starting emitter current to initiate oscillation.

The secondary of the interstage transformer, T2, is coupled to the base of the mixer, V2. The emitter is returned to ground by R8, which provides bias stability in a manner analogous to that provided by the emitter-return resistors of the amplifier stages. Approximately 0.4 volt rms of oscillator injection is applied to the emitter through capacitor C8; the corresponding average emitter current is 0.4 ma. The optimum magnitude of oscillator injection depends in part on the magnitude of the emitter-return resistor; in this instance optimum injection is typically 0.35 volt rms. If injection decreases below this value, the conversion gain falls rapidly, while the conversion gain decreases relatively slowly with increasing injection. Somewhat greater than optimum injection insures interchangeability of mixer transistors and minimizes variation of conversion gain with small changes in oscillator injection.

Since the coupling capacitor, C8, presents a low impedance at the intermediate frequency and the secondary of T3 has a low impedance at both signal and intermediate frequencies, R8 is effectively bypassed to ground for both the mixer input and output signals.

The mixer output impedance is typically 60,000 ohms and is relatively independent of signal frequency, while the input impedance is about 500 ohms at midband and decreases somewhat with increasing frequency. The mixer stage conversion gain is about 20 db at midband.

IF Amplifier[5]

The three IF interstage coupling networks consist of two capacitively-coupled double-tuned transformers, T4-T4 and T5-T5 and a single-tuned transformer T6. These transformers embody an experimental arrangement and are ferrite cored. They are approximately $\frac{3}{8}$ inch in diameter by $\frac{1}{2}$ inch high. The IF amplifier con-

[5] The IF, detector, and AGC circuits are similar to those described by the authors in a paper entitled "A developmental pocket-size broadcast receiver employing transistors," p. 662, this issue.

tributes 37 db of adjacent channel attenuation (ACA); the selectivity provided by the various interstage networks is so apportioned as to minimize over-all insertion loss for this ACA.

For the double-tuned circuits, the ACA is determined by the coefficient of coupling, k, as well as the operating Q's, as shown in Fig. 4(a). The insertion loss of the double-tuned circuits is determined by k, by the operating Q's, and by the unloaded coil Q, as shown in Fig. 4(b). That optimum combination of k and operating Q which results in minimum insertion loss for a prescribed ACA may be determined from these curves. The variation of minimum insertion loss vs ACA is shown in Fig. 5; a corresponding curve for a single-tuned circuit is plotted for comparison.

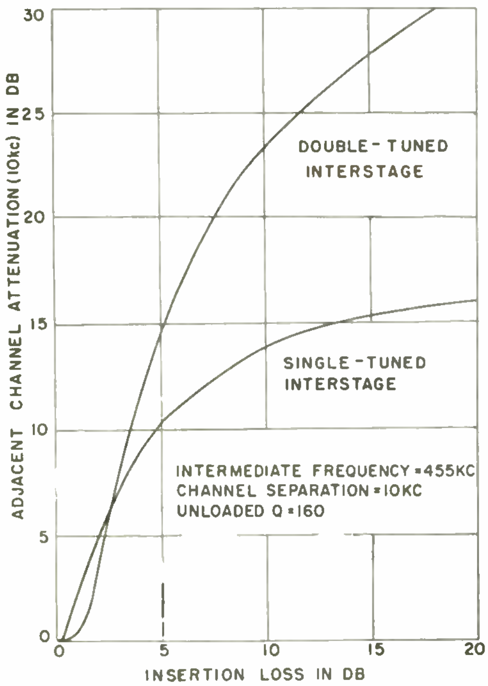


Fig. 5—Minimum insertion loss vs ACA for double- and single-tuned circuits.

The ACA's provided by T4-T4, T5-T5, and T6 are 15.5 db, 15.5 db, and 6 db respectively, and the insertion losses are 5 db, 5 db, and 2.5 db respectively. The amplifier provides about 50 db gain from the base of the first IF amplifier, V4, to the base of the second detector, V6.[6]

Biasing of the first IF stage, to which AGC is applied, is similar to that of the rf stage. The second IF stage is constant-emitter-current biased by means of the emitter resistor, R17. The IF stage bases return to the tap on a separate bleeder formed by R12 and R13. The bleeder tap is bypassed to ground by C17. This bleeder circuit, in combination with the collector-circuit decoupling of the second IF stage provided for by R16-C23, serves to isolate the IF amplifier from the rf and the mixer stages. Neutralization of the IF stages is provided for by C19 and C24.

[6] By way of example, IF transistors having $r_{bb'}=75$ ohms, $\alpha_{cb}=20$, $C_{b'c}=11$ μμf, and $C_{b'e}=0.001$ μf ("alpha cutoff" of 6 mc) would give typical performance.

Second Detector and AGC

As shown in the schematic, AGC is applied to the rf and first IF stages. Emitter current control is employed, and is obtained from the audio-AGC detector, V6, via R20, R15, and R5. The bias arrangement of V1 and V4 holds the currents in R4 and R14 essentially constant; the controlled stages are constant-emitter-current biased for the zero-signal condition. An increase in the detector collector current produces a proportionate decrease in the emitter currents of the controlled stages. The magnitude of these currents at zero signal (about 1.5 ma per stage) is sufficient that the initial decrease in current results in little change in gain, introducing the desired delay in AGC action. The control current at the flat portion of the AGC characteristic is 3 ma; an additional 1 ma of detector current flows through the shunt resistor, R19. The total detector current of 4 ma is then sufficiently large to minimize the effects of elevated-temperature saturation current in the detector transistor. In the design of an AGC circuit of this type, adjustment of the zero-signal emitter currents of the controlled stages provides a convenient level control of the flat portion of the AGC characteristic.

The rc networks in emitter circuits of detector, rf, and first IF amplifiers provide filtering of rf and audio components of AGC current generated in the detector.

Detector linearity is improved by degeneration provided by R18, in the return path for the audio component of detector emitter current. Audio output is taken from the collector through the volume control, R22. Decoupling from the collector supply is provided by R21 and C30.

Audio Amplifier

The audio output of the detector is applied, via C31, to the base of the audio driver, V7. Since the dynamic output impedance of the detector transistor is high, the output impedance of the detector stage is essentially the resistance of the volume control irrespective of volume control setting. The current available for driving the following stage is therefore proportional to the resistance intercepted between the slider and the common supply. Nearly all of this available output current flows into the base of the driver, since the driver presents an input impedance on the order of 75 ohms. A base-to-collector signal current gain of 40 to 50 is realized. For medium audio frequencies, the emitter of the driver is bypassed, by C32, to the common supply. The low-frequency response of the amplifier is down 3 db at that frequency at which the input impedance at the base of the driver (the product of the driver current gain and the reactance of C32) equals the parallel resistance of the volume control, R22, and the base-bias bleeder resistors, R23 and R24. The roll-off frequency, for operation into a 4-ohm dummy load, is approximately 120 cps. Response into speaker load remains elevated to below 80 cps due to mechanical low-frequency resonance of loudspeaker.

Control of the driver stage operating current is provided by the base-bias bleeder in conjunction with the emitter resistor R25. The driver collector current is approximately 15 ma at moderate temperatures; it increases to 30 ma at 80 degrees C. and drops to 10 ma at −40 degrees C.

The driver collector is transformer-coupled to the bases of the push-pull output stage by transformer T7. The interstage and output transformer data are given on the schematic.

In addition to the usual transformer design compromises, (efficiency, low-frequency response, etc.) the following requirements are met in the design of the interstage transformer:

1. The impedance reflected to the driver collector is low enough that driver overload does not occur before overload of the output stage.
2. The series resistance of the primary winding is low enough that the loss of driver supply voltage is tolerable.
3. The two halves of the secondary winding are sufficiently tightly coupled together to avoid transient voltages when current shifts from one output transistor to the other. This is accomplished by bifilar winding of the secondary.

A current gain of 5 is afforded by the interstage transformer from the collector of the driver to the bases of the output transistors. Nonlinearity in the cross-over interval is minimized by providing the output transistors with a small initial threshold bias. Experiment indicates that there is an optimum value of threshold emitter current, about 20 ma, which results in minimum nonlinearity. While this optimum value of current is essentially independent of temperature, the corresponding required base-to-emitter bias voltage varies with temperature at a rate of approximately −0.0025 volt per degree C. The resistor-thermistor networks, R27 through R33, provide a low-impedance bias source exhibiting the proper variation of voltage with temperature. The transistor-thermistor mounting arrangement,

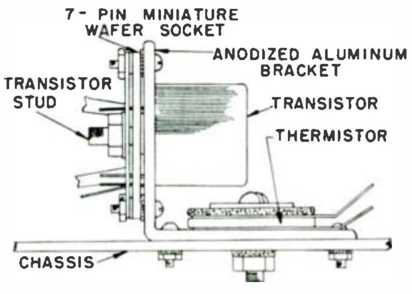


Fig. 6—Power-transistor and thermistor mounting arrangement.

shown in Fig. 6, provides for close thermal contact between the transistor case and mounting bracket, and thermistor, mounting bracket, and chassis.

The output stage affords large-signal current gain of 20 to 30, and transconductance in excess of 1 mho.[7]

[7] These characteristics were obtained with experimental transistors having substantially better high-current performance than that described by Armstrong and Jenny, *loc. cit.*

With a 6.6-volt collector supply voltage, a collector voltage swing of 6 volts peak is obtainable. The corresponding peak collector current, for 2.0 watts output (10 per cent distortion), is 0.67 ampere. The 4-ohm speaker voice-coil impedance is transformed to the required 9-ohm collector load by transformer T8. The use here of a bifilar wound autotransformer provides close coupling and high transformer efficiency. The power gain of the output stage is on the order of 24 db.

High-frequency roll-off of the audio amplifier is controlled by capacitor C34, which introduces inverse feedback in the output stage at frequencies above 2 kc.

The audio input circuits are decoupled from the power supply by R26 and C33.

Power and Interference Considerations

Conventional means are utilized to eliminate high-frequency interference on the power lead. A spark plate, C36, and an rf choke and capacitor L2, C35, remove the vhf and mf components respectively. The receiver chassis is insulated from the receiver case. Power connection for operation in automobiles with positive or negative polarity battery ground is obtained by employing plug P1A or P1B, respectively.

Rejection of high-frequency impulse type of ignition interference appearing on the antenna is accomplished by the choke, L1, in series with the antenna lead, which, together with the shunt capacitance across the antenna primary, forms a low-pass filter.

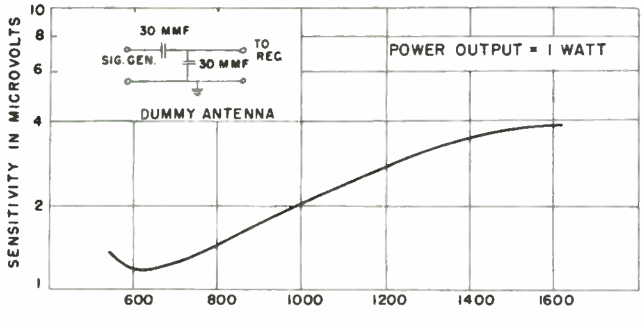
An additional potential source of interference arises from the periodic current drawn by the automobile ignition system, which may produce a low-frequency voltage fluctuation on the power lead. The presence of more than a few millivolts of this type of interference appearing between bases and emitters of the gain-controlled stages would produce objectionable modulation of the received signal. This is avoided by returning the bias bleeders, as well as the collectors of V1 through V6, to the decoupled voltage available at the junction of R26 and C33, and by the additional isolation of the base returns provided by C2.

Performance

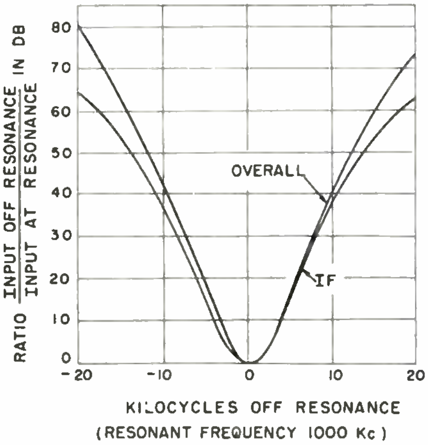
Various receiver performance characteristics are shown in Fig. 7 on the facing page.

The receiver sensitivity as a function of signal frequency appears in Fig. 7(a). The increase in sensitivity at the lower extreme of the broadcast band is due largely to the increased gain of the rf and mixer stages over that obtaining at the upper extreme of the band. The shape of this curve is also influenced by the tracking error of the oscillator transformer which is plus-or-minus 4 kc. This tracking error is due in part to the step approximation to a variable-pitch winding employed in this transformer.

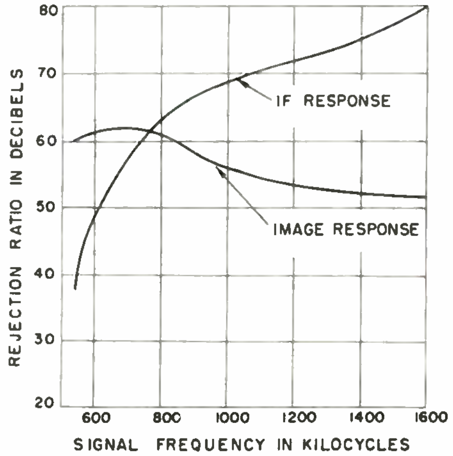
The IF and over-all selectivity at 1,000 kc, and the IF and image rejection as a function of receiver tuning are shown in Figs. 7(b) and 7(c) respectively. These curves



(a)

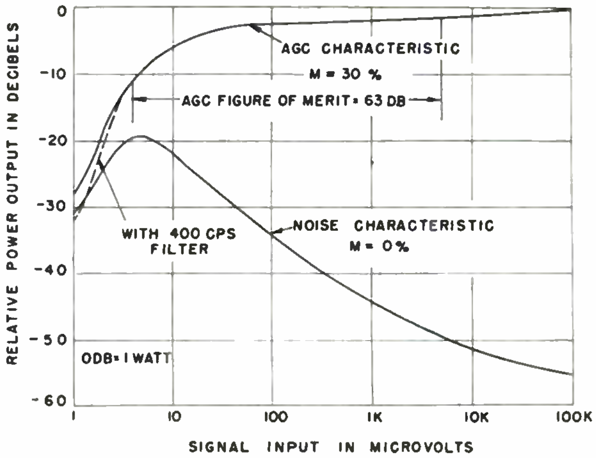


(b)

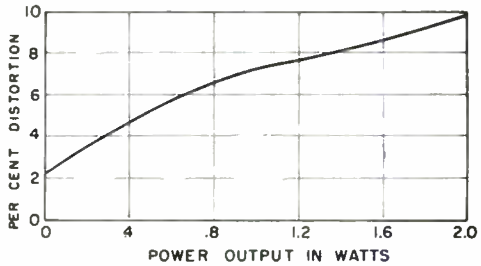


(c)

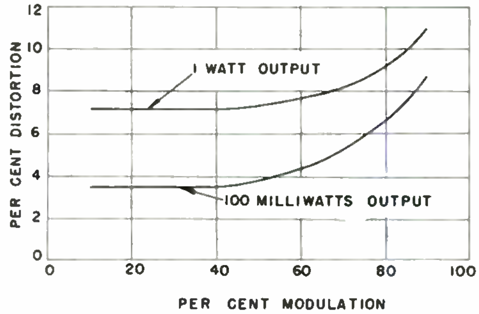
Fig. 7 (*above and right*)—Receiver performance characteristics. (a) Sensitivity vs signal frequency. (b) IF and over-all selectivity. (c) IF and image rejection characteristics. (d) AGC and noise characteristics. (e) Distortion vs power output. (f) Distortion vs per cent modulation. (g) Electric fidelity.



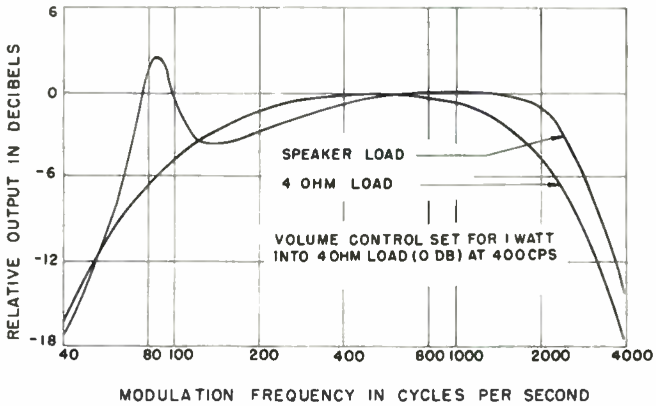
(d)



(e)



(f)



(g)

reflect the design compromises in the antenna, interstage, and IF coupling networks. The over-all adjacent channel attenuation at 1,000 kc is 41 db. Since the over-all selectivity is determined almost entirely by the IF coupling networks, the ACA is essentially independent of the signal frequency. The image rejection at 1,600 kc is 42 db and increases to about 60 db at the low end of the band. The IF rejection provided at 540 kc is 40 db and increases rapidly with increasing signal frequency.

The AGC and noise characteristics are shown in Fig. 7(d). The AGC figure of merit is 63 db. The signal input required for a 20-db signal-to-noise ratio is 12 μvolts; the corresponding equivalent-noise-sideband input is 0.4 μvolt.

The distortion vs power-output characteristic is displayed in Fig. 7(e). Total harmonic distortion reaches 10 per cent, mostly third harmonic, at approximately 2 watts output. At this level, distortion arises almost entirely from the curvature of the current-gain characteristics of the output transistors. Since the power output is not limited by the dissipation capabilities of the transistors, operation in automobiles with 12-volt electrical systems should permit a substantial increase of power output, utilizing the same transistors at the same peak currents.

The distortion vs per cent modulation characteristics for 1.0 watt and 100 milliwatts output are shown in Fig. 7(f). The distortion is substantially constant up to 50 per cent modulation and increases above this level due to an increasing detector contribution.

The electric fidelity into the speaker load and into a 4-ohm dummy load is displayed in Fig. 7(g). The mechanical resonance of the 6×9-inch speaker and the series inductance of the voice coil account for the elevated response into the speaker-load in the vicinity of 85 and 2,000 cps respectively. Volume-control compensation has not been employed; the electric fidelity is essentially independent of volume-control setting. The receiver acoustical performance is influenced by the effective baffle which is employed, and thus is different for various makes and models of automobiles.

Curves illustrating the performance of the receiver as a function of ambient temperature are shown in Fig. 8.

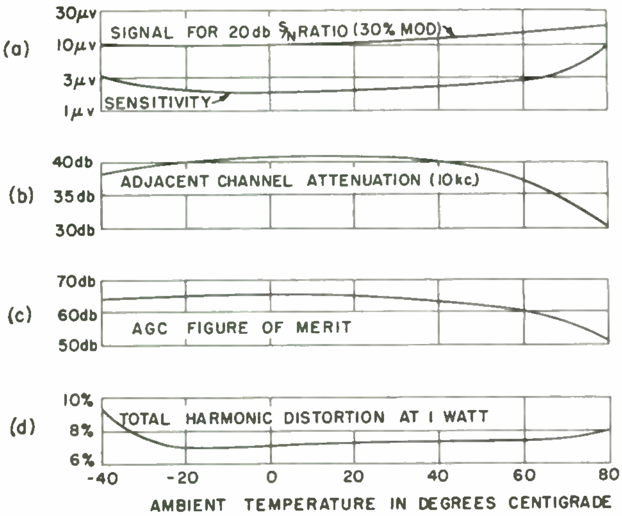


Fig. 8—Receiver performance as a function of ambient temperature.

Fig. 8(a) shows the receiver sensitivity and the signal required for 20-db signal-to-noise ratio as a function of ambient temperature. The sensitivity is 2.0 μvolts at 20 degrees C., and is below 10 μvolts over the range from −40 degrees C. to 80 degrees C. Loss of sensitivity at high temperatures arises from the reduced Q of the core materials in the tuned circuits, from detuning effects, and from a decrease in gain of the high-frequency transistors of approximately 1 db per stage. The receiver noise performance is substantially unaffected over this range of temperature.

Fig. 8(b) shows the receiver ACA as a function of ambient temperature. Differences in tuned-frequency shift of the individual IF interstages tend to reduce selectivity at the extreme temperatures. This tendency is offset by the increase in Q's at reduced temperatures, and aggravated by the reduction in Q's at elevated temperatures, resulting in a reduction in ACA from approximately 40 db, at low and moderate temperatures, to 31 db at 80 degrees C.

The variation of the AGC figure of merit with ambient temperature is shown in Fig. 8(c). The figure of merit is approximately 65 db at low and moderate temperatures, and drops to 51 db at 80 degrees C. The shape of the AGC characteristic at any temperature in this range is essentially the same as that shown in Fig. 7(d). The reduction in AGC figure of merit at elevated temperatures arises from the corresponding reduction in receiver sensitivity.

The total harmonic distortion at 1.0 watt output, shown in Fig. 8(d), and the maximum power output (10 per cent distortion), are relatively unaffected over the temperature range from −40 degrees C. to 80 degrees C. The increase in distortion at extreme temperatures arises for the most part from the imperfect temperature compensation provided by the output-stage biasing network. Additional distortion at −40 degrees C. originates in the detector-AGC circuitry, and is due mainly to the increased impedance of the sintered tantalum electrolytic capacitor, C29, in the AGC filter network. It is the use of this type of capacitor here, notable for good low-temperature performance, that permits operation to −40 degrees C. Substitution of a conventional aluminum-foil type capacitor restricts the useful lower-temperature limit to approximately −10 degrees C.

The change in oscillation frequency with change in supply voltage, shown in Fig. 9, is about 0.2 per cent per volt. The frequency variation of the oscillator with respect to ambient temperature is also shown in Fig. 9.

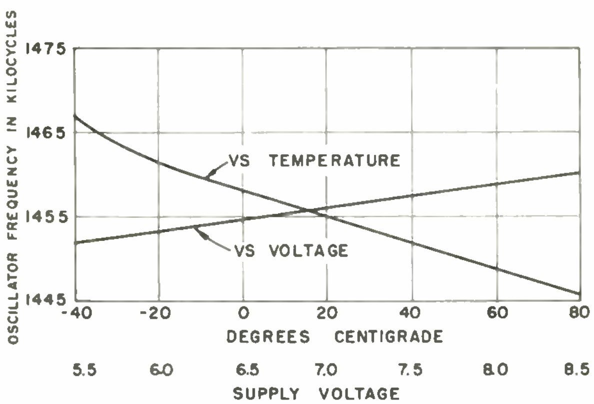


Fig. 9—Oscillator frequency stability.

The difference in oscillator-frequency shift and in the frequency shift of the signal tuned-circuits is in the same direction and of approximately the same magnitude as the shift in the IF tuned-circuits. The net tracking error as a function of ambient temperature is therefore smaller than either the oscillator- or intermediate-frequency shifts.

A Graphical Sinusoidal Analysis of a Nonlinear RC Phase-Shift Feedback Circuit*

DOMENICK BARBIERE†

Summary—In certain nonlinear systems, a form of frequency response has been observed in which the peak is bent over to the right or left depending on the type of nonlinearity. It will be shown that a similar response may be obtained in the RC phase-shift feedback circuit by inserting a nonlinear resistor, such as a thyrite, in the feedback loop. This circuit is discussed with emphasis on a graphical sinusoidal analysis of some general applicability which predicts its frequency response. An informative physical interpretation is also obtained from the analysis of the triple-valued region in the response and of the "jump" phenomena.

The method is essentially a graphical superposition of the voltage-resistance characteristic of the nonlinear resistor upon the voltage-resistance characteristics (with frequency as a parameter) of the circuit between the two points where the nonlinear resistor is intended to be placed. The points of intersection satisfy both sets of characteristics, and therefore yield the response of the circuit. The method is also applied to a series RLC circuit containing a nonlinear iron core inductor, for which current-inductance curves are superimposed.

Introduction

Graphical Technique and Illustrative Circuits

THE ANALYSIS of a nonlinear system generally presents an imposing problem. The familiar linear approaches, such as the Laplace transform or the principle of superposition, are not applicable. With the considerable interest in nonlinear analysis in recent years, various analytical and graphical methods have been developed. These methods have made possible the design of systems combining nonlinear with linear elements for optimum performance.

The graphical approach presented in this paper yields a solution for the first harmonic in a nonlinear circuit. In circuits of limited bandwidth and moderate nonlinearity, this solution is often sufficient. The method is described below as applied to an RC phase-shift feedback circuit containing a nonlinear resistor in the feedback loop and to an RLC circuit with a nonlinear iron core inductor. The latter circuit is well-known and has various applications.[1] Since the nonlinear RC phase-shift circuit was developed and investigated in this laboratory, this circuit and its analysis will be described in more detail.

The RC phase-shift circuit which is shown in Fig. 1 has been in use for some time as an audio oscillator. The oscillations are generated by a single amplifier tube and a phase-shift feedback network consisting of three or more RC sections in cascade. When the tube gain is maintained below the value necessary for oscillations, a frequency-selective response is obtained. In this application, however, the null type of feedback network such as the twin-T has been preferred.

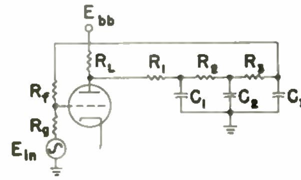


Fig. 1—Basic RC phase-shift feedback circuit.

The introduction of a nonlinear resistor for one of the resistors in the feedback loop results in a bending of the frequency response peak to the right or left depending on the type of nonlinear resistor. In the region of bending, the response curve is triple-valued with three possible values of voltage for one of frequency. This phenomenon, which is peculiar to nonlinear systems, has been observed in the RLC circuit with a nonlinear reactive element,[1] in mechanical systems containing nonlinear springs,[2] and in saturating servomechanisms.[3] One may consider the nonlinear resistor in the feedback loop of the RC phase-shift circuit as equivalent to a nonlinear reactor in the RLC circuit, since the resistor exerts a tuning effect through its influence on the feedback phase shift.

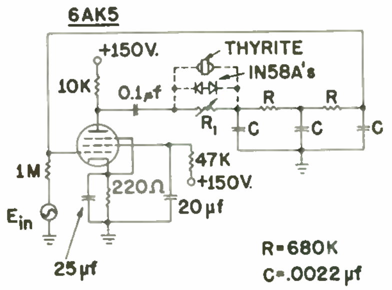


Fig. 2—Circuit leading to curves in Figs. 3, 4, 5 and 6.

The nonlinear version of the RC phase-shift circuit is shown in Fig. 2. The approach that first comes to mind for its analysis is to set up the differential equation for each node or mesh as in linear circuits. For the nonlinear resistor, some approximate volt-ampere characteristic must be assumed, such as several terms of a series expansion. The mathematics in solving these nonlinear equations is considerably involved, and at best, only an approximate analytical solution can be expected.

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[1] C. G. Suits, "New applications of nonlinear circuits to relay and control problems," *Trans. AIEE*, vol. 51, pp. 914–922; 1932.

[2] J. J. Stoker, "Nonlinear Vibrations in Mechanical and Electrical Systems," Interscience Publishers, Inc., New York, N.Y., chap. IV; 1950.

[3] E. Levinson, "Some saturation phenomena in servomechanisms with emphasis on the tachometer stabilized system," *Trans. AIEE*, vol. 72, part II, pp. 1-9; March, 1953.

In the graphical method, a variable linear resistor is first substituted between the two points of the circuit where the nonlinear resistor is intended to be placed, as shown by R_1 in Fig. 2. For some constant value of input voltage to the circuit, curves are plotted of the voltage across the resistor as a function of resistance with frequency as a parameter. These curves shall be termed the voltage-resistance characteristics of the linear circuit, and may be computed by linear methods, if desired. Upon these curves is superimposed the voltage-resistance characteristic of the nonlinear resistor to an ac voltage. The resistance of the nonlinear resistor is defined here as the amplitude of an applied sinusoidal voltage over the amplitude of the fundamental component of current. The points of intersection between the linear circuit characteristics and the nonlinear resistor characteristic satisfy both and, therefore, represent the solutions for the first harmonic in the circuit. For a circuit with a nonlinear iron core inductor, current-inductance characteristics may be superimposed. Similarly, for a circuit with a nonlinear saturating capacitor, voltage-capacitance curves may be superimposed.

The retention of only the fundamental component of current through the nonlinear resistor in defining its resistance, limits the method to circuits with moderate nonlinearity where the harmonics generated by the nonlinear element are either small, or are attenuated by falling outside of the circuit passband. In the nonlinear RC phase-shift circuit the distortion was very small, except for low-frequency inputs to circuit where harmonics then happened to fall in the region of resonance.

Comparison with Similar Methods

Basically, the method above assumes that for the first harmonic in a circuit a nonlinear element behaves as if it were a variable linear impedance, and solutions are obtained by a graphical comparison with an actual variable linear impedance. This procedure was first suggested mathematically by Kryloff and Bogoliuboff,[4] who derived formulas for the equivalent impedance of nonlinear elements. Their analysis and results, which form a significant contribution to nonlinear theory, will be outlined very briefly.

A nonlinear differential equation of the general form

$$m\frac{d^2x}{dt^2} + \epsilon f\left(x, \frac{dx}{dt}\right) + kx = 0 \tag{1}$$

is considered, where the nonlinear terms in x and the damping terms are grouped in ϵf, which is assumed small. For the linear undamped case ($\epsilon = 0$), solution of (1) is

$$x = a\sin(\omega t + \phi) \tag{2}$$

$$\frac{dx}{dt} = a\omega\cos(\omega t + \phi), \tag{3}$$

[4] N. Kryloff and N. Bogoliuboff, "Introduction to Nonlinear Mechanics," Translated from the Russian by S. Lefschetz, Princeton University Press, Princeton, N. J.; 1943.

where $\omega = \sqrt{k/m}$ and where a and ϕ are constants. For the "First Approximation" in a slightly nonlinear system, a solution of the same form as in (2) is tried, where a and ϕ are now allowed to be functions of time but are so selected that (3) is satisfied. A basic pair of equations is derived from which a and ϕ may be calculated and substituted in (2) for "First Approximation" solution.

The surprising result of this solution is, it enables transformation of (1) to the familiar linear equivalent,

$$m\frac{d^2x}{dt^2} + \bar{\lambda}\frac{dx}{dt} + \bar{k}x = 0 \tag{4}$$

to terms of order ϵ^2. The equivalent dissipation factor, $\bar{\lambda}$, and the equivalent spring constant, $\bar{k}$, are found to be given by

$$\bar{\lambda} = \frac{\epsilon}{\pi\omega a}\int_0^{2\pi} f(a\sin\psi, a\omega\cos\psi)\cos\psi d\psi \tag{5}$$

$$\bar{k} = k + \frac{\epsilon}{\pi a}\int_0^{2\pi} f(a\sin\psi, a\omega\cos\psi)\sin\psi d\psi, \tag{6}$$

where $\bar{k}$ and $\bar{\lambda}$ are thus functions of the amplitude a, rather than the instantaneous displacement x.

By replacing the mechanical constants in (1) and (4) with their electrical analogs, the equivalent impedance of various types of nonlinear elements may be derived. Consider a nonlinear resistor, an iron core inductor and a saturable capacitor, with characteristics that may be approximated by the following cubics, respectively,

$$\begin{aligned} e &= Ri(1 + \alpha i^2) \\ \Phi &= Li(1 - \beta i^2) \\ e &= \frac{q}{C}(1 + \gamma q^2), \end{aligned} \tag{7}$$

where Φ, e, i and q are the instantaneous total flux, voltage, current and charge, respectively, and α, β and γ are constants indicating the degree of nonlinearity. By applying the electrical analogs of (5) and (6), the following equivalent constants for the nonlinear elements are obtained:

$$\begin{aligned} R_e &= R\left(1 + \frac{3\alpha}{2}I^2\right) \\ L_e &= L\left(1 - \frac{3}{2}\beta I^2\right) \\ C_e &= C\left(1 - \frac{3\gamma}{2\omega^2}I^2\right), \end{aligned} \tag{8}$$

where I is the rms current. The characteristic of a nonlinear element, originally a function of the instantaneous current or charge, has thus been transformed to a more workable form for ac circuits dependent on the rms current. Considering, as an example, a series RLC circuit containing a linear and a nonlinear inductor, we may write

$$I = \frac{E}{\left\{R^2 + \left[\omega(L + Le) - \frac{1}{\omega c}\right]^2\right\}^{1/2}} \tag{9}$$

$$\omega = \sqrt{\frac{1}{C(L + Le)}}.$$

Kryloff and Bogoliuboff then show that the equivalent constants in (8) may be determined more directly from (7) by assuming i a pure sinusoid and solving for the first harmonic in e (or Φ). When an approximate mathematical expression, such as in (7), is not readily available for the nonlinear element, this approach may be carried out experimentally with a harmonic wave analyzer. The nonlinear resistor characteristic, as defined in the graphical analysis above, is therefore a plot of R_e, drawn versus E_{rms} rather than I_{rms}. It should be noted that C_e in (8) is a function ω as well as the rms current, so that a family of curves with frequency as a parameter is required to represent a nonlinear capacitor.

Mention should also be made of a similar technique which has been applied in an appropriate form to nonlinear feedback control systems.[3,5] The equivalent transfer function of a nonlinear element, such as a relay or a saturating component, is determined for the fundamental component in the system. This transfer function is generally referred to as the "describing function" and is a function of amplitude. The equation for the stability or the closed-loop gain of the system is manipulated so that the describing function appears alone on the left-hand side, and the solutions are then obtained graphically.

Procedure

Nonlinear RC Phase-Shift Circuit

The circuit shown in Fig. 2 was investigated for two types of nonlinear resistors in position R_1: a thyrite (General Electric No. 8399401G1) and a pair of 1N58A germanium diodes connected back-to-back. The circuit characteristics of the voltage across R_1 versus linear resistance were measured by inserting a decade resistance box for R_1. Since this resistor is floating at both ends, it was found expedient to measure the voltage across it by feeding the voltages at its ends to a calibrated difference amplifier. An abritrary even value of input voltage to the circuit was chosen: $E_{in} = 100$ millivolts. Since these curves apply to a linear circuit, appropriate scale manipulations may be applied for other inputs, i.e., halving the vertical scale adapts curves measured with $E_{in} = 100$ mv to those for $E_{in} = 50$ mv. The voltage-resistance characteristics of the circuit in Fig. 2 are shown as the solid line curves in Fig. 3.

The voltage-resistance characteristic of the nonlinear resistor was measured by applying a sinusoidal voltage of some reasonable audio frequency to the resistor and measuring the current component of the same frequency with a harmonic wave analyzer. Since the analyzer is a voltage-measuring device, a known negligibly small resistor was placed in series with the nonlinear resistor and the corresponding voltage developed across this small resistor was measured. The characteristics of the thyrite and of the diode back-to-back arrangement are shown as the broken line curves in Fig. 3. These curves are independent of the circuit.

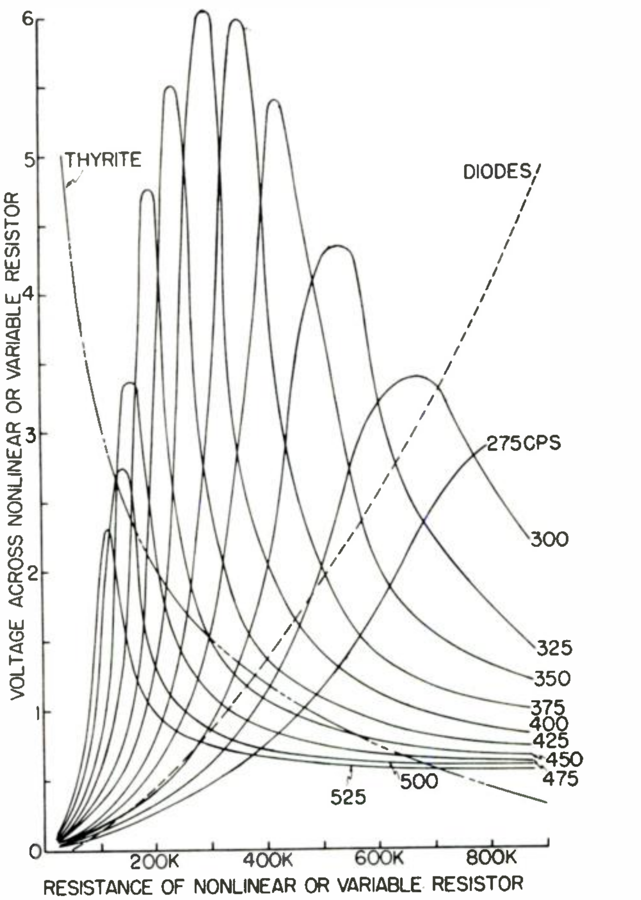

Fig. 3—Superposition of nonlinear resistor characteristic (broken line curves) on circuit characteristics (solid line curves). Circuit is shown in Fig. 2. Thyrite is General Electric No. 8399401G1. Diodes are two 1N58A's back-to-back.

The broken line curves in Fig. 3 show that the thyrite and the diodes represent opposing types of characteristics. The resistance of the thyrite decreases as the applied voltage is increased, while the resistance of the diodes increases with voltage. Referring back to (7) and (8), this is equivalent to α positive for the diodes and negative for the thyrite. Considering the back-to-back arrangement of the diodes more closely, we note that looking into this combination from either direction, the back resistance of one diode is seen with the forward of the other being negligible. Thus, the diodes characteristic shown in Fig. 3 represents the back resistance of a diode, symmetrically arranged. A characteristic similar to that of the thyrite is obtainable by working with the forward resistance of the diodes. This may be accomplished through a parallel arrangement with plate con-

[5] E. C. Johnson, "Sinusoidal analysis of feedback-control systems containing nonlinear elements," *Trans. AIEE*, vol. 71, part II, pp. 169–181; July, 1952.

nected to cathode. Looking into this combination from either direction, the forward resistance of one diode is seen with the back of the other being negligible. However, the low impedance level of this combination resulted in loading difficulties, whence thyrite was used.

The curve marked $E_{in}=100$ mv in Fig. 4 was plotted from the intersections of the thyrite characteristic with the circuit characteristics in Fig. 3. The curve for $E_{in}=50$ mv in Fig. 4 was obtained by the simple manipulation of halving the vertical scale in Fig. 3 for the circuit characteristics and drawing the same thyrite characteristic to this new scale. Fig. 5 shows the corresponding results for the diodes back-to-back arrangement.

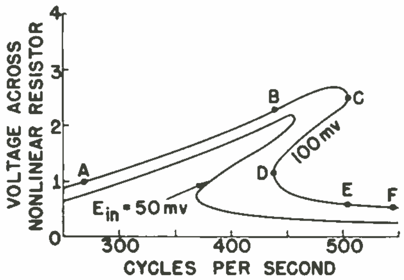

Fig. 4—Response curves from intersections of thyrite characteristic with circuit characteristics in Fig. 3. For $E_{in}=50$ mv, vertical scale in Fig. 3 was halved and thyrite characteristic drawn to new scale.

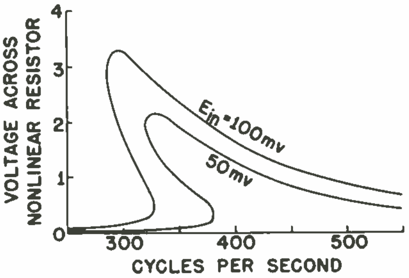

Fig. 5—Response curves from intersections of diodes characteristic with circuit characteristics in Fig. 3 For $E_{in}=50$ mv, vertical scale in Fig. 3 was halved and diodes characteristic drawn to new scale.

With the thyrite, therefore, the response of the circuit is bent over to the right, while with the diodes, the response is bent over to the left. A similar situation is found in mechanics, where nonlinear "hard" and "soft" springs result in responses like those in Figs. 4 and 5, respectively.[2]

In view of the unusual shape of the responses in Figs. 4 and 5, it would be well at this point to consider what is actually observed experimentally as the frequency of the audio oscillator input to the circuit is swept. Consider the curve for $E_{in}=100$ mv in Fig. 4. For increasing frequency, the response rides along A–B until point C is reached. The response then suddenly jumps down to point E and continues on to F. For decreasing frequency, the response rides along the portion F–E until point D is reached. It then suddenly jumps up to point B and continues on to A. The points C and D are generally referred to as the jump points of the circuit. The tangent to the response curve at these points is vertical. The region between points C and D is not observable experimentally, and therefore represents a set of unstable solutions. By drawing a vertical constant frequency line at about 450 cps in Fig. 4, the distinguishing feature of the unstable region can be seen to be that as the input (E_{in}) to the circuit increases, the output actually decreases.

The curves in Fig. 3 show in a striking manner why the triple values and the jumps occur. Following the diodes characteristic, we note that for each frequency parameter from 525 to 350 cycles, one point of intersection occurs between the diodes characteristic and the circuit characteristic. At 325 and 300 cycles, three points of intersection are found. These points represent three possible solutions at a single frequency, one of which has been found to be unstable. At 275 cycles, only one point of intersection is again found and at a low value of voltage. Somewhere between 300 and 275 cycles, the voltage was forced to jump from a high value to a low value. Observing the relationship of the diodes characteristic to the peaks of the circuit curves, it can be seen that for some frequency parameter between 300 and 275 cycles, there is a critical circuit curve for which the diodes curve just glances off the peak. For the circuit curve corresponding to the next frequency decrement, the diodes curve is above the peak, and an intersection can occur only at some low voltage at the lower left-hand corner of Fig. 3. The frequency parameter of the critical circuit curve represents the jump frequency. It was estimated that the jump should occur at about 290 cycles. When the circuit was put together with the diodes in place, the jump occurred at 292 cycles.

The above considerations suggest a quick graphical technique for obtaining the jump frequencies of the circuit directly. If the envelope of the peaks of the circuit curves in Fig. 3 is drawn, then the intersection of the nonlinear resistor characteristic with this envelope represents the approximate jump point. The point is approximate in that the nonlinear resistor characteristic is actually tangent to the side of the peak of the critical circuit curve, rather than to the peak itself. The envelope may be plotted directly by setting the variable linear resistor in Fig. 2 at some value and varying the frequency of the input audio oscillator until a maximum voltage across the resistor is noted. The value of this voltage and the frequency are recorded. The curve in Fig. 6 containing the frequency parameter was plotted in this manner for $E_{in}=100$ mv. Rather than re-plot this curve for $E_{in}=50$ mv and $E_{in}=200$ mv, the simpler diodes characteristic was manipulated. The remaining three curves in Fig. 6 represent the same diodes characteristic as in Fig. 3, except that the curve for $E_{in}=50$ mv was drawn with the vertical scale in Fig. 6 halved, and the curve for $E_{in}=200$ mv was drawn with the vertical scale doubled. Also shown in Fig. 6 is a comparison of jump points as estimated from the figure with those actually measured when circuit was put together.

Most of the labor in setting up the graphical construction in Fig. 3 is concerned with plotting the circuit characteristics. These curves represent the linear circuit, with which the engineer is more familiar. Once the circuit characteristics are drawn, the effects of various types of nonlinear resistors may be investigated.

Some generality may be ascribed to the circuit characteristics plotted in Fig. 3. The impedance level of the feedback loop in the circuit in Fig. 2 may be altered and the curves in Fig. 3 still apply as long as the two fixed resistors (R) are maintained equal to each other, and the three condensers (C) are maintained equal to each other. Changing the value of the three condensers results only in a frequency shift. The generalization may be accomplished by using the normalized variable R_1/R as the abscissa in Fig. 3 and ωRC as the frequency parameter. These considerations permit the fitting of the impedance level of the feedback network to that of the nonlinear resistor without necessitating continual replotting of the circuit curves.

JUMP FREQUENCIES

E_{in}	PREDICTED	MEASURED
50mv	325cps	325cps
100	293	292
200	266	270

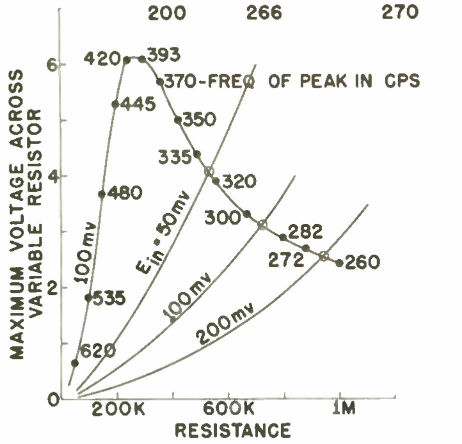

Fig. 6—Curve containing frequency parameter is the envelope of the peaks of the circuit characteristics in Fig. 3. Remaining three curves represent the same diodes characteristic, except that the curve for $E_{in}=50$ mv is drawn to one-half the vertical scale of the figure and that for $E_{in}=200$ mv to twice the vertical scale.

The RC phase-shift circuit shown in Figs. 1 and 2 is generally referred to as the shunt-C type, since the resistors are floating and the condensers are placed to ground. By interchanging the condenser and resistor positions, the shunt-R form is obtained. Results similar to those discussed above have been obtained with the shunt-R form.

Series RLC Circuit with an Iron Core Inductor

The graphical construction for this circuit is shown in Fig. 7. The solid line circuit characteristics of current versus inductance were computed from the formula

$$I = \frac{E}{\left\{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2\right\}^{1/2}} \tag{9}$$

for $E=100$ volts, $R=100$ ohms and $C=2$ mfd, where I, E represent rms quantities. A reasonable characteristic for the nonlinear inductor was assumed of the form $(1-3/2\beta I^2)$ henries as in (8), and β was set at 0.5. The inductance is therefore one henry for I small and decreases to 0.25 henries for $I=1$ ampere. The iron core inductor characteristic is shown as the broken line curve in Fig. 7. From the points of intersection of the inductor

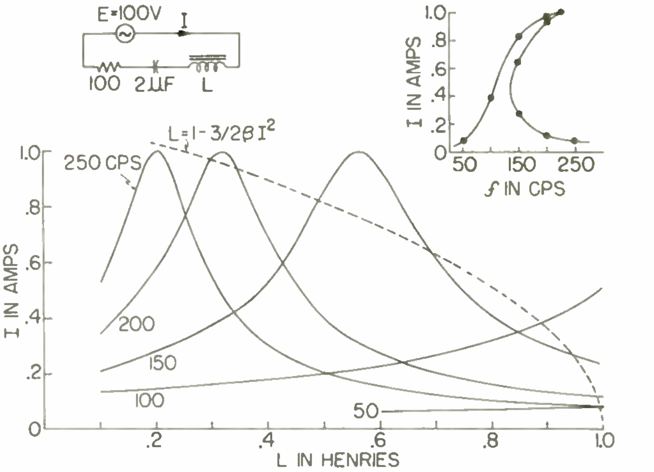

Fig. 7—Superposition of iron core inductor characteristic (broken line curve) on series RLC circuit current versus inductance characteristics (solid line curves). Response in upper right-hand corner was plotted from the points of intersection.

characteristic with the circuit characteristics, the response curve in the upper right-hand corner of Fig. 7 was obtained. As in the RC phase-shift circuit, the intersection of the inductor characteristic with the envelope of the peaks of the circuit curves yields the approximate jump point. In this case, the envelope is a straight horizontal line equal to $I=1$ ampere. The inductor characteristic intersects this line at $L=0.25$ henries, whence the jump frequency is established at

$$f = \frac{1}{2\pi\sqrt{LC}} = 225 \text{ cycles} \tag{10}$$

for $L=0.25$ henries and $C=2$ mfd.

The triple-valued I versus E curve for the circuit may be obtained directly as shown in Fig. 8. The solid line

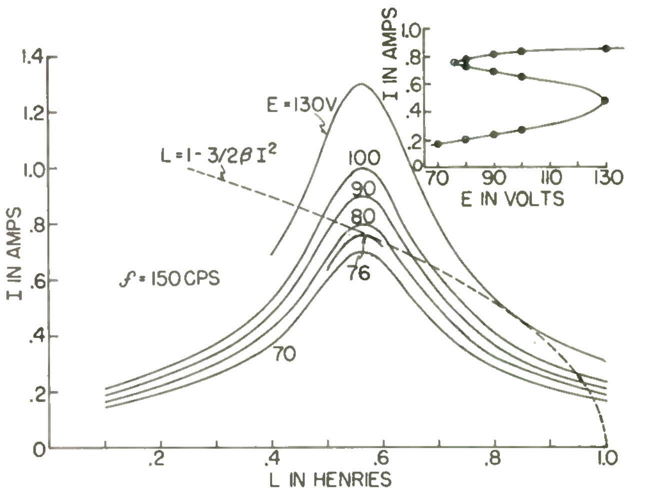

Fig. 8—Superposition of iron core inductor characteristic (broken line curve) on series RLC circuit current-inductance characteristics with applied voltage as a parameter (solid line curves). The curve in the upper right-hand corner was plotted from the points of intersection.

curves in Fig. 8 are the same as the 150 cycle curve in Fig. 7, except multiplied by an appropriate factor for different values of applied voltage. The intersections of these curves with the inductor characteristic yield the curve in the upper right-hand corner of Fig. 8. Large jump changes of current occur at the vertical tangencies of this curve with small changes of voltage. This phenomenon has been applied to the design of a voltage regulator.[1]

Conclusion

The method in this paper may be applied to circuits containing a nonlinear element where the first harmonic is predominant. For more than one nonlinear element in a circuit, the method becomes cumbersome. The experimental approach shows that the analysis may be carried out with a minimum of computation. In particular, the characteristic of the nonlinear element need not be known in mathematical form. The method is inapplicable to systems based on extreme nonlinearities, such as the rectifier where the dc component is actually used, and the magnetic amplifier.

Acknowledgment

The author wishes to express his appreciation to Dr. John S. Thomsen for his invaluable advice on some of the fine points of nonlinear theory and to Clayton W. Smith for his considerable assistance in preparing this report. In particular, credit is due to William M. Waters who constructed the first working model of the RC circuit discussed in this paper, but due to circumstances was unable to continue the work.

Backward-Wave Oscillators*

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Summary—**The Pierce traveling-wave tube theory is modified to apply to the backward-wave oscillator. Theoretical dependence of both starting current and frequency upon space charge and circuit loss is calculated, as is the dependence of efficiency upon these parameters. Good experimental confirmation is obtained with two tubes, one of which was of adjustable length. Pushing, pulling, stability, frequency gaps, and spurious oscillations are described and explained. Lack of oscillation at low-voltage end of tuning range of some tubes is tentatively assigned to a velocity-distribution effect.**

Introduction

THE BACKWARD-WAVE oscillator has several useful properties; it tunes fully electronically and rapidly over a very wide band of frequencies; it has an efficiency comparable to that of a reflex klystron; its frequency is rather independent of load impedance; it can be designed to be unusually stable in frequency.

This paper contains a brief review of some of the literature bearing on the oscillator (called O-type Carcinotron by the French) and a brief statement of the physical picture of backward-wave oscillator operation, followed by a derivation of the starting current, starting frequency, and efficiency of the device. Design curves based on this theory are presented, and some experimental results are compared with the theory.

Past Works

Several investigations of backward-wave interaction appear in the literature. The work of Pierce[1] contains an arithmetical mistake in calculation of a graph which renders the conclusions invalid, and is, in any case, restricted to the case of zero space charge and zero distributed circuit loss. Kompfner and Williams,[2,3] Epsztein,[4] and Warnecke, Guenard and Doehler[5] clearly explain backward-wave oscillation. The former assumes both loss and space charge equal to zero. Heffner[6] computes starting current as a function of space charge, but does not consider loss and space charge simultaneously. In addition, he uses Bernier's[7] assumption that current, rather than the derivative of current, is the driving

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[1] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Co., New York, N. Y., chap. II; 1950.
[2] R. Kompfner, and N. T. Williams, IRE Conference on Electron Devices, Ottawa, Ont., Can.; June, 1952.
[3] R. Kompfner and N. T. Williams, "Backward-wave tubes," Proc. IRE, vol. 41, pp. 1602–1611; November, 1953.
[4] B. Epsztein, IRE Conference on Electron Devices, Ottawa, Ont., Can.; June, 1952.
[5] R. Warnecke, P. Guénard et O. Doehler, "Phénomènes fondamentaux dans les tubes à onde progressive," *l'Onde Électrique*, no. 325; April, 1954.
[6] H. Heffner, "Analysis of the Backward-Wave Traveling-Wave Tube," Elec. Res. Lab. Rep. No. 48, Stanford Univ., Stanford, Calif; June 18, 1952, and Proc. IRE, vol. 42, pp. 930–937; June, 1954.
[7] J. Bernier, "Essai de théorie du tube électronique à propagation d'ondes," *Ann. Radioélect.*, vol. 2, pp. 87–101; January, 1947.

term in the transmission-line equation. This leads to a small discrepancy between Heffner's work and the present calculations; it may become large for large C and is consequently worth mentioning. The work of Walker[8] agrees with the present work despite his unusual notation, but he does not show calculations for the cases of loss, and simultaneous loss and space charge. Muller's[9] paper extends the analysis to crossed-fields devices.

Watkins and Grow[10] have applied Pierce's[11] current-limiting assumption to calculating the efficiency of the backward-wave oscillator, but they do not consider loss and space charge simultaneously either, nor the transition region between large and small space charge.

Publication of experimental data has been limited.[3,10,12,13] Our frequency pushing differs in sign from that observed by Kompfner and Williams.[3] There is no published work designed to check the theory of backward-wave oscillator starting current and frequency, nor the effect of tube length and operating voltage upon efficiency.

Physical Picture

The structure of the backward-wave oscillator is elegantly simple. It is similar to the conventional traveling-wave amplifier in that it consists of an electron beam and transmission line which interact for many cycles of the oscillation frequency, as shown schematically in Fig. 1. The main difference is that the circuit is capable of propagating a wave whose phase and group velocities are in opposite directions—a "backward wave." Such a wave can be propagated by the lumped circuit, by any periodic structure such as folded line or helix, by certain smooth structures such as the helical sheath, and presumably by other structures.

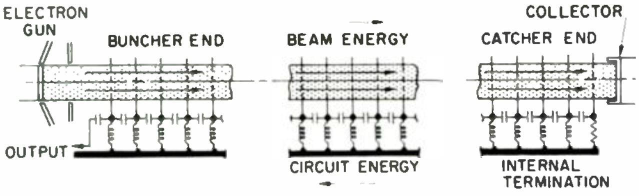

Fig. 1—Schematic drawing of backward-wave oscillator.

The principle of operation of this oscillator, as of most others, can best be understood by considering the buildup of oscillations. Suppose a small pulse of noise near a certain frequency f develops on the circuit at the end marked "internal termination" on Fig. 1; some of this energy will propagate along the circuit toward the end marked "output." This wave will have a phase velocity in the opposite direction, that is, from electron-gun end to collector end of the tube. If a small electron current at a velocity just greater than this phase velocity is introduced, average kinetic energy of electrons will be converted into wave energy, increasing energy of circuit wave and resulting in more energy appearing at "output" than before electron beam was turned on.

So far, this action is quite similar to the action of an ordinary traveling-wave amplifier, except that wave energy and electron-beam energy are traveling in opposite directions; however, this "backward-wave amplifier" is inherently regenerative even when the circuit is matched perfectly at both ends. The circuit fields at "output" produce velocity modulation of the incoming electron beam—the "buncher" end of the oscillator; this velocity modulation becomes current modulation toward the collector or "catcher" end of the tube. The current modulation induces energy in the circuit in the form of a wave whose phase velocity is nearly the same as the electron velocity or a wave whose energy is traveling back toward the gun or output end of the tube. This returned energy is not caused by reflection; if the output load is matched, there is no circuit energy at all traveling from "output" to "internal termination." The returned energy produces further bunching, which in turn produces more returned energy. For low beam currents, regenerative amplification occurs; for higher beam currents, the tube will oscillate.

The amplification frequency, and consequently the oscillation frequency, is restricted to that value for which the circuit velocity is nearly equal to the beam velocity. If the circuit wave is dispersive, as all backward waves must be, the oscillation frequency will be tunable over a wide range of frequencies through variation of the beam voltage. Some feedback-type traveling-wave tube oscillators used in the past have had this property also, except for the objectionable characteristic of mode jumping. To stay in the same mode, an oscillator must have a fixed integral number of wavelengths around any feedback loop, a condition difficult to satisfy in previous traveling-wave tube oscillators, but automatically satisfied in the backward-wave oscillator. This tube may be imagined to contain many feedback loops, the net length of each being zero wavelengths. The condition that it remained zero is fortunately the condition for gain, namely that the average velocity of the electron beam be approximately the same as the phase velocity of the circuit. Thus, there is a mechanism for automatic mode selection.

Starting Current and Frequency

We now proceed to calculate the starting current and frequency for the backward-wave oscillator. Pierce[14] discusses excitation of fields on a circuit by unidirectional

[8] L. R. Walker, "Starting currents in the backward-wave oscillator," *Jour. Appl. Phys.*, vol. 24, pp. 854–859; July, 1953.

[9] M. Muller, "Traveling-wave tube amplifiers and backward-wave oscillators," Proc. IRE, vol. 42, pp. 1651–1658; November, 1954.

[10] D. A. Watkins and R. Grow, "Backward-wave oscillator efficiency," IRE Conference on Electron Devices, Stanford, Calif.; June, 1953; and private communication.

[11] J. R. Pierce, *loc. cit.*, chap. 12.

[12] D. A. Watkins, and A. E. Siegman, "Helix impedance measurements using an electron beam," *Jour. Appl. Phys.*, vol. 24, pp. 917–922; July, 1953.

[13] J. W. Sullivan, "A wide-band voltage-tunable oscillator," Proc. IRE, vol. 42, pp. 1658–1665; November, 1954.

[14] J. R. Pierce, *loc. cit.*, chap. 6.

current in a beam passing nearby. The current is assumed to be $\mathrm{Re}[i \exp(j\omega t - \Gamma z)]$. The power flow P in a backward wave is negative, or opposite in direction to the phase velocity; thus, to keep $K>0$ we must rewrite Pierce's circuit equation (7.1) using the new definition

$$K = -E_z^2/2\beta^2 P. \qquad (1)$$

$$\frac{E}{\Gamma} = \left[-\frac{\Gamma_1 K}{\Gamma_1^2 - \Gamma^2} - \frac{j2QK}{\beta_e} \right] \Gamma i \qquad (2)$$

expresses the electric field E produced by a stream current i. In (2) the first term is Pierce's circuit voltage, and the second term is the voltage due to local space charge. Pierce also shows that a field $\mathrm{Re}[E \exp(j\omega t - \Gamma z)]$ will produce a current i according to the ballistic equation

$$i = \frac{jI_0\beta_e E}{2V_0(j\beta_e - \Gamma)^2}, \qquad (3)$$

and a velocity

$$v = -\frac{(j\beta_e - \Gamma)u_0 i}{j\beta_e I_0}; \qquad (4)$$

here $i/I_0 \ll 1$ and $v/u_0 \ll 1$ have been assumed (small-signal approximation) to render the ballistic equation linear. Eqs. (2) and (3) can be consistent only if the determinantal equation

$$(j\beta_e - \Gamma)^2(\Gamma_1^2 - \Gamma^2) = -j2\beta_e\Gamma^2\Gamma_1 C^3 + 4\Gamma^2 C^3 Q(\Gamma_1^2 - \Gamma^2) \qquad (5)$$

is satisfied, where

$$C^3 \equiv KI_0/4V_0. \qquad (6)$$

To place this in a form so that the simplifying approximation $C \ll 1$ can be applied, define b and d with

$$-\Gamma_1 \equiv -j\beta_e(1 + Cb) + \beta_e Cd, \qquad (7)$$

and δ with

$$-\Gamma \equiv -j\beta_e + \beta_e C\delta. \qquad (8)$$

Let x and y be defined by

$$\delta \equiv x + jy. \qquad (9)$$

Note that (7) differs in the sign of d from the definition used by Pierce for forward waves. Our definition is chosen so that $d>0$ corresponds to circuit attenuation for a backward wave. This will yield, for $C \ll 1$, the small-C determinantal equation

$$-\delta^2 = \frac{1}{-b - jd + j\delta} + 4QC. \qquad (10)$$

Here the wave with phase velocity opposite to the electron stream has been dropped, since in the approximation $C \ll 1$ it has the same velocity and rate of attenuation whether or not the electron stream is present.

The three roots of (10), namely, δ_1, δ_2, and δ_3, determine propagation constants of three waves. Associated with each wave is a circuit-voltage wave, beam-current wave, beam-velocity wave, and total-voltage wave. The left-hand side of (2) is called the total voltage of a wave, and the first term of the right-hand side is called the circuit voltage. V_1 is the total voltage of the wave associated with δ_1, V_2, with δ_2, etc. The sum of the three total voltages is designated

$$V \equiv V_1 + V_2 + V_3. \qquad (11)$$

For computing the power delivered to the load when the latter terminates the helix, only fields associated with the circuit voltage should be considered. V_{c1} is the circuit voltage of the wave associated with δ_1, V_{c2} with δ_2, etc. The above verbal definition of circuit voltage is equivalent to

$$\frac{V_{c1}}{V_1} \equiv \frac{\delta_1^2 + 4QC}{\delta_1^2}, \qquad (12)$$

and cyclic permutations. The voltage from which the actual circuit power should be computed is

$$V_c \equiv V_{c1} + V_{c2} + V_{c3}. \qquad (13)$$

The stream current is, from (3) and (8),

$$\left(-\frac{2V_0C^2}{I_0}\right) i = \frac{V_1}{\delta_1^2} + \frac{V_2}{\delta_2^2} + \frac{V_3}{\delta_3^2}; \qquad (14)$$

and the stream velocity is, from (4) and (8),

$$\left(\frac{ju_0C}{\eta}\right) v = \frac{V_1}{\delta_1} + \frac{V_2}{\delta_2} + \frac{V_3}{\delta_3}. \qquad (15)$$

Inverting (11), (14), and (15) we find

$$V_1 = \frac{V - (\delta_2 + \delta_3)(ju_0C/\eta)v + \delta_2\delta_3(-2V_0C^2/I_0)i}{(1 - \delta_2/\delta_1)(1 - \delta_3/\delta_1)}, \qquad (16)$$

and cyclic permutations. (It is hoped that use of η for charge-to-mass ratio of the electron will not conflict with later use of η for efficiency.) Eq. (16) is good for all values of z; we now apply it for $z=0$ where the entering stream is unmodulated, or

$$i = v = 0 \quad \text{at} \quad z = 0. \qquad (17)$$

Suppose some power is applied to the collector end of the circuit, traveling toward the gun end. We take the applied circuit voltage at the collector end of the tube ($z=l$) to be $V_c(l)$, and the resulting circuit voltage at $z=0$ to be $V_c(0)$. The voltage gain of this backward-wave amplifier is $|V_c(0)/V_c(l)|$. Using (11) through (17) we may write

$$\exp(j2\pi N)\frac{V_c(l)}{V_c(0)} = \frac{(\delta_1^2 + 4QC)\exp(2\pi CN\delta_1)}{(\delta_1 - \delta_2)(\delta_1 - \delta_3)} + \frac{(\delta_2^2 + 4QC)\exp(2\pi CN\delta_2)}{(\delta_2 - \delta_3)(\delta_2 - \delta_1)} + \frac{(\delta_3^2 + 4QC)\exp(2\pi CN\delta_3)}{(\delta_3 - \delta_1)(\delta_3 - \delta_2)}. \qquad (18)$$

N is the length of the beam in electronic wavelengths, or

$$\beta_e l = 2\pi N. \tag{19}$$

Let us take the distributed loss of the circuit as

$$L \equiv \frac{40\pi}{\ln 10} dCN = 54.575\ dCN \text{ decibels.} \tag{20}$$

Given L and QC as parameters, (10) implicitly gives δ_1, δ_2, and δ_3 as functions of CN and b. Imagining these values for δ_1, δ_2, and δ_3 substituted into (18), we have the gain of the backward-wave amplifier in terms of CN and b with QC and L as parameters. For each pair of values of the parameters, the gain proves to have an infinite value for certain critical pairs of values of CN and b. Consider a tube of certain length, with certain beam voltage V_0; as beam current I_0 is raised, a certain current I_s will be reached which makes CN exceed the lowest of the critical values at the frequency appropriate for the necessary value of b. At this point, the tube will break into backward-wave oscillation. Accordingly, the conditions for backward-wave oscillation are the roots CN and b of the complex equation

$$\frac{V_c(l)}{V_c(0)} = 0, \tag{21}$$

in which QC and L are parameters. We have solved this equation numerically for the lowest mode of oscillation (root of smallest CN) for a grid of values of the parameters. The results are presented in tabular form in Table I (right) and also in Figs. 2 and 3, following page. Here $H \equiv 2\pi N\omega_q/\omega, \beta \equiv -j\Gamma_1, (\beta-\beta_e)l = 2\pi CNb$, and $\omega_q/\omega = \sqrt{4QC^3}$; see Appendix for uses of some of these quantities. These results differ slightly from those of Heffner;[6] they agree with those of Walker[8] for the only cases he computes, namely various values of QC for $L=0$.

No matter whether QC, H, or Q/N is used as a space-charge parameter, this start-oscillation condition divides the CN vs space-charge plane into two regions; in one, oscillation can occur. As current I_0 is increased in a tube, a locus of operating points will be described in the CN versus QC plane. The locus will be a straight line of unity slope, passing through the origin, since both CN and QC for the tube are proportional to $I_0^{1/3}$. Where this locus intersects the CN vs QC characteristic of Table I, oscillation will begin. When Q/N is used as the space-charge parameter, the locus of operating points is a vertical line, since Q/N is independent of I_0. This property renders Q/N a useful parameter in design. It is interesting to note that the starting characteristic in this latter plane is a monotonically increasing function, and there exists a CN for every Q/N. This is in contrast to certain experimental results obtained with most backward-wave oscillators at low voltages; consequently, we look for another effect which may lead to an operating condition in which backward-wave oscillation is impossible for any current when beam voltage is low enough. This is described later under velocity distribution effects.

The preceding information easily can be used to compute the starting current I_s of a backward-wave oscillator of beam voltage V_0 whose electron transit time is N cycles. K, for a helix, can be determined from the paper of Watkins and Ash[15] or from the approximate formulas

$$K = \frac{16.7ka}{(1-0.923ka)^2}\frac{I_1^2(\gamma r)}{I_1^2(\gamma a)} \text{ (unifilar helix)} \tag{22}$$

TABLE I

THEORETICAL STARTING CONDITIONS FOR BACKWARD-WAVE OSCILLATOR

	L	b	CN	H	$(\beta-\beta_e)l$	QC/CN
$QC=0$	0	1.522	0.3141	0	3.003	0
	2	1.488	0.3275	0	3.062	0
	4	1.457	0.3414	0	3.125	0
	6	1.427	0.3556	0	3.188	0
	10	1.375	0.3847	0	3.324	0
	15	1.318	0.4229	0	3.502	0
	20	1.271	0.4627	0	3.695	0
	25	1.231	0.5040	0	3.898	0
	30	1.197	0.5470	0	4.114	0
$QC=0.25$	0	1.501	0.3434	2.158	3.238	0.7280
	2	1.472	0.3603	2.264	3.333	0.6939
	4	1.445	0.3774	2.371	3.426	0.6624
	6	1.421	0.3955	2.485	3.531	0.6321
	10	1.380	0.4329	2.720	3.754	0.5775
	15	1.341	0.4826	3.032	4.066	0.5180
	20	1.314	0.5348	3.360	4.415	0.4675
	25	1.294	0.5893	3.703	4.791	0.4242
	30	1.282	0.6451	4.053	5.196	0.3875
$QC=0.50$	0	1.533	0.3990	3.545	3.843	1.253
	2	1.526	0.4237	3.765	4.063	1.180
	4	1.526	0.4483	3.984	4.298	1.115
	6	1.530	0.4731	4.204	4.548	1.057
	10	1.543	0.5207	4.627	5.048	0.9602
	15	1.553	0.5775	5.132	5.635	0.8658
	20	1.555	0.6344	5.637	6.198	0.7881
	25	1.551	0.6934	6.161	6.758	0.7211
	30	1.547	0.7554	6.712	7.343	0.6619
$QC=0.75$	0	1.834	0.4660	5.071	5.370	1.609
	2	1.838	0.4857	5.286	5.609	1.544
	4	1.839	0.5062	5.509	5.849	1.482
	6	1.833	0.5273	5.739	6.073	1.422
	10	1.825	0.5730	6.236	6.571	1.309
	15	1.818	0.6358	6.919	7.263	1.180
	20	1.817	0.7023	7.643	8.018	1.068
	25	1.818	0.7686	8.365	8.780	0.9758
	30	1.821	0.8359	9.097	9.564	0.8972
$QC=1.00$	0	2.072	0.4914	6.1745	6.397	2.035
	2	2.064	0.5156	6.479	6.687	1.939
	4	2.059	0.5413	6.802	7.003	1.847
	6	2.058	0.5684	7.143	7.350	1.759
	10	2.063	0.6218	7.814	8.060	1.608
	15	2.067	0.6866	8.628	8.917	1.456
	20	2.065	0.7537	9.471	9.779	1.327
	25	2.063	0.8249	10.366	10.69	1.212
	30	2.064	0.8982	11.287	11.65	1.113
$QC=1.50$	0	2.499	0.5522	8.498	8.670	2.629
	2	2.498	0.5764	8.871	9.047	2.602
	4	2.495	0.6022	9.268	9.441	2.491
	6	2.491	0.6293	9.685	9.850	2.384
	10	2.490	0.6881	10.59	10.77	2.180
	15	2.493	0.7605	11.70	11.91	1.972
	20	2.492	0.8350	12.85	13.07	1.796
	25	2.492	0.9138	14.06	14.31	1.641
	30	2.492	0.9935	15.29	15.56	1.510

[15] D. A. Watkins and E. A. Ash, "The helex as a backward-wave circuit structure," *Jour. Appl. Phys.*, vol. 25, pp. 782–790; June, 1954.

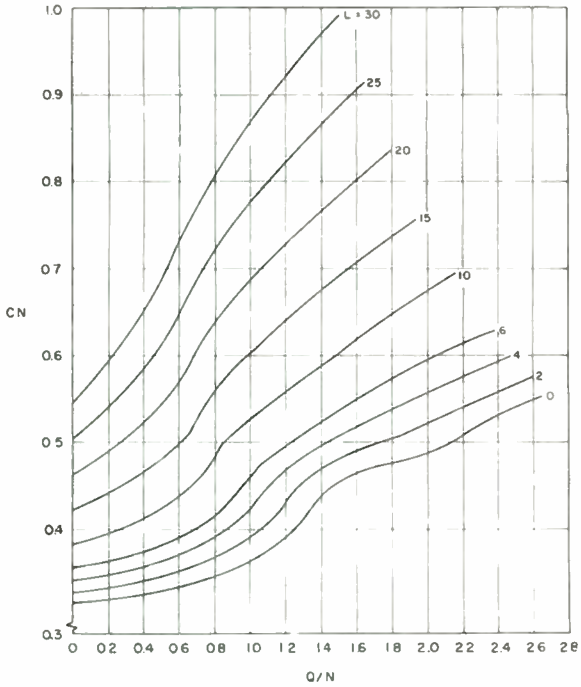

Fig. 2—Theoretical curve of CN vs Q/N at start of backward-wave oscillation.

$$K = \frac{18.4}{1 - 1.224ka} \frac{I_1^2(\gamma r)}{I_1^2(\gamma a)} \text{ (bifilar helix).} \tag{23}$$

These are good approximations to the results of the paper mentioned above for $0.2 < ka < 0.6$. We have assumed the tape width equal to the gap width, in addition to the assumptions made in that paper. Q can be determined from the Appendix of this paper. Q/N is then calculated and CN obtained from Fig. 2. C and I_0 can be obtained using (6); then Fig. 3 determines the frequency of oscillation.

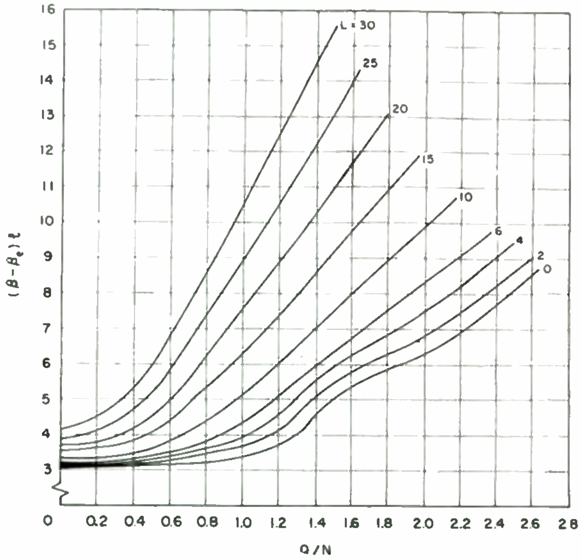

Fig. 3—Theoretical curve of $(\beta-\beta_e)l$ vs Q/N at start of backward-wave oscillation.

The computations for CN and $(\beta-\beta_e)l$ necessitated calculation of the roots δ_1, δ_2, and δ_3 for each value of QC and L. These values are important for extensions of the theory; in view of this possible use, and the considerable arithmetic labor involved in the computation, they are presented for reference as Table II (opposite).

It is useful to note that asymptotic expressions[6] for the starting conditions can be developed in the regime of large QC ($QC > 0.25$). In this regime, the second term of (18) is negligible compared to the first and third, as can easily be verified numerically using the information in Tables I and II, as shown in Fig. 4. If circuit loss is assumed to be zero,

$$\delta_2 \cong j\sqrt{4QC} \tag{24}$$

so this wave is essentially the unperturbed fast space-charge wave of Hahn and Ramo.[16] The magnitudes of the three terms of (18) at the start of oscillation are plotted in Fig. 4. For large space charge, the second term is obviously negligible; i.e., the interaction is between the slow space-charge wave and the circuit wave.

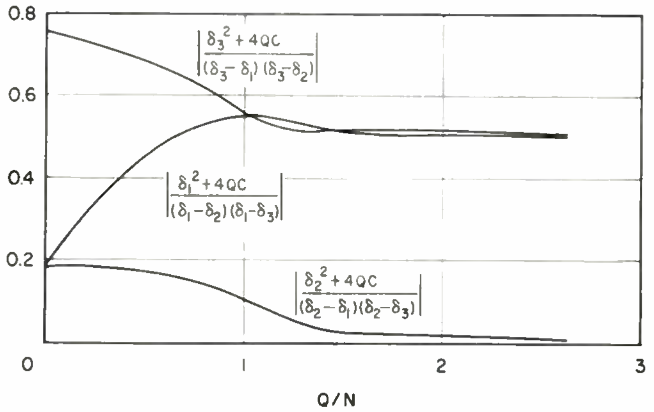

Fig. 4—Relative magnitudes of the three partial waves in the backward-wave oscillator at start of oscillation, vs space-charge parameter Q/N. $L=0$.

Furthermore, δ_1 and δ_3 are pure imaginaries so the magnitudes of the coefficients of the exponentials of the first and third terms of (18) must be equal in order that (21) can be satisfied. This can occur if

$$b \cong \sqrt{4QC}, \tag{25}$$

or slow space-charge wave must be in exact synchronism with circuit wave. Thus, using (24) and (10),

$$\delta_1 = -j\sqrt{4QC}\,[1 - 1/2(QC)^{3/4}] \tag{26}$$

$$\delta_3 = -j\sqrt{4QC}\,[(1 + 1/2(QC)^{3/4}]. \tag{27}$$

In order for (21) to be satisfied, the arguments of the exponential in the first term of (18) must differ from that of the last term by $j(2n+1)\pi$, so

$$CN = \frac{(2n+1)(QC)^{1/4}}{2}, \text{ or} \tag{28}$$

$$CN = [(2n+1)/2]^{4/3}[Q/N]^{1/3}. \tag{29}$$

This expression agrees with Fig. 2 for $L=0$, $Q/N > 1.4$.

[16] S. Ramo, "Space charge and field waves in an electron beam," *Phys. Rev.*, vol. 56, pp. 276–283; August 1, 1939.

TABLE II

THEORETICAL PROPAGATION CONSTANTS FOR WAVES IN A BACKWARD-WAVE OSCILLATOR AT START OF OSCILLATION

	L	δ_1		Fast space-charge wave δ_2		δ_3	
$QC=0$	0	0.72521	+.15046j	−.72521	+.15046j	0	−1.82293j
	2	0.74390	+.13938j	−.71527	+.16864j	0.083275	−1.79602j
	4	0.76168	+.12775j	−.70537	+.18423j	0.15838	−1.76898j
	6	0.77885	+.11591j	−.69589	+.19784j	0.22624	−1.74075j
	10	0.81036	+.09118j	−.67805	+.21964j	0.34389	−1.68582j
	15	0.84560	+.05976j	−.65872	+.23995j	0.46303	−1.61770j
	20	0.87594	+.02827j	−.64244	+.25459j	0.55850	−1.55386j
	25	0.90220	−.00246j	−.62890	+.26552j	0.63562	−1.49407j
	30	0.92476	−.03180	−.61775	+.27382j	0.69799	−1.43902j
$QC=0.25$	0	0	−.35593j	0	+.74479j	0	−1.88986j
	2	0.05419	−.34376j	−.01618	+.74146j	0.06369	−1.86970j
	4	0.10593	−.33755j	−.03159	+.74044j	0.11985	−1.84789j
	6	0.15427	−.33672j	−.04565	+.74147j	0.16938	−1.82575j
	10	0.24007	−.34616j	−.06908	+.74741j	0.25231	−1.78125j
	15	0.32703	−.37039j	−.08987	+.75792j	0.33234	−1.72853j
	20	0.39468	−.40127j	−.10346	+.76862j	0.39408	−1.68135j
	25	0.44779	−.43302j	−.11248	+.77793j	0.44209	−1.63891j
	30	0.48865	−.46539j	−.11831	+.78592j	0.48166	−1.60254j
$QC=0.50$	0	0	−.79876j	0	+1.28250j	0	−2.01674j
	2	0.04676	−.79680j	−.004494	+1.28231j	0.04423	−2.01151j
	4	0.08835	−.80145j	−.008465	+1.28272j	0.08361	−2.00727j
	6	0.12507	−.81025j	−.011934	+1.28350j	0.11926	−2.00325j
	10	0.18685	−.83442j	−.017663	+1.28558j	0.18272	−1.99415j
	15	0.24963	−.86576j	−.023288	+1.28806j	0.24956	−1.97530j
	20	0.30151	−.89517j	−.027729	+1.29016j	0.30392	−1.95000j
	25	0.34522	−.92156j	−.031282	+1.29186j	0.34666	−1.92130j
	30	0.38099	−.94662j	−.034054	+1.29334j	0.38076	−1.89371j
$QC=0.75$	0	0	−1.19645j	0	+1.64704j	0	−2.28459j
	2	0.03810	−1.19988j	−.001933	+1.64718j	0.039287	−2.28531j
	4	0.07316	−1.20315j	−.003706	+1.64728j	0.075342	−2.28113j
	6	0.10547	−1.20775j	−.005329	+1.64741j	0.10826	−2.27465j
	10	0.16333	−1.21765j	−.008181	+1.64762j	0.16465	−2.25497j
	15	0.22218	−1.23635j	−.011015	+1.64811j	0.22113	−2.22975j
	20	0.26825	−1.25934j	−.013193	+1.64874j	0.22674	−2.20640j
	25	0.30545	−1.28430j	−.014931	+1.64942j	0.30548	−2.18412j
	30	0.33544	−1.30921j	−.016322	+1.65006j	0.33838	−2.16285j
$QC=1.0$	0	0	−1.50029j	0	+1.93663j	0	−2.50834j
	2	0.036174	−1.49754j	−.001165	+1.93652j	0.036061	−2.50298j
	4	0.069266	−1.49845j	−.002223	+1.93650j	0.068358	−2.49705j
	6	0.099034	−1.50294j	−.003173	+1.93656j	0.097539	−2.49163j
	10	0.15005	−1.51868j	−.004804	+1.93685j	0.14936	−2.48117j
	15	0.20269	−1.54151j	−.006478	+1.93725j	0.20409	−2.46373j
	20	0.24689	−1.56200j	−.007839	+1.93752j	0.24715	−2.44052j
	25	0.28204	−1.58323j	−.008909	+1.93780j	0.28187	−2.41857j
	30	0.31087	−1.60445j	−.009779	+1.93807j	0.31091	−2.39762j
$QC=1.5$	0	0	−1.99950j	0	+2.40753j	0	−2.90703j
	2	0.031861	−2.00014j	−.0005534	+2.40753j	0.032273	−2.90539j
	4	0.061188	−2.00168j	−.001060	+2.40752j	0.061572	−2.90084j
	6	0.088230	−2.00415j	−.001523	+2.40751j	0.087993	−2.89437j
	10	0.13464	−2.01538j	−.002319	+2.40758j	0.13398	−2.88220j
	15	0.18205	−2.03483j	−.003135	+2.40771j	0.18249	−2.86588j
	20	0.22132	−2.05383j	−.003798	+2.40782j	0.22138	−2.84599j
	25	0.25274	−2.07319j	−.004327	+2.40792j	0.25289	−2.82673j
	30	0.27891	−2.09223j	−.004764	+2.40802j	0.27916	−2.80779j

$n=0$ corresponds to the fundamental oscillation just discussed, whereas $n=1$ corresponds to the first spurious frequency. It is interesting to note that the ratio of spurious to fundamental starting current at fixed Q/N is independent of Q/N, and equal to 81. This is the value which is approached asymptotically for large QC, as described in a later section.

EFFECT OF VELOCITY DISTRIBUTION ON STARTING CURRENT

There is reason to believe that velocity-distribution effects can produce a significant increase of starting current for backward-wave oscillation; in fact, under certain conditions the oscillation theoretically cannot start for any current. This velocity distribution arises because negative charge in the beam depresses the potential of the beam center below that at the outer edge, with consequent reduction in average electron velocity.

If the beam current is considerably less than that corresponding to limiting perveance, solution of Poisson's equation can be carried out for a hollow beam of mean radius b and thickness t. The velocity-distribution function is rectangular, roughly centered near u_0 and of total width Δu_0 where

$$\frac{\Delta u_0}{u_0} = \frac{t}{8\pi\epsilon_0 b}\sqrt{\frac{m}{2e}}\,\frac{I_0}{V_0^{3/2}}. \qquad (30)$$

Watkins and Rynn[17] and Pierce[18] have analyzed the effect of a velocity distribution of the electron beam upon the gain of a traveling-wave tube. The secular equation is modified by the addition of a term entering just as QC; the presence of a rectangular velocity distribution may be taken into account approximately by replacing QC in the secular equation with

$$Q'C = QC + \frac{1}{16C^2}\left(\frac{\Delta u_0}{u_0}\right)^2. \qquad (31)$$

The starting current of a backward-wave oscillator is implied by Fig. 2 and Table I; let us represent this relationship as

$$CN = f'\{Q'C, L\}; \qquad (32)$$

or using (6), (30) and (31),

$$CN = f'\left\{\left[\frac{Q}{N}(CN) + D(CN)^4\right], L\right\}, \qquad (33)$$

where

$$D = \frac{l^2 m}{128\pi^2\epsilon_0^2 b^2 e N^4 K^2 V_0} = 0.574 \times 10^8 l^2/b^2N^4K^2V_0. \qquad (34)$$

This expression can of course be plotted as

$$CN = f\{Q/N, D\} \qquad (35)$$

as has been done in Fig. 5. The interesting result should be noticed, when the velocity-distribution parameter $D \gtrsim 16$—*no* CN exists for backward-wave oscillation. That is,

$$N > 44\frac{l^{1/2}}{b^{1/2}K^{1/2}V_0^{1/4}} \qquad (36)$$

for oscillation to be possible at all.

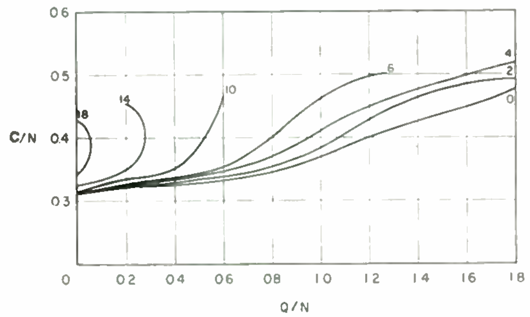

Fig. 5—CN vs Q/N at start of oscillation for backward-wave oscillator, with velocity-distribution parameter D as parameter; D for a hollow beam is given by (34). Here $L=0$.

Experimentally, we find that oscillation does indeed cut off at low voltages—there is a lowest voltage below which oscillation cannot be obtained. At the cutoff, we find N about four times the predicted value, so the above theory is verified only qualitatively. Perhaps a theory based upon a more realistic model than the preceding would improve the agreement.

This phenomenon may have two applications. For higher modes of oscillation, using (28), we predict

$$N > 44(2n+1)\frac{l^{1/2}}{b^{1/2}K^{1/2}V_0^{1/4}}, \qquad (37)$$

so that for

$$132\frac{l^{1/2}}{b^{1/2}K^{1/2}V_0^{1/4}} > N > 44\frac{l^{1/2}}{b^{1/2}K^{1/2}V_0^{1/4}} \qquad (38)$$

fundamental oscillation should be possible but the undesirable spurious oscillation ruled out.

Sometimes a traveling-wave tube amplifiers employ circuits with unwanted backward-wave components which may lead to parasitic backward-wave oscillations. R. Weglein has suggested that if (37) is not satisfied for these backward waves, the parasitic may be suppressed.

Efficiency

The efficiency of a backward-wave oscillator is much more difficult to compute exactly than are the starting current and frequency, since nonlinear or limiting processes are involved. Nevertheless, as Pierce[19] suggested, the linear theory may be used to shed some light on this important problem through the assumption that the linear equations apply up to the limit or saturation point

$$\frac{|i|_{\max}}{I_0} = \alpha. \qquad (39)$$

The linear ballistic equation (3) was derived from the nonlinear one by the assumptions $i/I_0 \ll 1$, $v/u_0 \ll 1$. Nevertheless, one might expect to obtain at least order-of-magnitude efficiencies by letting $\alpha = 1$ or thereabouts

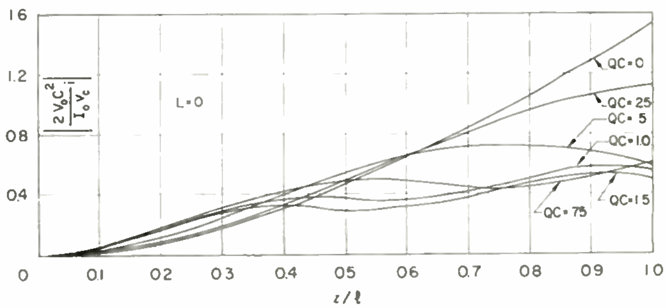

Fig. 6—Theoretical plot of alternating component of beam current $|(2V_0C^2/V_c(0)I_0)i|$ vs normalized distance z/l from gun end of tube. Pierce space-charge factor QC is parameter. Tube loss $L=0$ is shown, but $L=2$, and $L=4$ were also computed.

—Watkins and Grow[10] have suggested $\alpha = 2$ for $QC < 0.25$ and $\alpha = 1$ for $QC > 0.25$. Their calculations were based on $L=0$ and $z=l$ (collector end of the tube), but they have suggested that the maximum value of $|i|$ may, for intermediate values of QC, occur for $0 < z < l$.

[17] D. A. Watkins and N. Rynn, "Effect of velocity distribution on traveling-wave tube gain," *Jour. Appl. Phys.*, vol. 25, pp. 1375–1379; November, 1954.

[18] J. R. Pierce, private communication.

[19] J. R. Pierce, *loc. cit.*, chap. 12.

To calculate i as a function of z/l, (14) is combined with (16) with the result that

$$\exp(j\beta_e z)\left(-\frac{2V_0C^2}{I_0V_c(0)}\right)i = \frac{\exp(\beta_e z\delta_1)}{(\delta_1-\delta_2)(\delta_1-\delta_3)} + \frac{\exp(\beta_e z\delta_2)}{(\delta_2-\delta_3)(\delta_2-\delta_1)} + \frac{\exp(\beta_e z\delta_3)}{(\delta_3-\delta_1)(\delta_3-\delta_2)}. \quad (40)$$

Fig. 6 (previous page) shows several curves of $|[2V_0C^2/V_c(0)I_0]i|$ vs z/l to illustrate ($L=0$, 2, and 4 decibels were calculated, but only $L=0$ is shown). These curves may have QC and L as parameter; under assumption (39) their maxima may be used to determine [using (1)]

$$\frac{\eta}{\alpha^2C} \equiv \frac{|P|}{I_0V_0\alpha^2C} = \frac{1}{\left(2\dfrac{2V_0C^2}{I_0|V_c|}\right)^2|i|^2_{\max}}, \quad (41)$$

then

$$\frac{\eta}{\alpha^2(\omega_q/\omega)} = \frac{1}{4\sqrt{QC}\left(\dfrac{2V_0C^2}{I_0|V_c|}\right)^2|i|^2_{\max}}. \quad (42)$$

This derivation uses the fact that $V_c=V$ where $i=v=0$, which follows from (11) through (15) inclusive. Fig. 7 shows $\eta/(\alpha^2\omega_q/\omega)$ versus QC with L as parameter; the heavy line shows $\eta=\alpha^2\omega_q/\omega$ or $\eta/\alpha^2C=\sqrt{4QC}$.

$$\omega_q/\omega = \sqrt{4QC^3} \quad (43)$$

is the ratio of plasma frequency to operating frequency. Reduction in efficiency in the presence of distributed loss of two and of four decibels is also shown in Fig. 7. The effect of this loss is equivalent to that of a lumped attenuation $L'<L$ placed between a lossless tube and the load. The value of L'/L vs QC is plotted in Fig. 8.

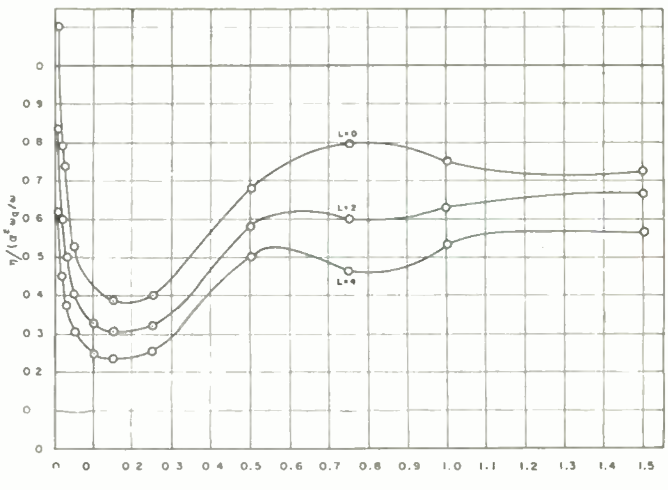

Fig. 7—Theoretical plot of normalized efficiency $\eta/(\alpha^2\omega_q/\omega)$ vs Pierce space-charge factor QC, with total distributed loss L in decibels as parameter.

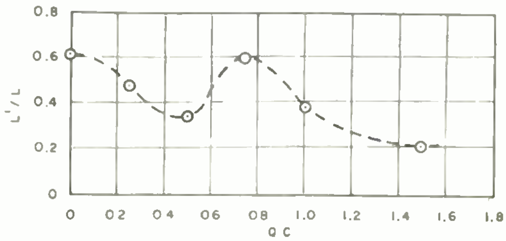

Fig. 8—Theoretical plot of L'/L vs Pierce space-charge parameter QC. L' is the value of loss in decibels in an external pad which will degrade the tube efficiency by the same amount as a loss L decibels distributed along the internal circuit. Efficiency η_0 in loss-free case is related to efficiency in presence of distributed loss L decibels through $\eta=\eta_0 10^{-\alpha L'/10}$.

Experiments on Starting Current and Frequency

An experimental check of the preceding theory for starting current and frequency has been carried out. A special tube constructed for this purpose is shown in Fig. 9. The slow-wave circuit was a helix, and a hollow annular beam of electrons was used. Interaction of the beam with the helix occurred over a length which could be adjusted while the tube was in operation. The interaction took place from gun end of helix to collector, a movable iron affair (Fig. 10, next page) which could slide on ceramic rods when it was pushed by the magnetic armature. This armature was moved by a small solenoid, operated on alternating current for ease of motion, and located outside the main magnetic focusing field. The helix was well terminated for rf by four ceramic rods coated with aquadag which slide along with the collector.

The boundary of the helix at the collector end was taken as the position of these attenuator rods. Power from the collector end of the helix was matched to a coaxial line with a transition developed by B. Keach.

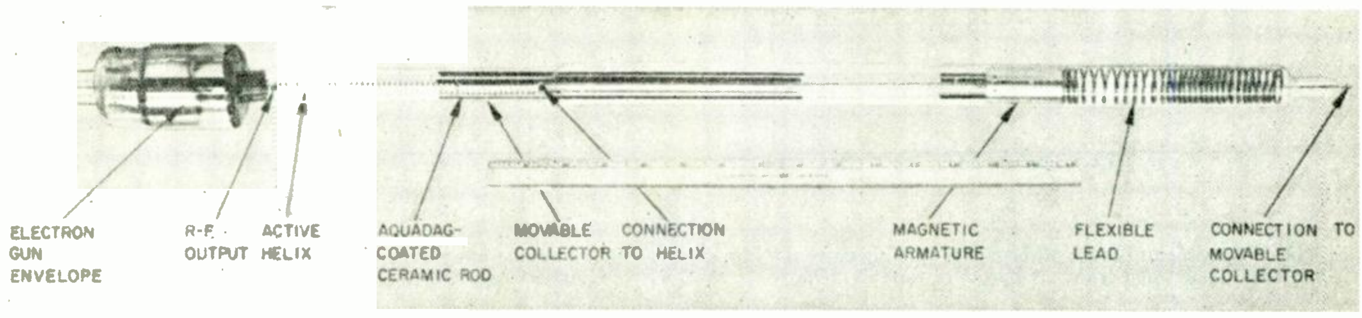

Fig. 9—Photograph of movable collector tube.

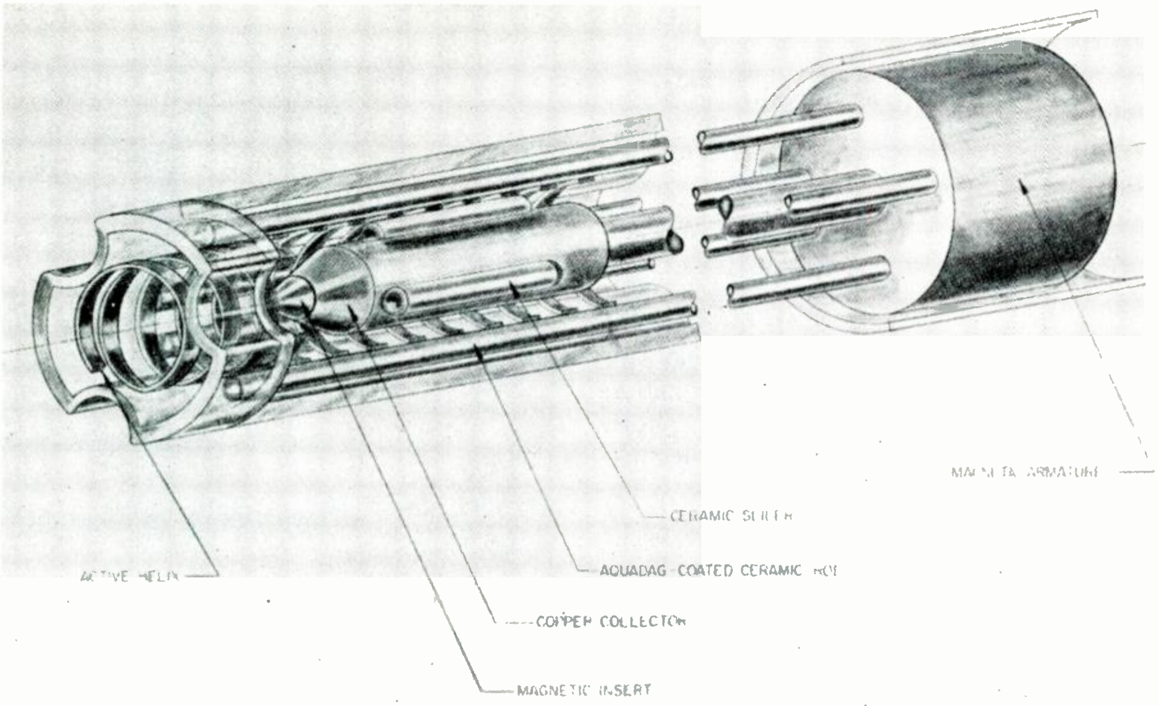

Fig. 10—Artist's sketch of the movable collector.

Pertinent dimensions are outlined in Table III.

TABLE III—Helix and Beam Dimensions of Experimental Tubes

Helix Pitch	0.1429 inch
Helix Tape Width	0.060 inch
Helix Tape Thickness	0.020 inch
Helix OD	0.508 inch
Helix ID	0.468 inch
Beam OD	0.444 inch
Beam ID	0.364 inch
DLF at 3,555 mc	0.923

Starting current for backward-wave oscillation at 2,965 mc was measured as a function of length l and the normalized results compared with the theory in Fig. 11. CN values were computed from

$$CN = \left(\frac{K}{4V_0}\right)^{1/3} \frac{506}{\lambda_0\sqrt{V_0}} I_s^{1/3} l. \tag{44}$$

The agreement is seen to be excellent despite the wide range of starting currents, 1.8 to 150 ma. Data were taken cw for lengths between 5.66 and 3.16 inches inclusive, and with 1-microsecond pulses for lengths between 3.16 and 1.66 inches inclusive. Points taken in both ways are shown for the 3.16-inch length. CN was computed for each length using the impedance $K = 5.63$ ohms chosen for best fit. K computed from Watkins and Ash[15] is 4.6 ohms. Q was obtained as outlined in the Appendix.

Starting voltage was also measured for each length. These data were reduced to a form suitable for comparison with the theoretical results of Table I, and plotted in Fig. 12. The agreement between theory and experiment is again good, except for the highest currents. The plot was made by noting that

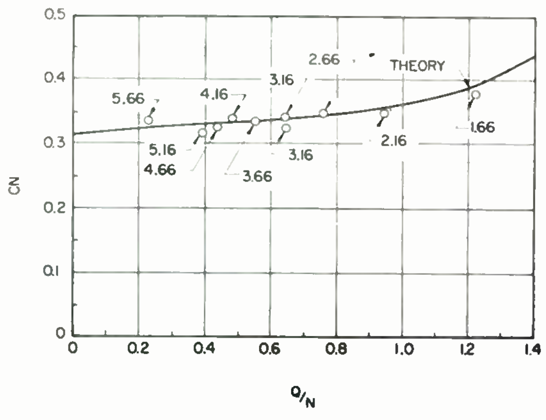

Fig. 11—Experimental check of theory for starting current. Numbers beside each point are length of tube in inches

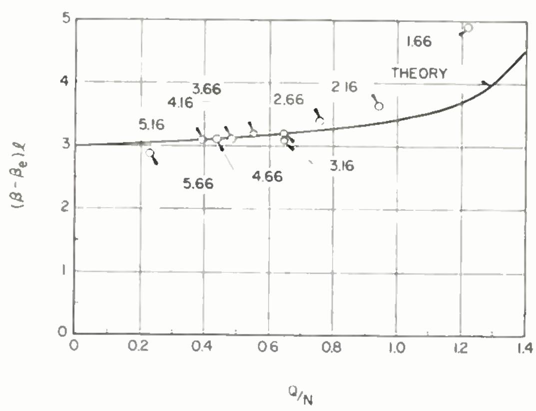

Fig. 12—Experimental check of theory for starting frequency. Numbers beside each point are length of tube in inches.

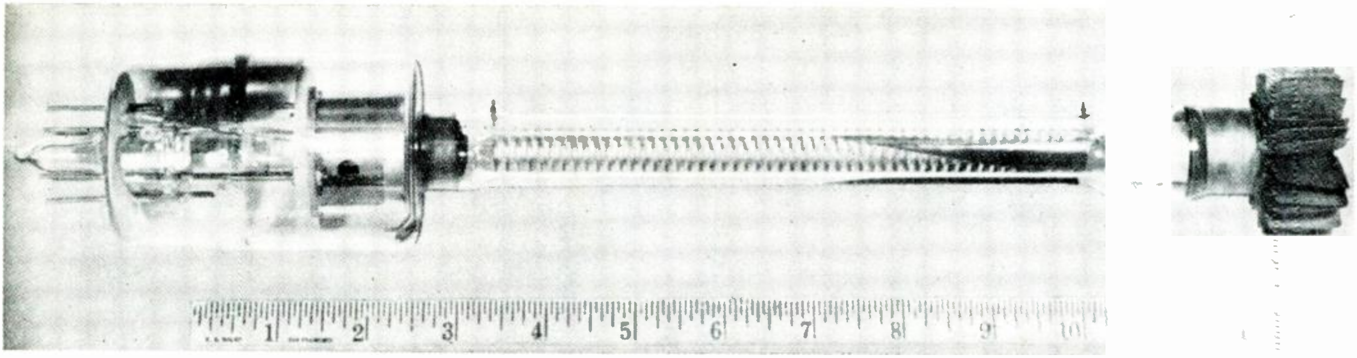

Fig. 13—Photograph of tube used for cw efficiency measurement of Figs. 14, 15, and 16.

$$(\beta - \beta_e)l = \frac{2\pi 506}{\lambda_0} \frac{l(\sqrt{V_0} - \sqrt{V_{0s}})}{\sqrt{V_0 V_{0s}}}, \tag{45}$$

where V_{0s} is the voltage necessary to accelerate electrons from rest to the phase velocity of the circuit at the free-space oscillation wave-length λ_0. λ_0, l, and V_0 are measured experimentally. V_{0s} is inferred from experimental measurement of the helix pitch p (which was uniform to 0.1 per cent) and the frequency f_1 for which $k_d a = 0.5$ in accordance with the equation

$$\frac{\sqrt{V_{0s}}}{506} = \frac{pf}{c(1 - f/2f_1)}. \tag{46}$$

Here f is the oscillation frequency. f_1 could easily be determined as 3,555 mc because of a gap in the oscillation spectrum about 50 mc wide, centered on this frequency (see later section entitled "Point $k_d a = 0.5$"). The result is $\sqrt{V_{0s}} = 31.13$. This, of course, assumes that the dielectric loading factor at the test frequency f is the same as at f_1.

An attempt has been made to reconcile the discrepancy occurring at highest currents by allowing for space-charge depression of average potential. The effect is in the right direction, but too small by a factor of about four.

Experiments on Efficiency

A set of measurements of the efficiency of a backward-wave oscillator was made to determine the range of validity of the preceding efficiency theory. The tube used (Fig. 13) is similar to the movable-collector tube, except that it has a fixed length of active helix and a large copper collector extending through the vacuum envelope.

The results of the measurements are shown in Fig. 14, along with theoretical plots based on $\eta = \omega_q/\omega$ or $\alpha \cong 1.2$. Toward the higher-frequency end of the band, the agreement is seen to be fairly good.

A possible reason for discrepancy observed at lowest frequencies might be distribution in electron velocity across beam because of space charge presence. As Wat-

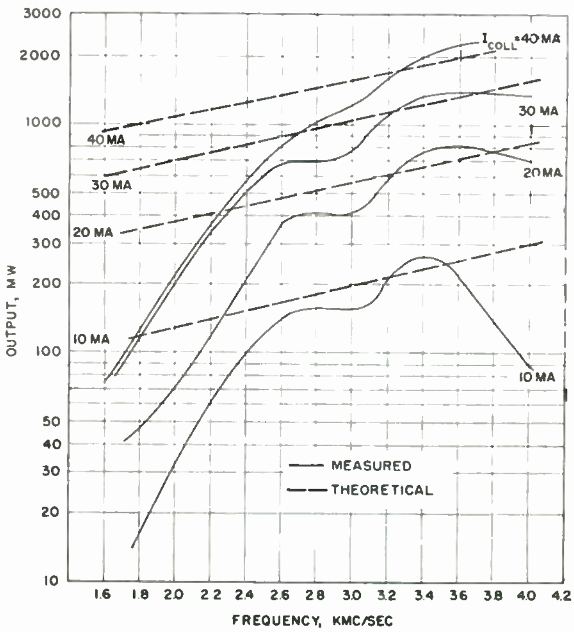

Fig. 14—Power output vs frequency for backward-wave oscillator, beam current as parameter; theoretical curves based on efficiency equal to ω_q/ω.

kins and Grow[20] pointed out, (31) suggests $\Delta u_0/u_0 C$ might be an appropriate parameter to use in an empirical search for such an effect. Hence, $\eta/(\omega_q/\omega)$ is plotted vs $\Delta u_0/u_0 C$ in Fig. 15, next page. Simple theory predicts a straight line $\eta/(\omega_q/\omega) \cong 1$. Apparently there is a distinct correlation between large values of $\Delta u_0/u_0 C$ and reduced efficiency. Fig. 16 contains a plot of helix loss vs frequency.

Effect of Tube Length on Efficiency

An empirical investigation of the effect of length of the tube on efficiency was made with the movable collector tube. This single run was made at a frequency of 2,950 mc and a beam current of 150 ma. Results are shown in Fig. 17. In this run, the pulse-power input was constant at 180 watts, and nothing was changed but the length of the tube. Evidently the output and efficiency

[20] D. A. Watkins and R. Grow, private communication.

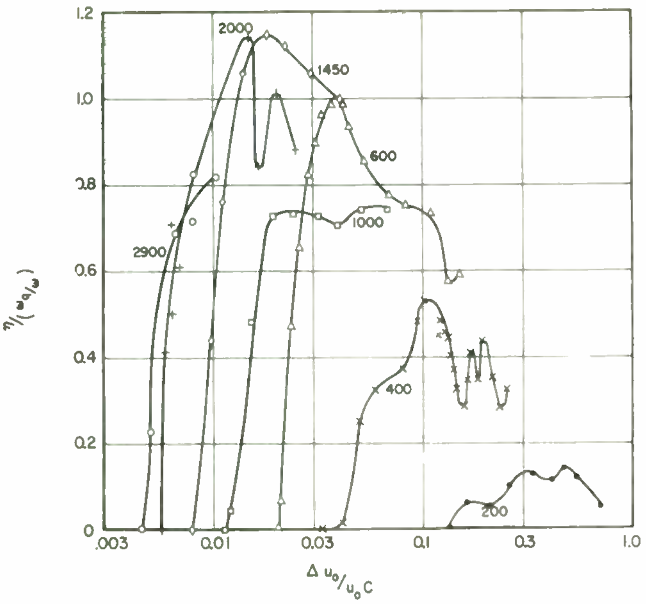

Fig. 15—Experimental plot of normalized efficiency $\eta/(\omega_q/\omega)$ vs quantity $\Delta u_0/u_0C$ related to velocity distribution.

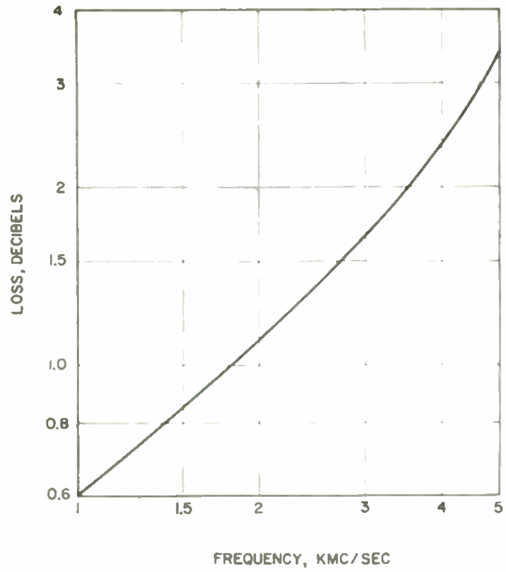

Fig. 16—Plot of measured loss of five inches of helix with zero beam current, tube shown in Fig. 13 and movable-collector tube (Fig. 9). Loss of movable-collector tube was 0.33 db/inch at 2,950 mc.

were constant over a wide range of lengths. This plateau ends on the small-length side when the length is such that the beam current is about twice starting current. It ends on the great-length side when the second frequency of oscillation starts—the total power output abruptly rises because of additional power at the other frequencies. In this case, the second oscillation began when the length was such that the current was about 80 times starting current. The phenomenon of spurious oscillation is more fully discussed in a later section.

Frequency Pulling

The backward-wave oscillator, if its internal load is a perfect match, should have very small frequency pulling. One tube showed, at 3,000 mc, a frequency pulling of only 0.8 mc as a short-circuited load was moved through all phases.

Frequency Pushing

Our experience has universally indicated frequency pushing such that higher beam currents at fixed helix-to-cathode voltage produce a drop in oscillation frequency. This is in conflict with data reported by Kompfner and Williams,[3] but in agreement with that reported by Sullivan.[13] For the tube with the 5-inch helix on which the efficiency measurements were made, at 3,000 mc and 20-ma beam current, the pushing figure was 1.25 mc/ma; with our electron gun, this is equivalent to 0.16 mc/volt tuning rate for the gun anode.

To summarize our results in more general terms, we find empirically that

$$(\beta - \beta_e)l \cong \alpha_2 H + \text{const}, \tag{47}$$

provided $I_0 > 2I_s$, where α_2 is generally about unity (1.5 observed for $V_0 = 200$ volts, 1.4 for 400 volts, and $0.8 \leq \alpha \leq 1.0$ for $600 \leq V_0 \leq 3{,}000$ volts). For $I_s < I_0 < 2I_s$, the empirical value of α_2 at each voltage is larger than for the higher currents. It is interesting to note that the theory for $(\beta - \beta_e)l$ vs H for start of oscillation (Table I) shows $\alpha_2 = 0$ for small H, but $(\beta - \beta_e)l \cong H$ for $H > 3$. These pushing measurements were made on the same tube as were the measurements of Figs. 13 through 16.

Point $k_d a = 0.5$

Several of our experimental tubes show a frequency gap in tuning at the frequency for which $k_d a = 0.5$. k_d is k times the DLF or dielectric loading factor. One, the one with the smallest frequency pulling, has no frequency gap. Another was observed by Mr. Mukaihata of this laboratory to exhibit a peculiar hysteresis effect. Oscillation on the backward harmonic would occur provided the $k_d a = 0.5$ voltage range were approached with the tube oscillating. If this voltage were initially established with zero beam current, and the beam current raised, backward-wave oscillation would not start. Instead, a forward-wave oscillation of the internal-feedback type, at a much lower frequency, would occur. This curious hysteresis was stable and repeatable. These effects are thought to be associated with coupling of forward and backward waves caused by imperfect matches and/or imperfect circular symmetry of the helix.

Stability

It appears that the backward-wave oscillator may be capable of unusual stability. Tests on two tubes at 3,000 mc with beam transit times of 30 and 60 cycles, respectively, showed drift less than 0.01 per cent over several weeks. Thermal stability better than that of a reflex klystron might be expected, since no power need be dissipated on the frequency-control electrode (the helix). As the magnetic field was changed from 750 to 1,500 gauss, the longer of the above tubes changed frequency only 0.01 per cent. Stability measurements require extremely stable power supplies because of the voltage-tuning feature of the tubes.

Spurious Backward-Wave Oscillations

If the beam current of a backward-wave oscillator is increased greatly above starting current, simultaneous oscillation at other frequencies may start. This was illustrated in the experimental plot of Fig. 17; spurious oscillation started when the beam current was 80 times the fundamental starting current. The text following (29) describes how the small-signal, large space-charge theory predicts the value 81. It is important to note that this number is obtained only when spurious and fundamental CN values are compared for the same value of Q/N, since Q/N is a space-charge parameter which is independent of current.

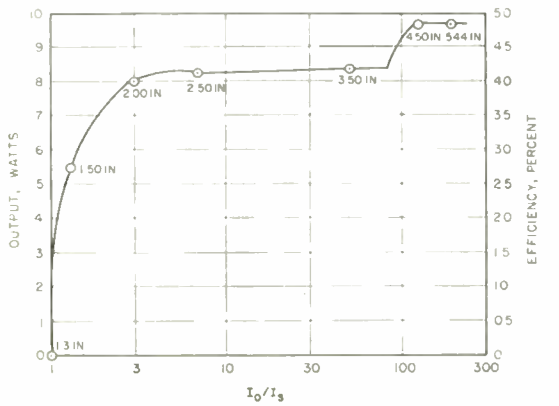

Fig. 17—Experimental plot of efficiency of movable-collector tube with beam current fixed at 150 ma, as tube length is changed. Abscissa is ratio of 150 ma to starting current for each length.

The ratio of spurious to fundamental starting current has been calculated and presented as Fig. 18. The abscissa, QC, is the value appropriate to the fundamental at start of oscillation. The calculation was facilitated by extrapolation of a table appearing in Walker's[8] paper. The extrapolation was accomplished by noting that CN for the spurious oscillation is 3.00 times CN for the fundamental if $QC>0.25$ and if the CN values are compared at the same value of QC. This factor of 3 can be explained physically by noting that a backward-wave oscillator operating in the first spurious mode is similar to a fundamental-mode oscillator of one-third the length, provided each is operating at the same value of QC. QC is the space-charge parameter which is independent of length.

Putz[21] has observed that there is a disagreement with the theory in that spurious backward-wave oscillations appear on the high-frequency side as well as the low-frequency side, whereas theory predicts them widely spaced in current, and only on the low-frequency side. We have made some observations which appear to resolve this difficulty; all the frequencies save two are simply intermodulation products of the original two. When the beam current was adjusted to a value barely sufficient to start the second oscillation (observed on a sensitive receiver), only one spurious frequency was observed. It was lower than the main oscillation frequency f by approximately the predicted amount, say δf. For beam currents slightly greater than this threshold value, new frequencies $(f+\delta f)$, $(f+2\delta f)$, $(f-2\delta f)$, etc., were observed. The equal spacing of these frequencies was checked with an accurate wavemeter. All the frequencies are thus of the form $nf-m(f-\delta f)$ as expected for modulation products (n and m are integers). Their amplitude is exceedingly small when the amplitude of the oscillation at $(f-\delta f)$ is small. A final check showed that when the beam current was below the oscillation threshold but the collector end of the helix was excited from an external source with a frequency $f-(\delta f)'$, the oscillator output did indeed contain many frequencies of the form $f\pm n(\delta f)'$.

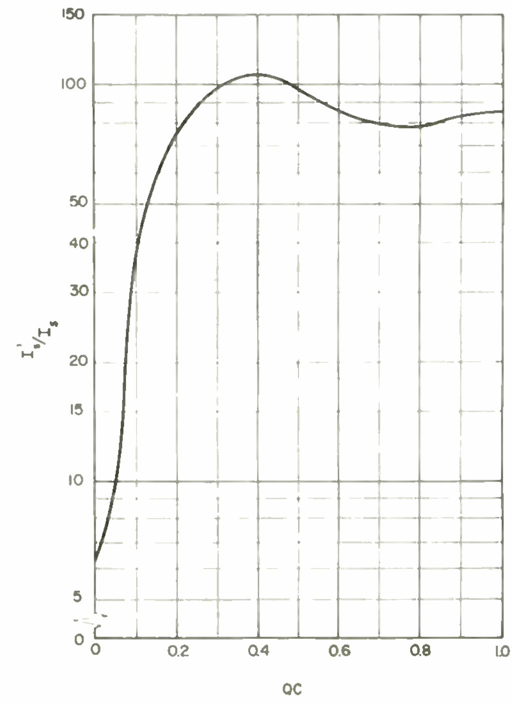

Fig. 18—Theoretical ratio I_s'/I_s of spurious to fundamental start-oscillation currents in backward-wave oscillator, vs Pierce space-charge parameter QC appropriate to start of oscillation for the fundamental.

Appendix

Evaluation of Q and ω_q/ω

Many quantities are used to express space-charge density in traveling-wave tubes, the Pierce "cut-off mode parameter" Q and "space-charge parameter" QC, as well as ω_q/ω, Q/N and H. Each of these can be useful in particular cases. QC is independent of tube length, Q/N is largely independent of beam current, and H is largely independent of circuit impedance. ω_q/ω is the ratio of the frequency of Hahn-Ramo[22] plasma oscilla-

[21] J. Putz, "Folded line traveling-wave amplifier," IRE Conference on Electron Devices, Ottawa, Ont., Can.; June, 1952.

[22] S. Ramo, *loc. cit.*

tion to the operating frequency. Let us derive the relation between the Pierce parameters and ω_q/ω. As Pierce[23] has pointed out, we might expect that when the circuit phase velocity is sensibly removed from the electron velocity ($|b| \gg 1$), the three waves of Pierce should be the circuit wave

$$\delta \cong -jb + d \tag{48}$$

and the two Hahn-Ramo space-charge waves. Indeed, for $|b| \gg 1$ the former is approximately a solution of (10). The other two are

$$\delta \cong +j\sqrt{4QC} \tag{49}$$

corresponding to propagation constants for the space-charge waves of

$$-\Gamma = -j\beta_e(1 \mp \sqrt{4QC^3}), \tag{50}$$

whence follows

$$\omega_q/\omega = \sqrt{4QC^3}. \tag{43}$$

Thus, ω_q/ω can be evaluated from a knowledge of Q, a quantity evaluated for the circularly symmetrical mode of the sheath helix from field-theoretical considerations by Fletcher.[24] Pierce[25] has suggested that this value of Q, which we shall call Q_s, should be corrected to give the Q value appropriate to a component of impedance K through the equation

$$Q = Q_s \frac{K_s}{K} \cdot \tag{51}$$

This can be justified by arguing that only in this way can ω_q/ω be held constant, or Pierce's[26] capacitance C_1 be held constant, as circuit impedance K is changed. K_s above is the sheath-helix impedance in the circularly symmetric mode.

$$K_s \cong \frac{1}{2} \frac{\beta_e}{k} F^3(\gamma_0 a) I_0^2(\gamma_0 b), \tag{52}$$

where k is the free-space propagation constant $\gamma_0 \cong \beta_e$, a is the helix radius, b is the beam radius and

$$F(\gamma_0 a) \cong 7.154 \exp(-0.6664\gamma_0 a). \tag{53}$$

A plot of Q_s calculated from Fletcher's theory is presented in Fig. 19.

A plot from which ω_q/ω can be computed directly, based on the use of Fig. 19 plus (43), (51), (52), and (53) is given in Fig. 20. The ordinate of this figure, when multiplied by the square root of the beam perveance, gives ω_q/ω directly.

Fletcher's work, on which Figs. 19 and 20 are based, is restricted to the $n=0$ mode of the sheath helix, whereas backward-wave oscillators more nearly employ the $n=-1$ mode. Also, Fletcher assumes zero thickness for his hollow beam. These considerations make little difference if $\gamma_0 a$ is larger than about 2, and (51) is used.

[23] J. R. Pierce, *loc. cit.*, pp. 126–127.
[24] J. R. Pierce, *loc. cit.*, pp. 242–251.
[25] J. R. Pierce, *loc. cit.*, p. 128.
[26] J. R. Pierce, *loc. cit.*, pp. 109–110.

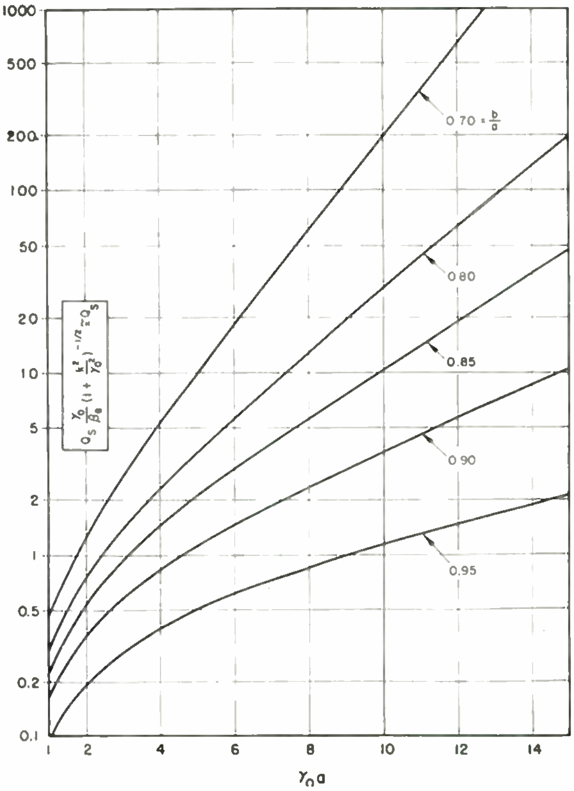

Fig. 19—Theoretical plot, essentially Q_s vs γ_0^2, b/a as parameter. In most cases, $\gamma_0^2 \cong \beta_e$. A zero-thickness hollow beam of radius b is assumed.

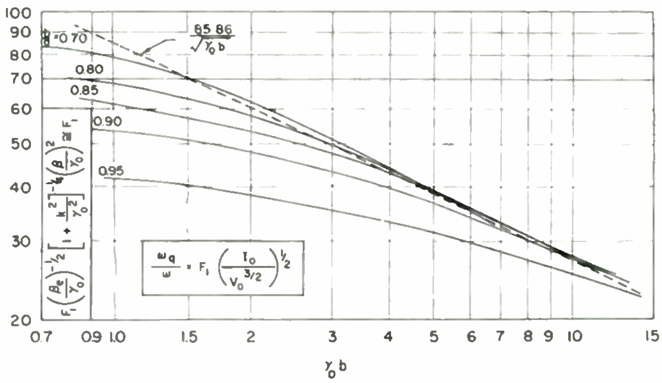

Fig. 20—Theoretical plot, essentially $(\omega_q/\omega)/(I_0/V_0^{3/2})^{1/2}$ vs $\gamma_0 b$, b/a as parameter. In most cases $\gamma_0 \cong \beta_e$. A zero-thickness hollow beam of radius b is assumed.

List of Symbols

b	Beam-velocity parameter defined by (7); beam radius for zero-thickness hollow beam.
b_{max}	Outer radius of hollow electron beam.
b_{min}	Inner radius of hollow electron beam.
$C^3 \equiv$	$I_0K/4V_0$.
C_1	Capacitance to account for effects of space charge; associated with equivalent circuit of Pierce.[25]
c	Velocity of light.

D Velocity-distribution parameter defined by (34).

DLF Ratio of fundamental phase velocity with dielectric loading to that without, in absence of beam current.

d Circuit-loss parameter defined by (7).

E Total axial electric field acting on the stream.

E_z Axial component of electric field of the appropriate space-harmonic of the circuit wave absence of beam, when the total power associated with the wave is P.

F Given by (53).

f, f' Letters denoting functional relation, e.g., $CN=f(QC, L)$.

$2f_1$ Frequency for which helix circumference is one free-space wavelength, times DLF.

$H\equiv$ $2\pi N\omega_q/\omega$, transit angle of beam in radians of plasma frequency. This quantity, while not used by Pierce, is used by Heffner.[6]

I_0 Average beam current; Bessel function.

I_s Beam current for start of lowest, or fundamental, mode of oscillation.

I_s' Beam current for start of first spurious mode of oscillation.

i Alternating current on stream is $\text{Re}\,[i \exp(j\omega t-\Gamma z)]$.

$|i|_{\max}$ Maximum value of $|i|$ along length of tube.

K Impedance parameter $-E_z^2/2\beta^2P$.

K_s Impedance of $n=0$ mode of sheath helix.

k Propagation constant of free space, $2\pi/\lambda_0$. Pierce uses β_0 for this quantity, but most of the other literature uses k.

k_d Propagation constant of free space times DLF.

L Total cold-circuit distributed loss in decibels.

L' Loss in decibels of an external pad necessary to reduce efficiency of lossless tube to efficiency of tube of loss L.

l Length of active portion of circuit.

n Index equal to zero for fundamental mode of oscillation, and to one for first spurious mode.

$N\equiv$ $\beta_e l/2\pi$.

P Power flux associated with wave of propagation constant Γ_1, taken positive if in direction of increasing z, so negative for backward waves.

p Helix pitch.

Q Passive mode parameter defined by Pierce.

Q' Total effective Q in presence of velocity distribution.

Q_s Q for $n=0$ mode of sheath helix, plotted in Fig. 19.

u_0 Average electron velocity.

V Total voltage, defined by (11).

V_0 Average voltage to accelerate the stream to velocity u_0.

V_{0s} Voltage to accelerate electrons to speed equal to circuit phase velocity.

v Alternating component of stream velocity is $\text{Re}\,[v \exp(j\omega t-\Gamma z)]$.

V_{c1}, V_{c2}, V_{c3} Circuit voltages associated with each of the three waves.

V_c Total circuit voltage.

V_1, V_2, V_3 Total voltages associated with each of the three waves.

$V(0)$ Total voltage at $z=0$, equal to circuit voltage at $z=0$.

$x=$ Re δ.

$y=$ Im δ.

z Axial co-ordinate, measured from gun end of circuit.

α Factor equal to ratio between peak alternating current along beam at saturation, and average current; defined by (39).

α_2 Empirical constant in (47).

β Propagation constant of the cold circuit, $\beta=-j\Gamma_1$.

β_e Propagation constant of electrons, $\beta_e\equiv\omega/u_0$.

Γ Axial propagation constant of wave in presence of beam.

Γ_1 Propagation constant of circuit in absence of electron beam; axial electric field component with which interaction occurs is $\text{Re}\,[E_z \exp(j\omega t-\Gamma_1 z)]$.

γ Defined by $-\Gamma^2=\gamma^2+k^2$.

γ_0 Defined by $\beta^2=\gamma_0^2+k^2$.

δ Wave propagation parameter defined by (8).

η Charge-to-mass ratio of the electron; efficiency, or ratio of rf power output to I_0V_0.

λ_0 Wavelength in free space, $2\pi c/\omega$.

ω Radian frequency of oscillation.

ω_q Frequency of Hahn-Ramo plasma oscillation.

Acknowledgment

The author wishes gratefully to acknowledge many extremely profitable technical discussions with his colleagues, and with Profs. D. A. Watkins and P. Hartman. The tubes were built by G. Lee, A. M. Anderson, W. Perkins and others of our engineering section The measurements of starting conditions were made by Prof. Hartman. Many of the mathematical computations were made by K. Higa.

Direction Sensitive Doppler Device*

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Summary—**A simple double-doppler device is described which makes it possible to determine the direction of motion in addition to measuring velocity. The same principle can be employed to measure distances, temperature, or small frequency differences. In addition an application of the device for moving-target indication is described.**

Introduction

THE DOPPLER phenomenon has been used extensively for detecting moving targets and for determining the radial component of the relative velocity between transmitter and target. Very often, it would be advantageous to determine not only this velocity, but to know whether the target is "coming or going." This information is not furnished by a simple doppler device because it cannot distinguish between positive and negative doppler frequencies.

Theoretically, the direction of motion can be found by determining exactly the frequency of the return signal or by watching its amplitude. In practice, neither method can be applied. It is impossible to separate the return signal from the outgoing energy well enough to measure its frequency. Furthermore the amplitude can change at random during motion because of a changing aspect angle so that it is very well possible that the amplitude diminishes at a given instant while the target is approaching.

A method without the above drawbacks will be described which makes it possible to determine direction with very simple means.

The New Method

Two signals at doppler frequency are produced. One is the standard doppler signal. For the second one, an additional phase-shift of 90 degrees is produced between the local signal and the return voltage. This way, a phase-shift of 90 degrees exists between the two doppler signals and it will be shown that, if this shift is positive for increasing distance it is negative for diminishing distance. Hence, a synchronous two-phase motor will turn, say, clockwise for an approaching target and counterclockwise for a receding target. In Fig. 1, it is assumed that a cw transmitter is located at A. A reflecting target T starts moving at time zero from location B with a velocity V which is supposed to be positive for increasing, and negative for decreasing, distance.

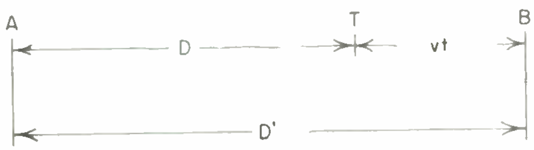

Fig. 1—Location of transmitter and target.

E_1 is the transmitted signal voltage and E_2 is the received voltage.

$$E_1 = E \sin \omega t$$

$$E_2 = KE \sin \omega\left(t - \frac{2D}{c}\right),$$

whereby K is an attenuation factor, c, velocity of light.

$$D = D' \pm vt$$

$$E_2 = KE \sin \omega\left(t - \frac{2D' \pm 2vt}{c}\right)$$

$$E_2 = KE \sin\left(\omega t - \frac{2D'\omega}{c} \mp 2\frac{v}{c}\omega t\right)$$

$$\frac{2D'\omega}{c} = \alpha \text{ represents a fixed phase angle.}$$

$2(v/c)\omega = \omega_d$ represents the angular doppler frequency. The signals E_1 and E_2 are fed into a mixer so that a third signal E_m is produced with the amplitude of the received signal E_2 and with a phase-angle which is the difference between the angles of E_1 and E_2.

$$E_m = KE \cos(\alpha \pm \omega_d t).$$

A second mixer is arranged in such a way that an additional phase-shift of $\pi/2$ is produced between E_1 and E_2.

$$E'_m = KE \cos\left(\alpha + \frac{\pi}{2} \pm \omega_d t\right).$$

Neglecting the constant phase-angle α, we have the following conditions. For increasing distance:

$$E_m = KE \cos(\omega_d t)$$

$$E'_m = KE \cos\left(\omega_d t + \frac{\pi}{2}\right).$$

For decreasing distance:

$$E_m = KE \cos(\omega_d t)$$

$$E'_m = KE \cos\left(\omega_d t - \frac{\pi}{2}\right).$$

A rotating magnetic or electrostatic field can, therefore, be produced whose direction of rotation depends on the direction of the radial relative motion of the target.

Fig. 2 shows a block diagram of the first experimental arrangement. An x-band klystron K is employed as the cw transmitter. The energy is radiated by horn A_T, reflected by the moving target, and received by horn A_R. A small part of the transmitted signal is branched off

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and fed to the two detectors D and D'. The delay line L produces a phase-shift of $\pi/2$ between the two local signals. The return signal is split symmetrically and fed to the detectors. The mixers produce, according to the theory described, the two doppler signals E_m and E'_m.

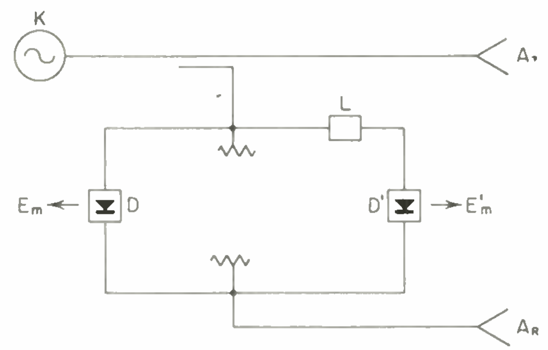

Fig. 2—Block diagram of the first experimental arrangement.

Instead of using mathematics, the operation of the device can also be explained by the use of rotating phasors as shown in Fig. 3. E_1 and E'_1 represent the two local signals, $\pi/2$ radians out of phase. E_2 and E'_2 are the return signals. They rotate with the angular doppler frequency and the mixer output is represented by their projection on the local phasors E_1.

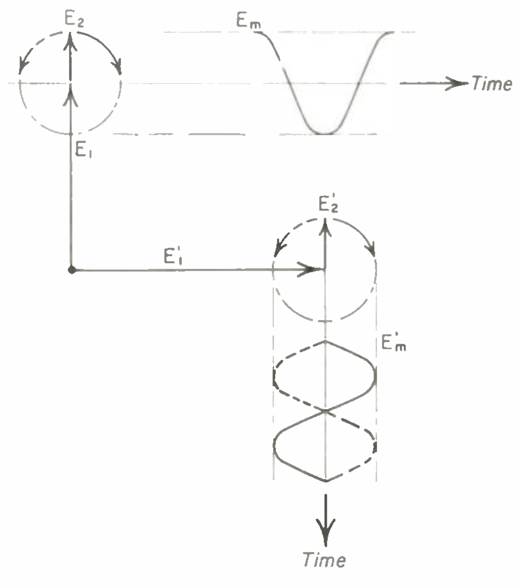

Fig. 3—Phasor representation of the two doppler signals.

Increasing distance corresponds to clockwise rotation and diminishing distance to counterclockwise rotation. Hence, the mixer output voltage E_m is the same for both directions. E'_m, however, is $\pi/2$ radians advanced or delayed with respect to E_m, depending on the sense of rotation of E'_2. A device as shown in Fig. 2 was built and performs according to theory.

Single Antenna Device with Gyrator

The first experimental setup had the disadvantage that two separate antennas had to be employed. A low-loss duplexing scheme with a single antenna can be built, using a gyrator.

Fig. 4 shows the block diagram and the position of the electric field vectors in space. K is the klystron, produc-

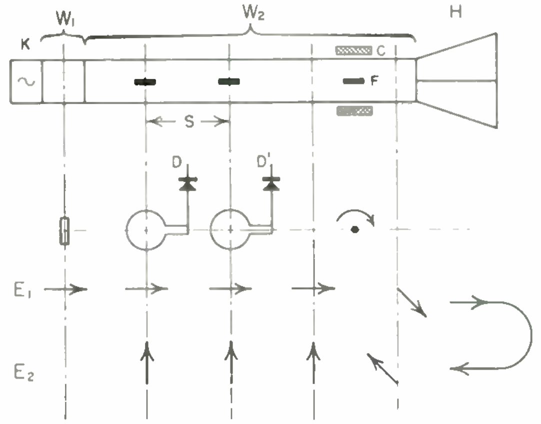

Fig. 4—Single antenna device with gyrator and position of electric fields in space.

ing a horizontally polarized wave E_1 which passes the rectangular waveguide W_1 with the standard TE_{01} mode. Next, the wave transverses the round guide W_2 with the TE_{11} mode. The gyrator F, consisting of a ferrite-rod, surrounded by the coil C, turns the polarization plane clockwise by 45 degrees, so that the energy is radiated by horn H, shifted by a 45-degree angle with respect to W_1. The two detectors D and D' are arranged to receive a very small part of E_1 which serves as the local signal.

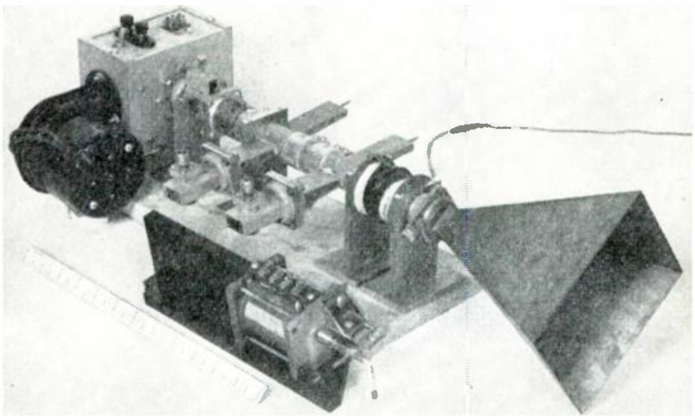

Fig. 5—Photograph of the single antenna device with gyrator.

The signal is reflected by the target and enters the horn with its plane of polarization shifted by -135 degrees with respect to W_1. It is represented by vector E_2. After passing the gyrator, the wave is again turned by 45 degrees so that the plane of polarization is now vertical. The two detectors receive freely the reflected signal and the two doppler signals are produced. The detectors are spaced by the distance $S=(2n+1)\lambda/8$, so that, if the phase-shift between local and return signal is ϕ in D, it is $\phi+(\pi/2)$ in D'. It was shown before that this is the condition for the production of a rotating field by the two doppler signals. Fig. 5 shows the arrangement.

Single-Antenna, Single-Detector Device

Duplexing schemes can be replaced by an arrangement in which the two doppler signals are produced alternately in quick succession. This method is especially applicable if vhf and uhf carriers are employed.

Use of a triode oscillator, the rf output of which is fed to the antenna through transmission line L, is shown in Fig. 6. The reflected signal is mixed with the local signal by the diode action of the grid of the triode. The doppler signal is fed into amplifier A. There is a delay network N inserted in the transmission line which is switched in and out periodically at a rate much faster than the doppler frequency. A synchronously-driven switch feeds the amplifier output alternately to the integrators I_1 and I_2 so that, from their terminals, the two doppler signals can be derived. The mechanical switch shown can, of course, be replaced by an electronic device with an inherently higher rate. In this case, it is advantageous to employ saturable ferrite reactors in the delay line, so that the delay time can be changed alternately by applying a square wave to the reactors.

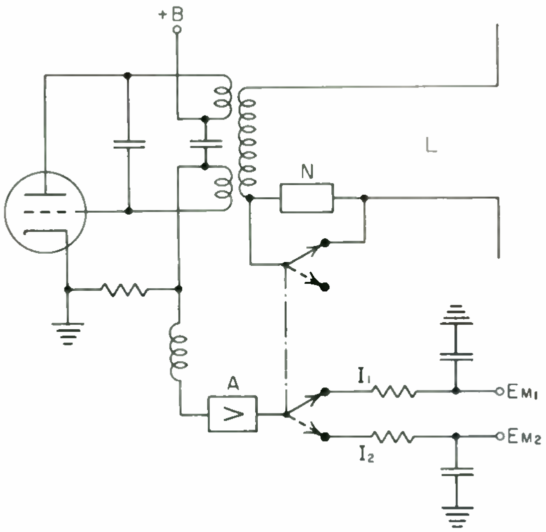

Fig. 6—Single antenna, single detector device.

Applications

Distance Determination

The device can be used to measure the actual distance from the antenna to the target if the original distance is known. Hence, it can be employed as an altimeter for an aircraft. The angular velocity of the synchronous motor is proportional to velocity, whereby, in distinction to conventional doppler devices, positive or negative velocity is measured by clockwise or counterclockwise rotation. The number of armature revolutions is determined by a counter so that, by this integrating process, distance is measured. If an x-band generator is employed, one revolution is obtained for traversing a distance of 1.5 cm so that a high accuracy can be achieved.

If the doppler frequency is so high that the motor response is limited by mechanical inertia, electronic phase comparison and counting devices can be applied.

The new method is especially applicable for the altitude control of a missile in level flight. The motor shaft output can be used directly to correct the altitude (to control the hydraulic system).

Moving-Target Indicator

Continuous revolution of the armature is produced only by continuously moving targets. Hence, it becomes possible to discriminate against ground clatter. The armature inertia is actually put to good use in this application because it serves as a memory device for the doppler frequency. The same end can be achieved electronically only by many narrow-band circuits (vibrating reeds) or equivalent complicated arrangements.

Temperature Measurement

When a sonic wave is transmitted between two fixed transducers in a gas atmosphere, a change in the frequency of the received wave can be observed when the temperature is changed. This "doppler" effect is due to the change of the propagation velocity of sound.[1] If now, instead of conventional doppler methods, the described double-channel system in combination with a synchronous motor is used, the temperature of the gas can be determined at any time if the initial temperature is known. The counter on the motor, or the cycle counter in an electronic system, acts as an integrator for temperature changes in such a way that the correct temperature is indicated independent of whether preceding changes were positive or negative. In this way, a fast, direct-acting thermometer can be designed, which makes it possible to measure temperature in a very short time. Work is now in progress to determine the temperature of the gas in the explosion chamber of a recoilless rifle or in the interior of cylinders in internal combustion motors.

Frequency Measurement

If the frequency of a wave has to be determined, it is normal to compare the unknown frequency with a known one by beat methods. This way, the difference frequency can be measured, but it is not easy to find out whether this difference frequency is positive or negative. By the use of two detectors, arranged in such a way that the two beat-notes are 90 degrees out-of-phase, it becomes readily possible to determine the sign of the difference frequency. Again, a two-phase motor can be employed or an electronic equivalent thereof.

Acknowledgment

The help of H. Dropkin, who built the first model of the doppler device and also designed the required waveguide arrangement, is acknowledged.

[1] H. P. Kalmus, A. L. Hedrich, and D. R. Pardue, Diamond Ordnance Fuze Labs., Washington, D. C., Tech. Rep. No. TR-72.

High-Stability Bridge-Balancing Oscillator*

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Summary—**A crystal-controlled frequency standard having a frequency stability of a few parts in 10^{10} per day is described. The oscillator employs a 1-megacycle GT-cut crystal unit in a high-sensitivity bridge-balancing frequency-correction system. The crystal serves both as the resonant element for the oscilaltor and for a bridge whose unbalance is an indication of departure of the oscillator frequency from the crystal series resonance. The unbalance, greatly amplified, is used to AFC the oscillator.**

Introduction

High-stability frequency standards are required for use in navigation systems, precise clocks, and for many types of scientific measurements and experiments. A frequency stability of 1×10^{-9} per day is often desirable. The oscillator to be described here is of interest because its frequency is almost entirely independent of tube, component, and supply-voltage changes. The first model, which employs relatively inexpensive, commercially available components, has a stability of a few parts in 10^{10} per day.

The resonant frequency of a quartz crystal-unit can be made to exhibit a high frequency stability. If it is connected to a suitable source of energy, such as an amplifier-limiter combination, a stable oscillator can be produced. Such an oscillator always departs from the resonant frequency of the crystal unit because of the phase shift present in the amplifier. There are two basic methods, which may be equivalent under some conditions, for decreasing the amplifier phase shift.

Oscillator Systems

One method employs a low-impedance coupling between the resonator and the amplifier.[1–4] It can be shown that the phase shift is directly proportional to the magnitude of the coupling impedance. However, with a limited transconductance, this impedance cannot be decreased below the point where gain is obtained. The second method employs inverse feedback in the amplifier to decrease the effective phase shift.[5] Recognizing the fact that phase shifts add, but gains multiply, in a multistage amplifier, it might appear that a large feedback factor could be employed to obtain excellent phase stability. Unfortunately it becomes difficult to control the sign of the feedback outside the pass band of the amplifier, particularly if three or more interstage couplings are required, and instability may be encountered. In a practical system it is difficult to obtain a phase-shift reduction of more than 50 to 1 by using inverse feedback.[6]

It should be noted here that the fractional (percentage) frequency change in a feedback oscillator is proportional to frequency, making it difficult to fully realize the capabilities of modern 1-megacycle and 5-megacycle AT-cut crystal units.[7] The use of high-frequency crystal units is desirable because they are less subject to the effects of mechanical shock and aging than are 100-kilocycle units. Furthermore they appear to be less expensive and more readily produced.

Bridge-Balancing Oscillator

The characteristics of a crystal unit can be measured to a high degree of precision in a resonance bridge. If an oscillator is adjusted to resonance with the crystal unit, frequency can be measured with a resolution of 1×10^{-10} or better.[8] Such a system constitutes a useful frequency standard, particularly if continuous frequency adjustment is made by automatic means.[9,10] Fig. 1(a), next page, is a block diagram of a suitable system. The crystal unit is connected in a low-impedance bridge so constructed that the effects of parasitic reactances are very small. The output of the null amplifier is used, in conjunction with a phase detector, to control the frequency of the oscillator exciting the bridge. In this manner the oscillator can be made to approach resonance with high precision. One limitation to the degree of precision is the noise in the system. In Appendix I it is shown that the equivalent fractional frequency deviation $\Delta f/f$ resulting from bridge thermal noise is

$$\frac{\Delta f}{f} = \frac{2}{Q}\sqrt{\frac{kT}{P_{R}t}}, \qquad (1)$$

where Q is the figure of merit of the resonator, k is Boltzmann's constant, 1.37×10^{-23} joule per degree

* Original manuscript received by the IRE, January 17, 1955; revised manuscript received, February 25, 1955.

† National Bureau of Standards, Washington 25, D. C.

1 J. K. Clapp, "An inductance-capacitance oscillator of unusual frequency stability," Proc. I.R.E., vol. 36, pp. 356–358; March, 1948.

2 W. A. Roberts, "An inductance-capacitance oscillator of unusual frequency stability," Proc. I.R.E., vol. 30, pp. 1261–1262; October, 1948.

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9 Such an oscillator was described by Norman Lea of British Marconi during his recent visit to the U.S.A.

10 T. A. Pendleton, "A System for Precision Frequency Control of a One-Hundred Kilocycle Oscillator by Means of a Quartz-Crystal Resonator," M.S. thesis, University of Maryland, May, 1953. (Summary of work done at the National Bureau of Standards in fall of 1952 and spring of 1953.)

Kelvin, T is the absolute temperature of the bridge, P_R is the power dissipated in the resonator, and t is the time required to make a measurement. For the oscillator to be described, $Q=10^6$, $T=320$ degrees K, $P_R=2.5\times10^{-6}$ watts, and $t\approx0.06$ sec. therefore, $(\Delta f/f)\approx3\times10^{-13}$.

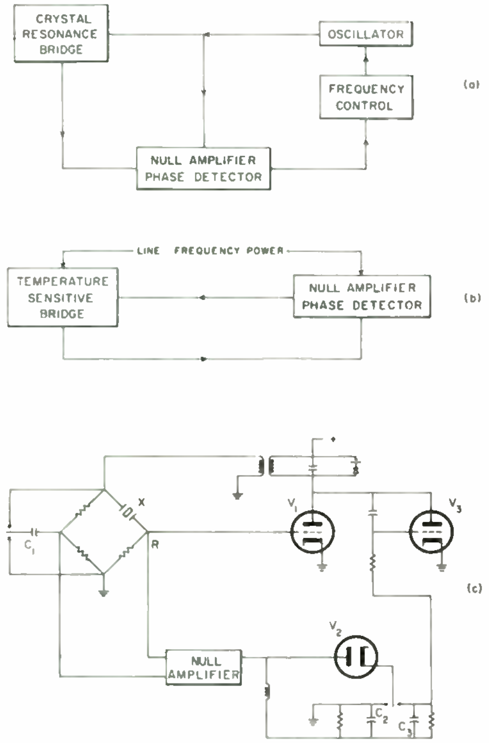

Fig. 1—(a) Block diagram of bridge-balancing oscillator, (b) block diagram of temperature-control system, (c) simplified schematic diagram showing components of bridge-balancing oscillator.

In Appendix II the feedback frequency-control system of Fig. 1(a) is analyzed, and it is found that the ratio of fractional frequency deviation with feedback $(\Delta f/f)_c$ to that without feedback, $\Delta f/f$, is given by

$$\frac{\left(\frac{\Delta f}{f}\right)_c}{\frac{\Delta f}{f}} = \frac{1}{1+\frac{1}{4}K_1QAV_1}, \tag{2}$$

where K_1 is twice the fractional frequency change produced by the frequency control per unit control voltage, Q is the figure of merit of the resonator, A is the effective voltage gain from the bridge output to the frequency-control input, and V_1 is the bridge-input voltage. For the experimental oscillator $K_1=10^{-7}$/volt, $Q=10^6$, $A=10^7$, and $V=0.01$ volt. Therefore

$$\frac{\left(\frac{\Delta f}{f}\right)_c}{\frac{\Delta f}{f}} \approx \frac{1}{2500}.$$

It is comparatively easy to produce a simple oscillator with a stability of a few parts in 10^7, and this can then be improved by a factor of 10^3 or more by a bridge-balancing frequency control, subject to the limitations of thermal noise and the stability of the crystal unit itself.

Temperature Controller

It must be pointed out that in order to realize the capabilities of such a system the crystal unit must be operated in a proper environment. Specifically, a constant, low driving power must be employed, and a constant temperature must be maintained. The crystal-unit power can be controlled by the use of the proper oscillator circuitry; however, the temperature controller deserves some additional consideration. The limitations and performance of a temperature controller are given below. Before proceeding with the analysis, however, it should be pointed out that such factors as bridge-arm instability and heat loss through leads entering the controlled chamber may make the performance of the oven poorer than an analysis assuming ideal conditions might indicate.

A simple method is shown in Fig. 1(b). A temperature-sensitive bridge is employed, and the output of a high-gain bridge-unbalance amplifier is fed back to the bridge input to provide the necessary heating power. As in the crystal bridge, one limitation to the sensitivity of this system is the thermal noise from the bridge arms themselves. In Appendix III it is shown that the thermal noise is equivalent to a temperature difference

$$\Delta T = \frac{2\sqrt{2}}{\alpha}\sqrt{\frac{kT}{P_Bt}}, \tag{3}$$

where α is the temperature coefficient of resistance of the material used in two of the bridge arms, and P_B is the signal power dissipated in the bridge. (The other quantities are defined above.) Here P_B is a constant power applied to the bridge to produce a signal for driving the amplifier, and is to be distinguished from the output of the amplifier, which is a function of signal level. In the circuit to be described below, P_B is obtained from a small alternating voltage applied to the bridge, while the major part of the bridge power, which must necessarily vary as a function of the bridge unbalance, is obtained by rectifying and filtering the amplifier output. A temperature controller has been constructed with the following constants: $\alpha=0.0045$ per degree C., $P_B=0.02$ watt, and $t=1/10$ second. Therefore $\Delta T\approx1.4\times10^{-6}$ degree C. This places a lower limit on the temperature resolution to be obtained with such a system.

In Appendix IV the feedback temperature controller is analyzed, and it is found that the ambient-temperature reduction factor S is given by

$$S = \frac{1}{A\alpha}\sqrt{\frac{K_2}{P_B(T_1-T_2)}}, \tag{4}$$

where S is the ratio of an oven-temperature change to the corresponding temperature change outside the oven, and K_2 is the power loss through the oven walls per unit temperature difference $T_1 - T_2$. If $A = 10^6$, $\alpha = 4.5 \times 10^{-3}$ per degree C., $K_2 = 0.02$ watt per degree C., $P_B = 0.02$ watt, $T_1 - T_2 = 25$ degrees C., and $S \approx (1/22{,}000)$. Thus a 22 degree C. change in ambient temperature would produce a change of but 10^{-3} degree inside the oven.

Practical Bridge-Balancing Oscillator

There are many possible choices for the components to fill the blocks of Fig. 1(a). Thus it would have been possible to use a commercial high-quality oscillator to drive the bridge, with the addition of a motor-driven frequency control, and an ordinary communications-type receiver might have been used as a null detector. However, in the interest of simplicity the circuits outlined in the incomplete schematic of Fig. 1(c) were chosen. Here V_1 is a single-stage amplifier, which is transformer-coupled to a bridge containing a one-megacycle AT-cut crystal X and a resistance R, which is made equal to the resistance of X. The other two bridge arms are equal resistors of a value somewhat higher than R. The oscillator alone tends to operate at the series-resonant frequency of X, with an over-all frequency stability of about one part in 10^7. Amplitude control is obtained by connecting a biased diode limiter across the plate circuit of V_1. The plate voltage of V_1 is approximately one volt rms. The tuned transformer decreases this to 10 millivolts at the bridge input. A crystal current of approximately 500 microamperes is obtained with a 10-ohm crystal unit.

A small capacitor C_1 is switched alternately across one or the other of the equal bridge arms by means of a chopper to obtain the sense of the unbalance. The bridge output is amplified by a factor of 10^7, rectified by V_2, and capacitors C_2 and C_3 are charged synchronously with the switching of C_1. The difference of the voltages across C_2 and C_3 is used to drive a reactance tube V_3 which controls the frequency of the oscillator over a small range. If there is no difference in frequency between the oscillator and the series-resonant frequency of X, equal outputs are obtained corresponding to the two positions of C_1, and the reactance tube is not actuated. If a frequency error does occur, a direct voltage of the proper polarity is applied to V_3 to produce a correction. The amount of correlation is given by (2), and the constants apply to this oscillator. Although the noise bandwidth of the null amplifier is approximately one kilocycle, the chopper and detector are followed by a resistance-capacitance filter with a band-width of approximately $\frac{1}{4}$ cycle. It can be shown that for small signal-to-noise ratios, the effective bandwidth reduction radio by post-detection filtering is the square root of the ratio of the pre- and post-detection filter bandwidths. Here the detector does not function as a carrier-synchronous detector, but rather as a nonsynchronous detector during each chopper half-cycle. Here the bandwidth-reduction factor is 63, with a resultant effective bandwidth of 16 cycles, and therefore the measurement time t is approximately 0.06 second.

The switching of capacitor C_1 should produce a square-wave modulation of the oscillator. The degree of modulation is very small, however, and the modulation has not been detected after multiplication to 1,000 megacycles.

One interesting feature of the oscillator is that the crystal unit is common to the bridge and the controlled oscillator, permitting some simplification of the equipment.[11] Another point worth mentioning is the use of a diode limiter rather than an automatic-gain-control system for amplitude control. The diode limiter maintains the crystal current constant to ± 1 per cent over a two-to-one plate-supply-voltage range. An automatic-gain-control system would probably require two or three tubes for equal performance. It should also be mentioned that the reactance-tube control voltage is metered, so that the departure from resonance is indicated at all times. The sensitivity of the meter is 1×10^{-11} per small division. This permits the readjustment of the oscillator if large, permanent phase shifts are encountered.

Fig. 2 shows frequency change vs plate-supply voltage for the oscillator (V_1) alone, for the oscillator plus reactance tube with the null amplifier disabled, and for the complete system. It will be noted that the reactance tube degrades the stability of the oscillator. The performance of a two-tube Meacham oscillator employing the same type of crystal unit is shown to permit comparison.

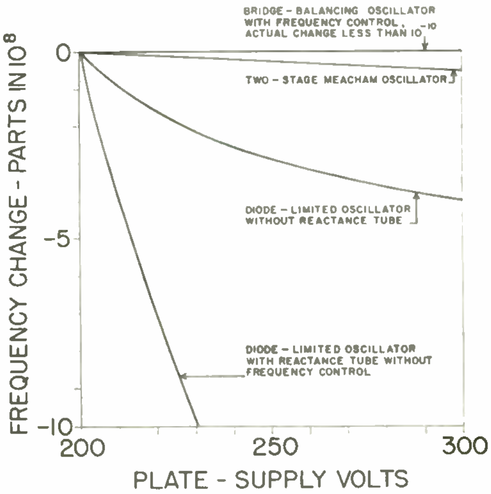

Fig. 2—Frequency change vs plate-supply voltage.

Fig. 3 (next page) is a 24-hour section of a record of beat frequency between the 1-megacycle oscillator and a reference oscillator. The outputs of the two oscillators were mixed after multiplication to a frequency of 1,000 megacycles, and the resulting beat was counted for 100-second intervals, producing a resolution of 10^{-11}. Each minor division represents a frequency change of 10^{-10}. It can be seen that the apparent stability of the oscillator, whose frequency changed $\pm 2 \times 10^{-10}$ during the

[11] N. Lea, "Quartz resonator servo," *Marconi Review*, vol. 11, pp. 65–73; 3rd quarter, 1954.

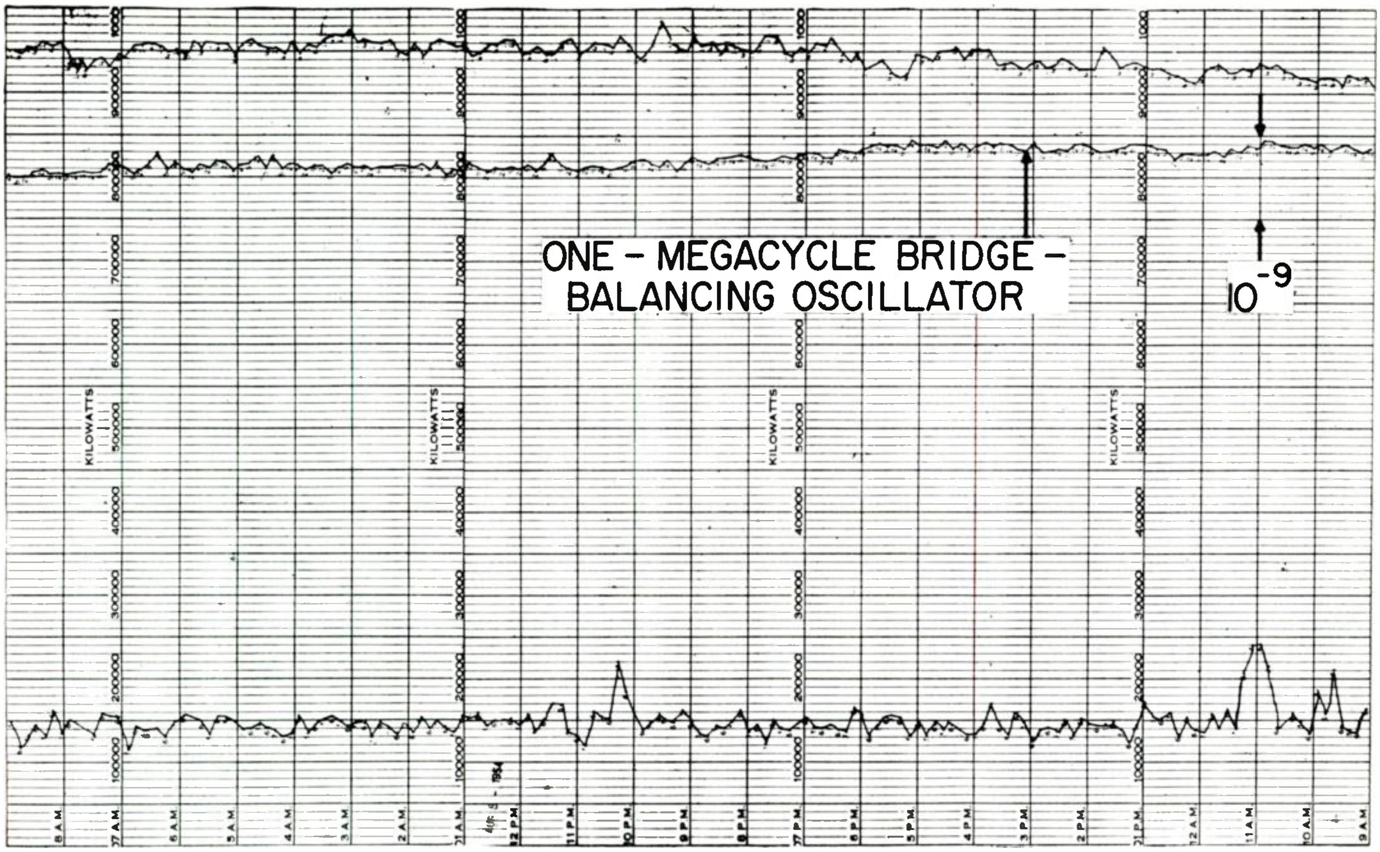

Fig. 3—24-hour frequency vs time record for the bridge-balancing oscillator.

24-hour period, compares favorably with that of the other two oscillators whose records are shown. These two oscillators, as well as the reference oscillator, are of the Meacham type, employing high-precision GT-cut crystal units. Actually the short-time stability (15 minutes) of the one-megacycle oscillator may be better than it would appear because of possible variations in the reference oscillator.

Fig. 4 is a schematic diagram showing the oscillator, reactance tube, and output (buffer) amplifier. Fig. 5 shows the null detector, which contains a three-stage tuned-radio-frequency amplifier and detector. Although a crystal filter is shown, this was not used during the tests and could well have been omitted. Such a filter would be necessary with a higher-gain amplifier. Fig. 6 is a schematic diagram of the temperature controller, which contains a conventional high-gain audio-frequency amplifier and phase detector.

The bridge arms are noninductively wound on an aluminum cylinder of $\frac{1}{4}$-inch wall thickness containing the crystal unit and its trimmer capacitor. The oven is insulated by means of a Dewar flask, whose effectiveness is demonstrated by the fact that replacing the flask with one inch of felt increased the power consumption by a factor of 10 for a given temperature difference.

Heat loss through the oven leads was minimized by the use of small-diameter wire.

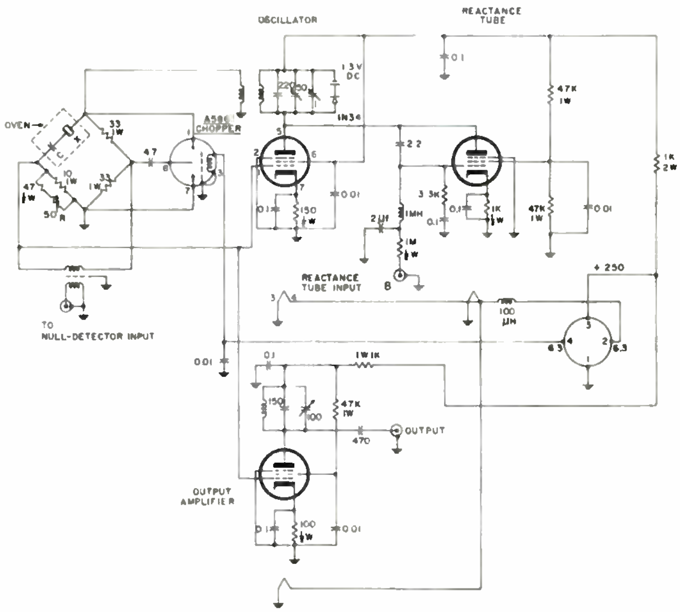

Fig. 4—Schematic diagram of the oscillator, reactance tube, and output amplifier. All tubes type 6AU6. Capacitances greater than 1 in $\mu\mu f$; others in μf unless noted. $x = 1$mc at cut crystal with resistance of 10 ohms ($Q = 10^6$). $C = 1,500$ $\mu\mu f$ to adjust crystal to operating frequency.

Fig. 7, which appears on page 706, is a photograph of the experimental oscillation.

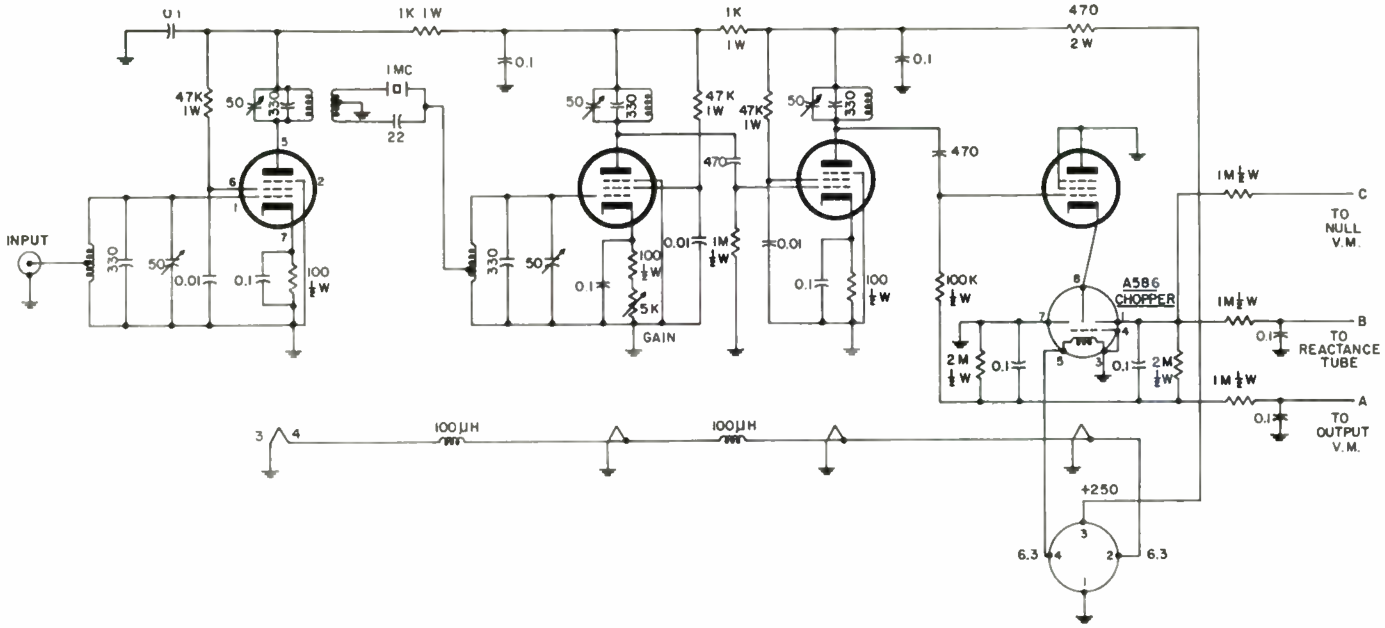

Fig. 5—Schematic diagram of the null amplifier and detector. All tubes type 6AU6. Capacitances greater than 1 in $\mu\mu f$; others in μf.

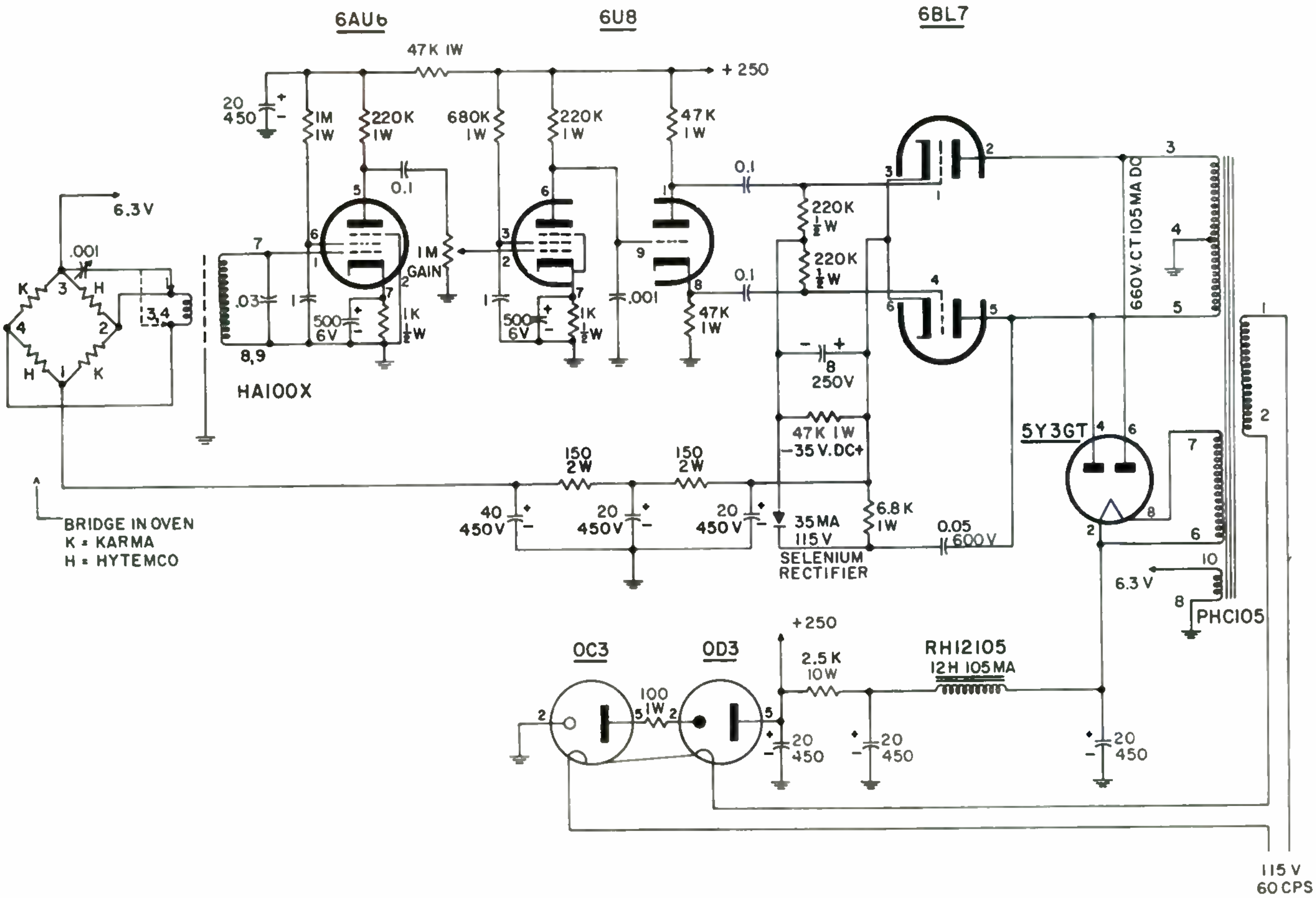

Fig. 6—Schematic diagram of the temperature controller. All bridge arms = 2,500Ω at 50 degrees C. All capacitances in μf. All paper capacitors 600 volts working.

Appendix I

Crystal-Bridge Signal-to-Noise Ratio

The available noise power from the bridge is $P_n = kTB$, where k is Boltzmann's constant, T is the absolute temperature of the bridge, and B is the noise bandwidth of the null detector. Considering an equal-arm bridge, the equivalent bridge input voltage is $V_1 = 2\sqrt{P_R R}$, where P_R is the power dissipated in the resonator, and R is the resistance of each arm. The open-circuit bridge-output voltage resulting from a small fractional detuning $\Delta f/f$ is then $V_2 \cong Q\sqrt{P_R R}\,\Delta f/f$, and the available signal power is $P_S \cong \frac{1}{4}Q^2 P_R\,(\Delta f/f)^2$. It is assumed that the resolution of the system is limited to the condition of equal signal and noise powers, although this could be exceeded by special means. Equating the signal and noise powers, and assuming that making a measurement in t seconds requires a bandwidth $B = 1/t$ cycles,

$$\frac{\Delta f}{f} = \frac{2}{Q}\sqrt{\frac{kT}{P_R t}}\,.$$

This neglects the noise contribution of the null-detection system, which can be neglected because its noise figure can be made very low at medium frequencies.

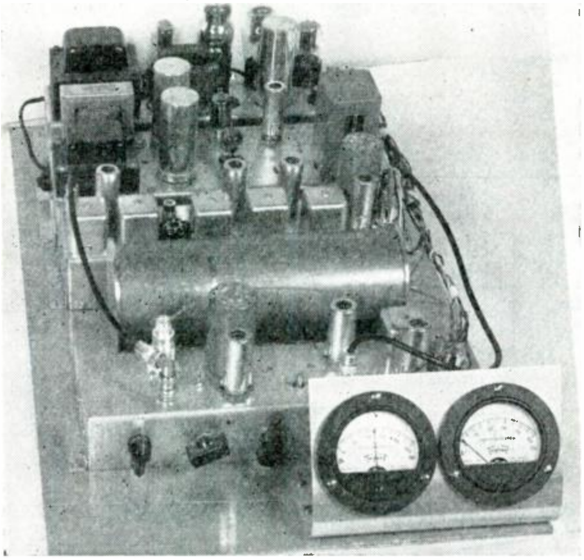

Fig. 7—Photograph of the experimental bridge-balancing oscillator.

Appendix II

Frequency-Control System

The output voltage V_2 of an equal-arm resonance bridge with input V_1 is given by

$$V_2 \cong \tfrac{1}{4}QuV_1 \qquad (5)$$

where

$$u = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega},$$

and ω_0 is the resonant frequency of the resonator.

The output voltage of V_3 of the null-detection system is then

$$V_3 \cong \tfrac{1}{4}QuV_1 A, \qquad (6)$$

where A is the effective voltage gain of the null detector.

A reactance tube is driven by V_3. The change produced by the reactance tube is

$$u_R = K_1 V_3. \qquad (7)$$

If the original (uncorrected) error is u_0, the new (corrected) error is

$$\begin{aligned} u_c &= u_0 - u_R \\ &= u_0 - K_1 V_3 = u_0 - \tfrac{1}{4}K_1 Q u_c V_1 A. \end{aligned} \qquad (8)$$

Solving for u_c/u_0,

$$\frac{u_c}{u_0} = \frac{1}{1 + \frac{1}{4}K_1 QAV_1}\,. \qquad (9)$$

Since

$$u \cong 2\,\frac{\omega_0 - \omega}{\omega_0} = \frac{\Delta\omega}{\omega_0} = \frac{\Delta f}{f},$$

$$\frac{\left(\frac{\Delta f}{f}\right)_c}{\frac{\Delta f}{f}} = \frac{1}{1 + \frac{1}{4}K_1 QAV_1}$$

Appendix III

Temperature-Control Signal-to-Noise Ratio

The available noise power from the bridge is $P_n = kTB$, where B is the noise bandwidth of the system following the bridge, and the other quantities are defined above. If the signal power supplied to the bridge is P_B, the equivalent input voltage V_1 is $V_1 = \sqrt{P_B R}$, when R is the resistance of each arm at the balance temperature. The open-circuit output voltage V_2 is then $V_2 = \frac{1}{2}\alpha\Delta T V_1 = \frac{1}{2}\alpha\Delta T\sqrt{P_B R}$, where α is the temperature coefficient of resistance of the material used in two of the bridge arms, and ΔT is the difference between the balance and operating temperatures. Thus the available signal power P_S is $P_S = \frac{1}{8}\alpha^2 P_B(\Delta T)^2$. Equating P_n and P_S, and letting $B = 1/t$, where t is time required for a measurement,

$$\Delta T = \frac{2\sqrt{2}}{\alpha}\sqrt{\frac{kT}{P_B t}}\,.$$

Appendix IV

Temperature-Control System

T_0 = bridge-balance temperature
T_1 = bridge temperature (very nearly the oven temperature)
T_2 = ambient temperature
Pb = constant signal power supplied to bridge
P_H = variable bridge-heating power
R = resistance of each bridge arm at T_0.

V_2 = bridge-output voltage (alternating)
V_3 = detector-output voltage (direct)
$A = V_3/V_2$
α = temperature-coefficient of resistance of two opposite bridge arms
K_2 = coefficient of power transfer through oven walls
S = ambient-temperature reduction factor.

The temperature-sensing bridge is supplied with a small power P_B at the line frequency. The bridge output voltage is amplified and rectified in a synchronous detector. The detector output, which is a direct voltage proportional to bridge unbalance, is filtered and applied to the bridge to supply the necessary heat, P_H. The principal advantage of rectification is the avoidance of oscillation at the signal frequency. An additional benefit is the fact that, with synchronous detection, the effective bandwidth of the system is twice that of the filter following the detector, rather than that of the audio amplifier preceding the detector. It is easier to obtain a narrow bandwidth with a low-pass filter following the detector than with a band-pass filter at the power line frequency. Actually the effective bandwidth of the system may be somewhat less than this because of the thermal inertia of the oven itself, which could produce some additional filtering.

It is easily shown that the bridge output voltage is

$$V_2 = \frac{\sqrt{P_B R}\,\alpha}{2}(T_0 - T_1) \tag{10}$$

$$V_2 = AV_2 \tag{11}$$

$$\text{power in} = \text{power out} \tag{12}$$

$$P_H = K_2(T_1 - T_2) \tag{13}$$

$$V_3 = \sqrt{P_H R} = \sqrt{RK_2(T_1 - T_2)} \tag{14}$$

Combining (10) and (14).

$$T_0 - T_1 = \frac{2}{A\alpha}\sqrt{\frac{K_2}{P_B}}(T_1 - T_2). \tag{15}$$

Differentiating (15), the ambient-temperature reduction factor S is obtained,

$$S = \frac{d(T_0 - T_1)}{d(T_1 - T_2)} = \frac{1}{A\alpha}\sqrt{\frac{K_2}{P_B(T_1 - T_2)}}. \tag{16}$$

An Inflection-Point Emission Test*

E. G. HOPKINS† AND K. K. SHRIVASTAVA†

Summary—The paper describes a method by which the inflection-point-emission of an oxide-cathode tube may be indicated as a meter reading after subjecting the tube to a single triangular current pulse. The testing current rises from zero at a constant rate and the anode voltage is differentiated twice with respect to time. The time which elapses between the commencement of the test and the appearance of the first positive voltage from the second differentiator is indicated on a meter calibrated directly in inflection-point emission. The testing current is arrested at the inflection point. An experimental equipment is described which tests receiving-tubes for inflection-point emission using a rate of current rise of 4 amperes per millisecond. The duration of the test is of the order of one millisecond and the emission readings tend towards pulse emission values rather than dc emission values. Experiments are described which demonstrate the potential usefulness of the test for quality control in tube manufacture and for taking readings of emission while subjecting a tube to particular operating conditions.

Introduction

THE CURRENT-VOLTAGE characteristic of an oxide-cathode tube executes a gradual increase in slope throughout the space-charge-limited region followed by a gradual decrease during transition to the temperature limited region. "Emission" can be defined only as the current corresponding to some arbitrarily chosen point on this characteristic such as "breakway point," "inflection point," or "flection point." The location of such points has hitherto required the interpretation of a simultaneous display of both current and voltage,[1,2] over a large portion of the tube characteristic. This is too lengthy a procedure for the commercial testing of receiving tubes or for rapid observations of emission during tube operation.

The system of emission testing described in this paper enables inflection-point emission to be read directly on a meter by subjecting the tube under test to a single current pulse. It was developed to enable the emission of a self-heating tube[3] to be measured with the minimum of interruption to the normal electron bombardment of its cathodes and is applicable to the investigation of oxide-cathode behavior under operating conditions and

* Original manuscript received by the IRE, November 8, 1954; revised manuscript February 11, 1955.
† School of Electrical Engrg., N.S.W. University of Technology, Broadway, Sydney, Australia.

[1] Standards of the IRE, Electron Tubes, Part I, "Methods of Testing," Proc. I.R.E., vol. 38 pp. 922–924; August, 1950.
[2] L. A. Marzetta, "High-power pulser aids cathode studies," *Electronics*, vol. 27, pp. 178–180; March, 1954.
[3] E. G. Hopkins, "Self-heating thermionic tubes," *Proc. A.I.E.E.*, Part III, p. 77; March, 1954.

to the high speed testing of receiving tubes. In particular it could improve the emission testing of small high-vacuum rectifiers.

In the new method, a current which increases from zero at a fixed rate is passed through the tube under test. Current is therefore directly proportional to the time which has elapsed since the commencement of the test. The voltage across the tube is differentiated twice with respect to time and the time taken for the resultant voltage to reach zero is indicated on a meter calibrated directly in inflection-point emission current. The current through the tube is interrupted at the inflection point so that the cathode is never subjected to temperature-limited operation. Such operation rapidly impairs emission and its avoidance is therefore important if emission tests are to be made repeatedly to discover the effect on emission of other variables.

Principle of Operation

Consider the anode voltage versus anode current characteristic of a typical oxide cathode diode. In the space-charge-limited region, d^2e_p/di_p^2 which is proportional to $i_p^{-4/3}$ increases from its zero-current value of minus infinity but never reaches zero. With the onset of saturation, d^2e_p/di_p^2 increases more rapidly and passes through zero to become positive at the inflection point. If a current which increases linearly with time ($i_p=Kt$) is passed through the diode and the voltage across it is differentiated twice with respect to time, the resultant voltage is proportional to d^2e_p/di_p^2 and the time at which this voltage reaches zero is I_s/K where I_s is the inflection-point emission current. This time and hence the inflection-point emission current may be measured by a combination of: (1) a current source having a constant known rate of rise from zero, (2) two differentiating circuits followed by an amplifier which will detect the appearance of a small positive second-differential of anode voltage, and (3) a time-measuring circuit which will indicate the time which elapses between the commencement of the test and the appearance of an inflection-point signal from the amplifier. The accuracy of such a system of measurement depends on the precision with which the inflection point may be located. The differentiators and amplifier must be capable of producing without significant delay a noise-free pulse in response to an increment of d^2i_p/dt^2 from zero to a positive value, during an increment of time which is small compared with the time required to reach the inflection point.

Description of the Experimental Testing Equipment

The operation of the experimental testing equipment is illustrated by the block diagram of Fig. 1. The testing current is obtained by discharging a 60 microfarad capacitor previously charged to 2,000 volts through a series combination of a thyratron, a 0.5 henry air-core inductance and the tube under test. The rate of current rise is assumed to be constant at 4 amperes per millisecond over the test period which always occupies less than 10 degrees of the test circuit discharge wave. Actually, the curvature of the current wave is sufficient to cause a significant error through displacement of the inflection point when measuring tubes which exhibit a very gradual transition to the temperature limited region. The voltage across the tube under test (which may suitably rise to between 100 and 200 volts maximum) is differentiated by an RC circuit (NET-1) whose output is amplified by a single pentode amplifier (BST-1) which also reverses the polarity of the signal before passing it on to a second RC differentiator (NET-2). A second amplifier (BST-2) produces a positive output of 100 volts in response to a 2-millivolt negative signal from the second differentiator. The test is initiated by firing the series thyratron and is ended by firing a second thyratron in parallel with the tube under test in response to either an inflection-point signal from the second amplifier or a pulse from a timer which determines the maximum test period.

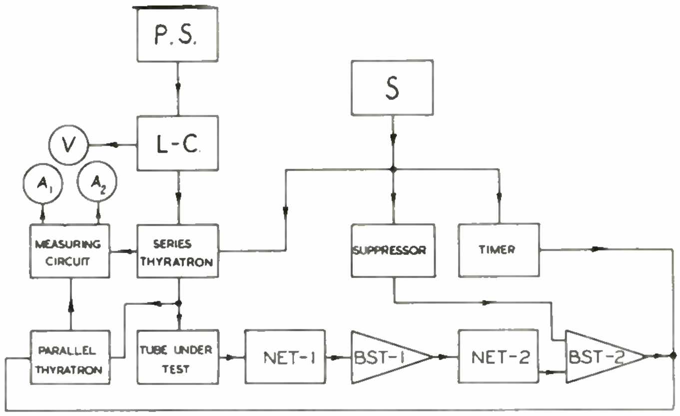

Fig. 1.—Block diagram illustrating the operation of the experimental emission-testing equipment. The meters V, A_1, and A_2 indicate the capacitor voltage, the inflection-point emission and the charging current of the time-measuring capacitor.

The second amplifier consists of two pentodes and a triode and is of conventional design with the exception that series grid resistors and large interstage coupling capacitors are used. This enables the zero reference level to be preserved over a period of several milliseconds in the presence of large signals from the second differentiator. The gain of the second amplifier is constant from ten cycles per second to 50 kilocycles per second and falls by 50 per cent at 100 kilocycles per second.

A suppressor circuit applies a paralyzing voltage to the second amplifier for the first 20 microseconds of the test to render it insensitive to a large negative pulse from the second differentiator which accompanies the firing of the series thyratron.

The time taken to reach the inflection point is measured by charging a capacitor during the period of the test with the constant anode current of a pentode. The capacitor voltage is indicated unchanged for a period of at least 20 seconds after the test by a vacuum-tube voltmeter (A_1) using an inverted triode[4] to attain the necessary low input leakage and having a linear scale reading 0 to 5 amperes.

The equipment is controlled by a three position key switch S. In the upper position, the time-measuring capacitor is short-circuited and the pentode bias removed to allow checking of the zero setting of the vacuum-tube voltmeter and the pentode anode current (A_2). The timer and suppressor circuits are also reset in this position. As the switch is depressed through the center position all circuits are returned to their operating condition; the lower position a single pair of contacts simultaneously applies a triggering pulse to the series thyratron and the suppressor and timer circuits.

The rate of current rise of 4 amperes per millisecond was chosen for the experimental equipment to facilitate observation of single oscilloscope sweeps and to enable relatively simple thyratron and amplifier circuits to be used. In any future designs it would be advisable to abandon the present method of current-control in favor of a high-vacuum tetrode circuit of the type used for radar modulation.[5] Such a modification would allow greater utilization of storage capacitance with less curvature in the current waveform and would also make possible the use of higher rates of current rise leading to the possibility of testing larger cathodes. A higher rate of current rise might be desirable to bring about a reduction in the heating of the tube under test and so to permit more frequent readings of rapidly changing emission to be taken.

Performance

Commercial receiving tubes measured at their normal operating heater voltages by the experimental equipment give inflection-point emission densities of 2 to 8 amperes/cm². These values are greater by a large factor than the dc emission usually obtained with commercial tubes, indicating that in the present method the cathodes are probably not greatly affected by the passage of the testing current.[6] Readings on a particular tube may be repeated indefinitely with variations of less than 2 per cent, provided that the time between tests is greater than 10 seconds.

The oscillograms of Fig. 2 show the anode voltage of the tube under test and the amplifier output as functions of time for two values of heater voltage when testing a Type 6X4 tube. To obtain these oscillograms, the amplifier output was disconnected from the parallel thyratron which was fired by the timer. Although the change in curvature is apparently very gradual at the higher filament voltage, a satisfactory inflection-point signal is produced by the amplifier. However, in most

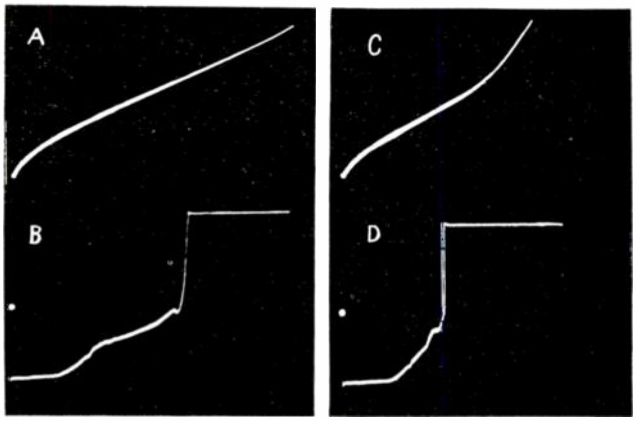

Fig. 2.—Oscillograms A and C of anode voltage verses time of a Type 6X4 tube undergoing test, with corresponding oscillograms B and D of amplifier output. Oscillograms A and B were made with a higher filament voltage than those of C and D. In obtaining these oscillograms the duration of the test pulse was set by means of the timer at 1.25 milliseconds for A and 0.9 milliseconds for C to prevent damage to the tube under test. In normal operation the test pulse is terminated at the inflection point.

of the tubes tested over a wide range of filament voltage, the inflection point signal decreased to zero above a certain critical filament voltage, which in a few tubes was as low as 6.3 volts, indicating that the negative curvature of the tube characteristic was, at the most, insufficient to cancel the small opposing curvature of the current pulse.

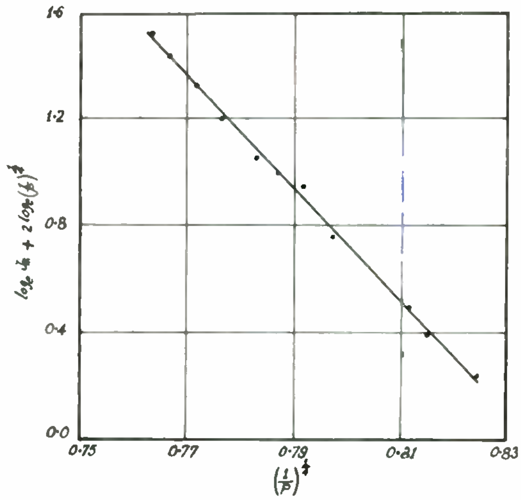

Fig. 3—The values of emission amps and heater power watts refer to one square centimeter of cathode area.

Fig. 3 is a plot of

$$\log_e (\text{emission}) + 2 \log_e \left(\frac{1}{\text{heater power}}\right)^{1/4}$$

against

$$\left(\frac{1}{\text{heater power}}\right)^{1/4}$$

for one of a batch of Type 6X5-GT tubes tested over the range of filament voltage 5.4 volts of 6.6 volts. If it is assumed that the cathode work function and thermal

[4] F. E. Terman, "The inverted vacuum tube—a voltage-reducing power amplifier," Proc. I.R.E., vol. 16, p. 477; April, 1928.
[5] G. N. Glasoe and J. V. Lebancqz, "Pulse Generators," McGraw-Hill Book Co. Inc., New York, Ch. 3; 1948.
[6] P. A. Wright, "A Survey of present knowledge of thermionic emitters," *Proc. A.I.E.E.*, Part III, vol. 100, p. 125; May, 1953.

emissivity remain constant over this range, the experimental points might be expected to lie on a straight line.[7] A straight line may be fitted to the points so that the maximum deviation from it corresponds to an error in emission-reading of 5 per cent. The other tubes of the batch either followed a similar linear law or smoothly departed from linearity in the direction of reduced emission above a certain heater power.

The equipment was used to measure the variations in inflection-point emission which occur during the testing of Type 6X4 tubes by the conventional 50-volt 3-second method.[8] A mechanical relay was arranged to disconnect the tube from the 50-volt circuit one millisecond prior to the firing of the series thyratron. A number of separate tests were made on each tube at various intervals up to 4 seconds after the application of 50 volts dc to the anode, the tube being allowed to recover completely its initial emission between tests. From these tests it was shown that the inflection-point emission of a typical tube fell to 45 per cent of its initial value within the first 0.5 second of the 50-volt test, rose between 0.7 second and 2 seconds to 100 per cent of the initial value and then fell steadily to less than 15 per cent at 4 seconds. This type of variation was observed in all the tubes and is thought to be due to the counteracting effects of the poisoning of the cathode surface and an increase in cathode temperature. The above experiment demonstrates the ability of the new test to sample the emission of a tube rapidly while it is operating under a particular set of conditions. By the use of suitable mechanical or electronic switching, standard production tests might be arranged to sample emission under any specified operating conditions or to observe the relative effects on emission of currents drawn to each electrode of a multi-electrode tube.

A batch of Type 6X4 tubes known to be slightly gassy and having previously shown signs of sparking during a warming test contained some tubes having particularly high inflection-point emission readings. Oscillograms showed that the anode voltages of these tubes at any current were below normal and in some cases anode voltage tended to become independent of anode current in the region 3 amperes to 4 amperes. The fact that the presence of gas may increase the apparent inflection-point emission could prove to be a serious weakness in any attempt to rely solely on this or any other emission test as a measure of tube quality.

The relatively large inter-electrode spacings of rectifiers such as Types 6X4 and 6X5-GT preclude their being tested in the temperature-limited region by simple dc tests, because of the excessive anode heating which would result. The 50-volt 3-second test referred to above is at present in widespread use for the commercial and governmental acceptance testing of these tube types. Although referred to as an "emission test," it is recognized by the tube manufacturers to be a test of perveance which eliminates very poor tubes but which provides little information on the effects of changes in manufacturing procedures on cathode quality. The following experiment demonstrates the potential superiority of the new test as an indication of cathode quality. One hundred Type 6X4 rectifier tubes taken from one production batch were tested both with the new equipment and the conventional 50-volt 3-second emission check. The distributions of the readings of both tests are illustrated by frequency diagrams in Fig. 4. A scatter diagram failed to indicate any noticeable relationship between the two sets of readings. The fact that the equipment always functioned was checked by comparing the inflection-point emission reading of each of the hundred tubes with an oscillogram similar to those of Fig. 2 taken immediately after the emission reading.

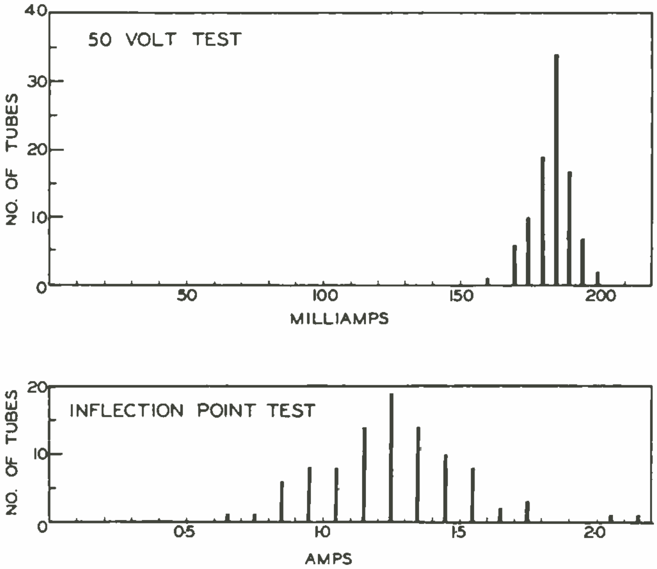

Fig. 4—Frequency diagrams illustrating the distribution of inflection-point emission readings and 50-volt emission readings among a batch of one hundred Type 6X4 tubes.

A number of types of multi-electrode tubes were tested with the equipment, using the diode connection. In the case of tubes having very high perveance it was necessary to connect a series resistor of 500 to 1,000 ohms between the control grid and the other electrodes in order to produce sufficiently high anode voltages. While this procedure upset the mathematical correctness of the test, good comparative emission readings could still be obtained. The inability of the experimental equipment to test a high perveance tube is due to the fact that it was designed specifically for rectifiers and does not indicate an inherent weakness of the method.

Conclusion

The system described in this paper enables a rapid reading of inflection-point emission to be made without significantly heating the tube under test. Although it

[7] W. G. Dow, "Fundamentals of Engineering Electronics," (John Wiley and Sons Inc., New York, N.Y. pp. 159–160; 1937.

[8] F. Langford-Smith, "Radiotron Designers Handbook," Amalgamated Wireless Valve Co. Pty. Ltd., Sydney, Australia (published in U.S.A. by R.C.A.), pp. 94–95; 1952.

cannot be assumed that all commercial tubes exhibit a definite inflection point at the normal operating cathode temperature, an instrument based on the system would probably be of use to tube manufacturers for rapid factory or laboratory measurements and investigating variations of emission occurring during tube operation.

Acknowledgment

The project described above was partially supported by the Technical Co-operation Scheme of the Colombo Plan. The authors wish to thank Professor C. J. Milner and Mr. J. B. Douglas of this University for their advice during the preparation of the paper and the former for some valuable discussions during the course of the work. The authors are indebted to Professor R. E. Vowels, Head of the School of Electrical Engineering of this University for his interest and to Amalgamated Wireless Valve Company Pty. Ltd. for their co-operation in testing the experimental equipment.

X-Ray Emission from High-Voltage Hydrogen Thyratrons*

S. SCHNEIDER† AND B. REICH†

Summary—**In the operation of high-voltage hydrogen thyratrons, the bombardment of the anode by high-energy electrons causes the production of X-rays. The X-ray output of three types of high-voltage hydrogen thyratrons has been studied to determine the generation of X-rays as a function of the operating parameters. The beam pattern and the energy and intensity of the emitted X-rays are included. It has been determined that emission occurs during the breakdown of the pulse and during the interpulse interval as a result of grid emission. The beam pattern is defined by the geometry of the tube and is a narrow beam emanating in a circle from the grid-anode region. Values for the X-ray energy and intensity and for protective shielding are given.**

Introduction

THE BOMBARDMENT of a substance by cathode rays causes the production of X-rays. In the operation of high-voltage electron tubes, this action is present where electrons emitted from the grid or cathode are accelerated by an electric field toward the anode.

At the Signal Corps Engineering Laboratories, Fort Monmouth, New Jersey, high-voltage thyratrons are being tested. Personnel working with these tubes were therefore issued film badges to detect any X-radiation exposure. The results for the most part were negative. However, further examination showed that due to the physical configuration of the testing apparatus X-radiation was being emitted from these tubes at head level, and therefore never intercepted the film badges at chest level. A study was conducted to determine the radiation levels and characteristics of the X-rays emitted from the thyratrons.

Operation of Thyratrons

The hydrogen thyratron, as shown in Fig. 1, consists of an indirectly heated cylindrical cathode with an inner and outer cathode shield and cathode baffle. The nickel grid structure consists of a cylindrical portion and a perforated grid disk with a solid grid baffle below it. The molybdenum anode is enclosed by the grid disk, the

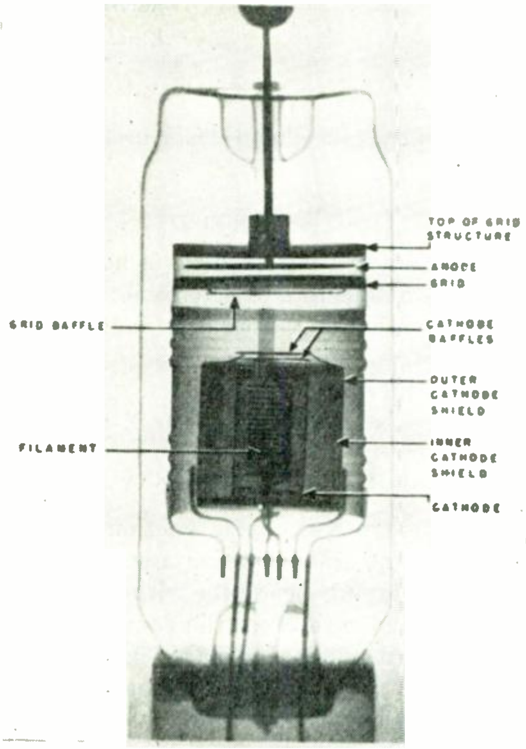

Fig. 1—Hydrogen thyratron.

grid mesh, and the top of the grid structure, with the spacing between anode and grid less than one mean free electron path at a pressure of 500 microns mercury. The tube is operated in a line-type pulse-modulator circuit with resonant charging of the pulse-forming network to

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† Sig. Corps Engrg. Labs., Fort Monmouth, N.J.

a high peak forward voltage. A charging diode is used to maintain this voltage on the anode until the tube is fired. Fig. 2 shows voltage and current waveforms.

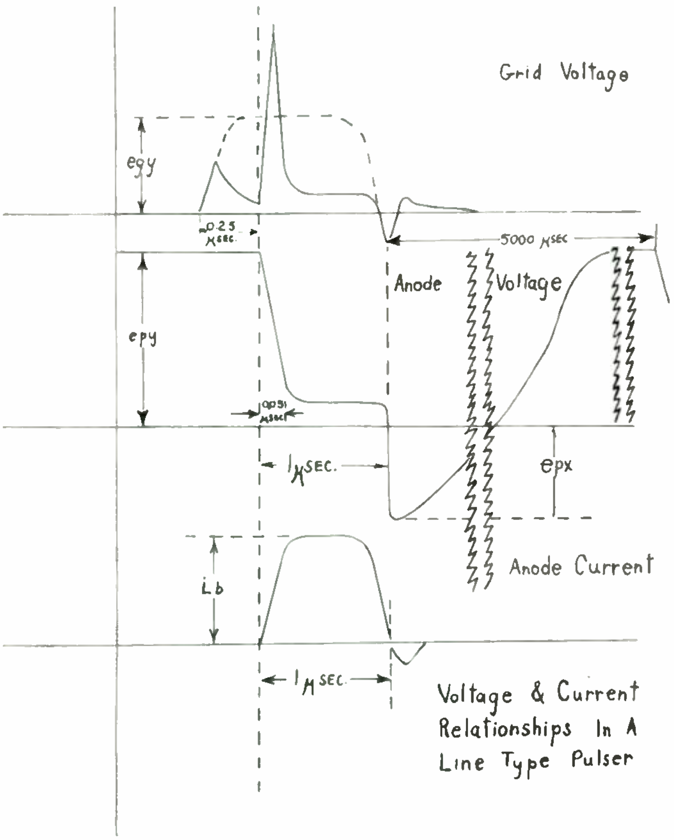


Fig. 2—Voltage and current relationships in a line-type pulser.

X-Ray Output

The X-ray output was studied using photographic and scintillation detectors to determine:

1. The generation of X-rays as a function of the operating parameters,
2. The X-ray beam pattern,
3. The energy and intensity of the X-rays emitted.

Generation of X-Rays

The generation of X-rays as a function of tube operating conditions was studied, using a scintillation detector. A 5819 photomultiplier tube with a sodium iodide, thallium-activated crystal for the conversion of the X-ray pulses to light pulses was placed in the beam of the X-rays opposite the grid-anode region. The output of the photomultiplier was amplified and observed on a synchroscope. The charging cycle of the plate voltage applied to the thyratron was fed into the scope simultaneously. Fig. 3 shows the observed pattern of the radiation output as negative pulses superimposed on the charging cycle. This photograph was taken of a hydrogen thyratron type 1257 operating at 38 kilovolts peak forward voltage, 2,000 amperes peak forward current, with a 2.5 microsecond current pulse at a repetition rate of 200 pulses per second. An analysis of the photograph shows at time zero, which corresponds to the initiation of the discharge, there is a sharp negative pulse of X-rays. This is due to a considerable current being drawn to the anode before the anode voltage has dropped below the level where the X-rays can be detected. During the pulse there are no X-rays emitted since the anode potential is only several hundred volts. Immediately

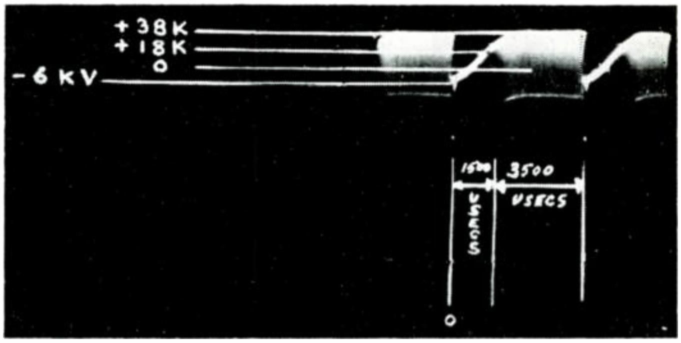


Fig. 3—Observed pattern (scintillation detector).

after the pulse, the plate voltage becomes negative, 6 kilovolts, as a result of the mismatch between the load and the pulse forming networks. After this the pulse forming network recharges. Fifteen-hundred microseconds after the pulse, the plate voltage has risen to 18 kilovolts and X-ray pulses are again observed from the tube and this continues during the entire interpulse interval until the tube fires again. The X-ray output during the period when the tube is not conducting is apparently due to thermionic emission from the grid as a result of migration of barium from the cathode during tube operation. It has been observed that tubes with excessive grid emission will generate considerable more X-rays than normal tubes. No measurement to completely correlate X-ray with grid emission has been made.

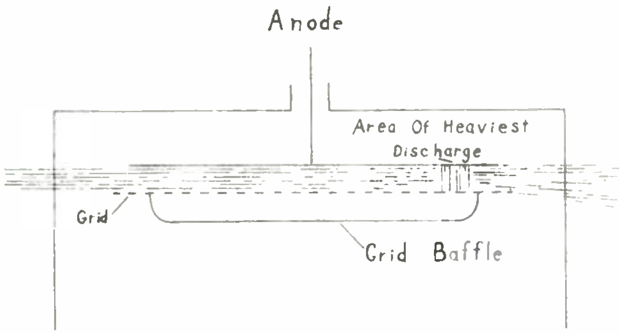


Fig. 4—Tube discharge and X-ray output.

Beam Configuration

Since the generation of X-rays occurs in the grid anode region by bombardment of the anode with high-energy electrons and since there is considerable shielding in the tube due to the massive grid structure, the major portion of the X-ray beam should emanate in a circle through the screen mesh of the grid-anode region, as shown in Fig. 4. The main area of generation of X-rays should be at the point where the discharge is the heaviest for the following two reasons:

1. During the breakdown period of the tube, most of the current is being drawn at this point,
2. Heavy localized heating in the area of heaviest discharge causes the greatest amount of grid emission during the interpulse interval.

Fig. 5—X-ray pattern (photographic).

Fig. 5 shows a typical X-ray pattern taken at one foot from the center of the tube (a type 1257). On the side where the discharge is heaviest, the beam is $\frac{5}{8}$ inch wide. The opposite side is only $\frac{3}{16}$ inch. X-ray output was noted at other points, which were less intense than the main beam.

X-Ray Energy and Intensity

In energy and intensity measurements, dental films, Minimax Type EFS without lead, were used.

Energy Measurements

The films were set at an arbitrary distance from the center of the thyratron. The initial measurement was made without an external absorber, the remaining measurements being made with a series of calibrated aluminum absorbers. In this manner the data for an absorption curve was gathered. Fig. 6 is a graph of the absorption curves of one each of three hydrogen thyratrons types 1257, 5948/1754, 5949/1907. It is noted that the curves plotted on semilogarithm paper are straight lines, indicating that the half-value layers of aluminum are constant. On this basis the exponential absorption law will be applied as follows:

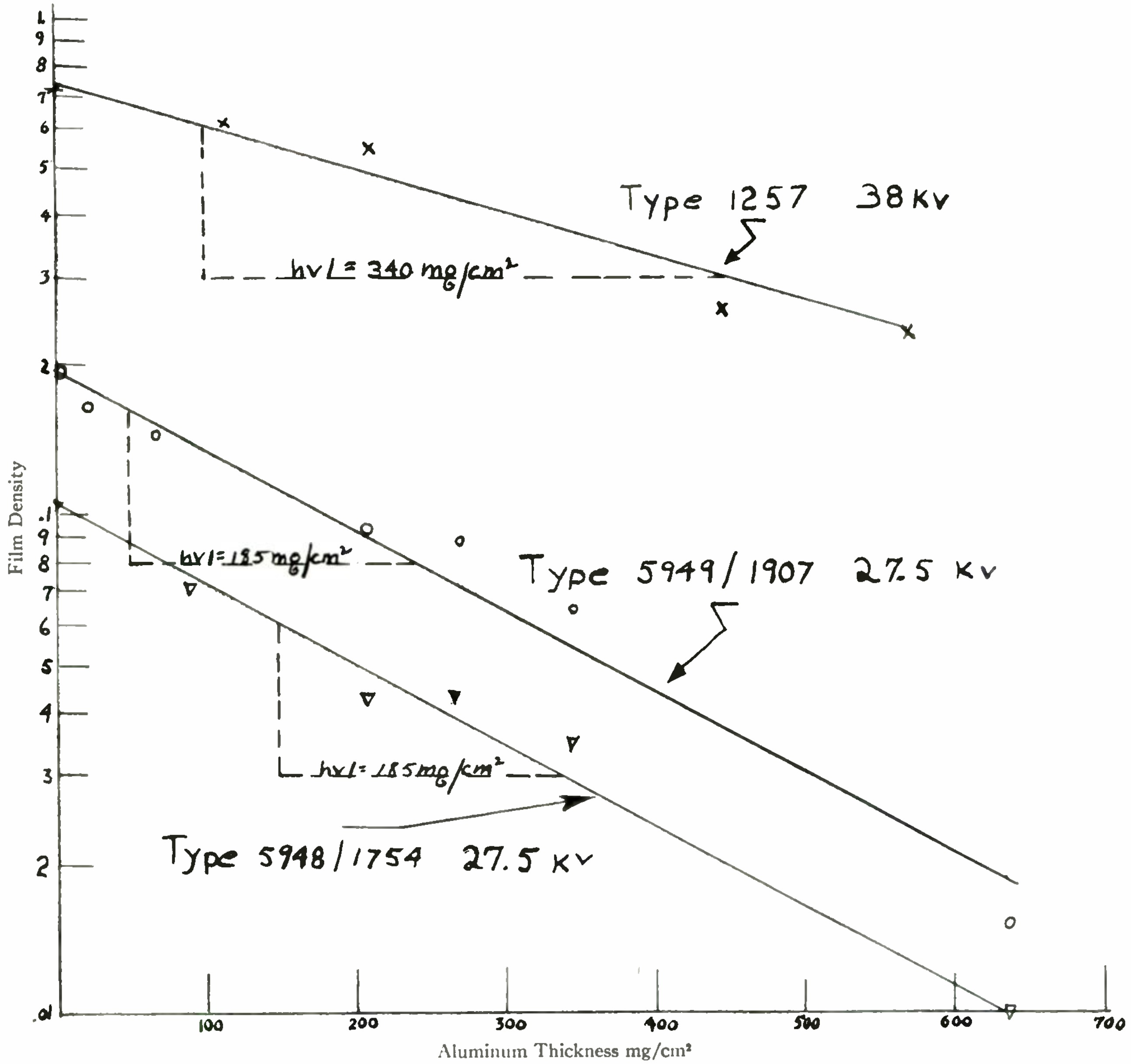


Fig. 6—Absorption curves for thyratron types 1257, 5948/1754, 5949/1907.

$$I = I_0 e - \frac{u}{p} X,$$

where

I_0 = initial intensity, no external absorber

I = final intensity with absorber

u/p = mass absorption coefficient (cm²/g)

X = thickness of absorber (g/cm²)

With this assumption of the exponential law, the following sample calculation will be made. Considering the type 1257 hydrogen thyratron, the following information is available from Fig. 6.

$$\frac{I}{I_0} = 0.5 \text{ when } X = 0.340 \text{ g/cm}^2$$

therefore

$$0.5 = e^{-u/p(0.340)}$$

$$2.04 \text{ cm}^2/\text{g} = u/p.$$

Consulting a table of mass absorption coefficients the effective energy of the radiating beam is found to be 25 kilovolts.[1] Similarly, the above calculations were applied to the other types. Table I is a summary of this data.

TABLE I

EFFECTIVE ENERGIES

Tube type	Peak operating voltage (kv)	hvl of AL (g/cm²)	Effective energy (kev)
1257	38	0.340	25
5948/1754	27.5	0.185	21
5949/1907	27.5	0.185	21

Intensity Measurements

From the above information the dental films used to monitor the X-ray beam were calibrated. The reason for this calibration is that the films used are extremely energy-dependent in this region and that in order to ascertain the dosage properly this calibration is necessary. A constant potential X-ray machine, General Electric Model XRD-1, was used. Table II is a listing of the radiation levels of one each of the three types of thyratrons at various distances as noted.

TABLE II

RADIATION LEVELS

Tube type	Peak voltage	Distance	Radiation intensity
	kilovolts	feet	milliroentgens per hour
1257	42	1	1230
	42	2	360
	38	1	570
	38	2	120
5948/1754	27.5	1	30
5949/1907	27.5	1	30

From the limited data in Table II, it is noted that the inverse square law seems to be valid for this experimental setup in the case of the thyratron type 1257.

[1] G. R. White, "X-Ray Attenuation Coefficients from 10 Kev to 100 Mev," *National Bureau of Standards Report* 1,003, May 13, 1952.

There may be considerable variation in X-ray emission from tube to tube under similar operating conditions due to variations in grid emission. It has been observed at other laboratories[2] that tubes may emit considerably more X-radiation than that measured during these experiments. The greatest X-ray emission observed was in the case of a 1257 which could not operate for any long period of time due to excessive grid emission. In this case the rate of emission was 10,000 milliroentgens per hour.

Protective Shielding

In determining personnel exposure, it should be noted that there are important factors to be considered such as exposure distance, time, and protective shielding. In addition to these it should be pointed out that the portion of the body that is irradiated and the area of exposure are important in determining any radiation damage.

As stated in the earlier portions of the report, the radiation beam is relatively narrow. In the operational set up at these Laboratories, the beam intercepts the upper portion of the body, in some cases the eyes.

In computing radiation intensity the exponential law is used and in addition suitable and convenient materials should be chosen when necessary.

A sample calculation will now be made to indicate the method of determining radiation intensity.

From Table II the radiation intensity at one foot from the center of the tube operating at 38 kilovolts is 570 milliroentgens per hour. Assuming that it was decided to reduce the level from 570 to 7.5 milliroentgens per hour, then

$$\frac{I}{I_0} = \frac{7.5}{570} = 0.0132.$$

For aluminum at 25 kev

$$u/p = 1.8 \text{ cm}^2/\text{g},$$

therefore

$$X = \frac{4.3}{1.8} = 2.4 \text{ g/cm}^2.$$

The density of aluminum is

$$p = 2.7 \text{ gm/cm}^3,$$

and the thickness

$$t = \frac{X}{p} = \frac{2.4}{2.7} = 0.89 \text{ cm}.$$

Therefore to reduce the radiation intensity from 570 to 7.5 milliroentgens per hour would require 0.89 centimeters of aluminum.

[2] M. H. Shamos, Chatham Electronics, Livingston, New Jersey; private communication.

Conclusions

It has been observed that X-rays are emitted from high-voltage hydrogen thyratrons. The X-Rays are intimately related to the mode of operation, the geometry of the tube, and the grid emission developed by migration of barium from the oxide-coated cathode to various parts of the tube. There may be considerable variation in the intensity of the X-ray beam with variations in grid emission, which is the dominant factor in determining intensity. Based on the most serious case of grid emission (10,000 milliroentgens per hour at one foot with an effective energy of 25 kilovolts), these Laboratories have established a minimum shielding requirement for the protection of personnel testing these tubes.

The shielding is as follows.

1. For viewing: $\frac{1}{4}$-inch lead glass which will attenuate 10,000 milliroentgens per hour to a negligible quantity.

2. For paneling: $\frac{1}{16}$-inch steel housing which will attenuate 10,000 milliroentgens per hour to 0.45 milliroentgens per hour. In case of inadequate shielding at the corners, 10 mil lead foil which has the same attenuation as the steel should be used.

Analysis of a Broadband Detector Circuit*

M. H. BROCKMAN†, SENIOR MEMBER, IRE

Summary—This paper presents a theoretical analysis of the steady-state behavior of an ideal linear detector with a broadband output circuit. Published analyses of a diode feeding a load resistance R_1 shunted by a capacitance C_1 show that detection efficiency D to the carrier frequency is a function of the ratio of R_1 to the reactance of C_1, or of R_1C_1 to $1/f$, at the frequency of the applied signal, as well as of the ratio of R_1 to the diode conduction resistance R_d. The broadband detector circuit considered here extends the range, at low values of R_1/X_{C_1}, over which D is maintained practically constant for values of R_1/R_d less than ten. This constant carrier detection efficiency characteristic is realized by inserting the proper size inductance in the detector output circuit.

The solution of the steady-state behavior of the broadband detector circuit is obtained by the method of successive approximations. As a first approximation, carrier detection efficiency is calculated for the case of an assumed infinite inductance in the output circuit. To obtain a practical solution, a second approximation is made for the case of a finite inductance in the detector output circuit. This solution is compared with the solution obtained with the assumed infinite inductance to obtain a safe minimum inductance value for any broadband detector design. Calculated data are presented for four values of the ratio R_1/R_d.

Introduction

ANALYSES have been made of a diode detector feeding a load resistance R_1 shunted by a capacitance C_1. The analysis of such a second detector circuit is greatly simplified if it is assumed that (1) the current-voltage characteristics of the diode is a straight line, and (2) the capacitance shunting the detector load resistance is either negligible or infinite. Various analyses have been made where condition (1) has been allowed but not condition (2). These analyses[1-3] have shown that detection efficiency D to the carrier frequency does vary with carrier frequency for a given ratio of R_1 to diode conduction resistance R_d for intermediate values of capacitance C_1 or intermediate values of R_1/X_{C_1}.[1,4]

The steady-state behavior of an ideal linear detector with a broadband output circuit is here analyzed theoretically. The analysis is important because the output circuit extends the region of constant carrier detection efficiency (realized at higher values of R_1/X_{C_1}) to considerably lower values of R_1/X_{C_1} (or carrier frequency) than for an rc output circuit with the same value of load resistance and capacitance. This characteristic, which is realized for values of R_1/R_d less than 10, is obtained by inserting the proper size inductance in the detector output circuit.

Results

The results of this analysis are shown in Fig. 1. This figure presents, for the case of an ideal linear detector, a comparison of the carrier detection efficiency D of the broadband detector output circuit with that of the standard rc detector output circuit for values of R_1/R_d $=2.5$ and 5.0. (Equations used to obtain the detector conduction angles wt_1 and wt_2 by graphical means for the rc case and to calculate D from wt_1, wt_2, and $K=R_1/X_{C_1}$ are contained in the literature.[5]

The lower curves of calculated carrier detection efficiency D for both the broadband and the rc detector

* Original manuscript received by the IRE, October 18, 1954; revised manuscript received, February 24, 1955. The material in this paper was originally presented as a Master's thesis in June 1953 and the basic data and conclusions are taken from the author's dissertation submitted to the Polytechnic Institute of Brooklyn in partial fulfillment of the requirements for the degree of Master of Electrical Engineering.

† Airborne Instruments Lab., Inc., Mineola, N. Y.

[1] Jean Marique, "Notes on the theory of diode rectification," *Wireless Eng.*, vol. 12, p. 17; January, 1935.

[2] K. R. Sturley, "Radio Receiver Design," vol. 1, John Wiley and Sons, Inc. New York; 1943.

[3] W. B. Lewis, "The detector," *Wireless Eng.*, vol. 9, p. 487; September, 1932.

[4] K. R. Sturley, *loc. cit.*, Fig. 8.13(a).

[5] K. R. Sturley, *loc. cit.*, pp. 367–369.

output circuits shown in Fig. 1 are plotted as a function of K (designation for R_1/X_{C_1}) for R_1/R_d equal to 2.5. At $K=4.67$, the calculated D for the broadband detector is 37 per cent and for the rc detector D is 36 per cent. At $K=2.59$, the calculated D for the broadband detector is 35 per cent while for the rc detector D is 30 per cent. In fact, carrier detection efficiency drops only to 33.5 per cent at $K=2.0$ (see Fig. 4) for the broadband detector circuit. As an example, if $R_1=820$ ohms and $C_1=10\ \mu\mu$f, values of K of 2.0 and 4.67 correspond to carrier signal frequencies of 38.8 and 90.5 mc respectively. Fig. 1 shows a similar effect on carrier detection efficiency D for the broadband and rc detector output circuits for R_1/R_d equal to 5.0. In this case, the curve for the rc detector output circuit was copied from reference 2 for purposes of comparison.

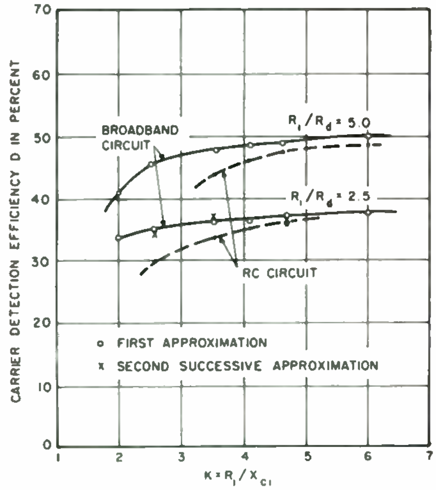
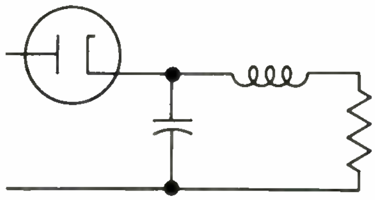


Fig. 1

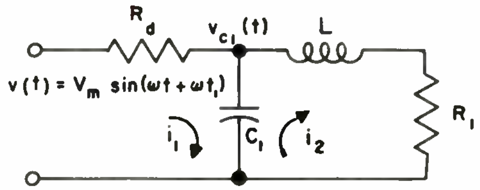
The broadband detector output circuit of Fig. 2(a) has been used as the output circuit of second detectors designed for operation with broadband high-frequency amplifiers. Whereas, to obtain a wide video band in the detector output circuit, the resistance and capacitance of the usual detector rc output circuit must be so chosen that detection efficiency varies with the frequency of the applied carrier. The broadband output circuit tends to eliminate this bandwidth restriction by maintaining constant detection efficiency within the amplifier pass band.

Equivalent Circuit

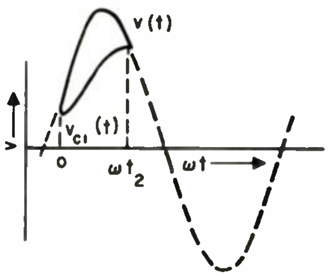
The equivalent circuit of the broadband detector circuit during capacitor charge (diode conducting) is shown in Fig. 2(b); the equivalent detector circuit during capacitor discharge (diode nonconducting) is shown in Fig. 2(d). The behavior of this circuit is analyzed for a sinusoidal signal input by examining the relationship of the dc component of the output voltage across detector capacitance C_1 to the peak voltage of the applied signal as the signal frequency varies. The voltage waveforms during capacitor charge and discharge are shown in Figs. 2(c) and 2(e) respectively.



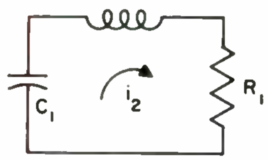
(a) BROAD-BAND DETECTOR CIRCUIT



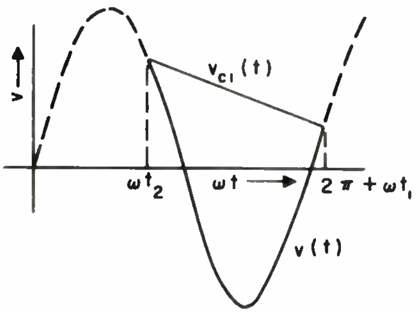
(b) EQUIVALENT BROAD-BAND DETECTOR CIRCUIT DURING CAPACITOR CHARGE



(c) VOLTAGE FOR $0 \leq t \leq (t_2 - t_1)$



(d) EQUIVALENT BROAD-BAND DETECTOR CIRCUIT DURING CAPACITOR DISCHARGE



(e) VOLTAGE FOR $t_2 \leq t \leq (\frac{2\pi}{\omega} + t_1)$

Fig. 2

The applied signal, which is supplied from a constant voltage source, is

$$V(t) = V_m \sin (wt + wt_1). \tag{1}$$

It is assumed that the diode is an ideal linear device whose resistance R_d is constant during conduction and infinite during nonconduction (detector capacitor discharge). In addition, it is assumed that there is no capacitance across the diode. The constants in the detector output are designated as R_1, C_1 and L.

A general solution of the equivalent circuits during charge and discharge can be made separately to obtain expressions for the voltage across the detector capacitance as a function of time. However, the expressions are complicated and their simultaneous solution to determine the conduction period of the diode during steady-state conditions is not particularly attractive. Simplifying assumptions are made therefore and a solution is evolved by successive approximations.

As a first approximation, it is assumed that L is infinite and therefore i_2 is a constant direct current i_{20}. The solution obtained from this first approximation is used to examine the behavior of the broadband detector circuit. As a second approximation, it is assumed that L is finite and that i_2 consists of two terms, a constant direct current i_{20} and an ac component at the applied signal frequency. The behavior of the broadband detector circuit for this second approximation is compared with the behavior for the first approximation to obtain a safe minimum value for the inductance L.

First Successive Approximation

The applied voltage to the equivalent detector circuit during capacitor charge (diode conducting) is given in (1) [see Fig. 2(b)]. The current i_2 is constant at all times and is designated as i_{20}. The circuit equation for the voltage across detector capacitor C_1 becomes (during charge) from inspection

$$V_{C1}(t) = \frac{q_0}{C_1} - \frac{1}{C_1}\int_0^t i_2(t)dt + \frac{1}{C_1}\int_0^t i_1(t)dt. \quad (2)$$

This expression is defined for $0 \leqq t \leqq (t_2 - t_1)$. Initially at $t=0$ (the beginning of charge),

$$\frac{q_0}{C_1} = V_m \sin \omega t_1. \quad (3)$$

The circuit equation for voltage across C_1 during discharge (diode nonconducting) can again be written, from inspection of Fig. 2(d),

$$V_{C1}(t) = \frac{q_0}{C_1} - \frac{i_{20}}{C_1}(t - t_2). \quad (4)$$

This expression is defined for $t_2 \leqq t \leqq (t_1 + 2\pi/\omega)$. Initially, at $t = t_2$ (the beginning of discharge),

$$\frac{q_0}{C_1} = V_m \sin \omega t_2. \quad (5)$$

The solution of the two circuit equations above for steady-state conditions is outlined in Appendix I. Solution of (2) results in the following expression:

$$\left[\sin\phi\cos(\omega t_1 - \phi) + \frac{i_{20}R_1}{V_m}\frac{R_d}{R_1}\right]\epsilon^{\omega t_1/\tan\phi} = \left[\sin\phi\cos(\omega t_2 - \phi) + \frac{i_{20}R_1}{V_m}\frac{R_d}{R_1}\right]\epsilon^{\omega t_2/\tan\phi}, \quad (6)$$

where $\phi = \tan^{-1}(R_d/X_{C_1})$.

This is the equation during charge for steady-state conditions. Solution of (4) results in

$$\sin(\omega t_1) + \frac{i_{20}R_1}{V_m}\frac{1/\omega C_1}{R_1}(2\pi + \omega t_1) = \sin(\omega t_2) + \frac{i_{20}R_1}{V_m}\frac{1/\omega C_1}{R_1}(\omega t_2). \quad (7)$$

This is the equation during discharge for steady-state conditions.

Examination of (6) and (7) reveals that there are three unknowns, ωt_1 (the angle at which detector capacitor discharge ends and capacitor charge-diode conduction starts), ωt_2 (the angle at which diode conduction stops and capacitor discharge starts), and $(i_{20}R_1)/V_m$, which is the detection efficiency D. Now if the assumption is made that the average rectified voltage across C_1 for the charge period is the same as for the discharge period, the value of this average rectified voltage can be written from inspection of Fig. 2(e)

$$D = \frac{i_{20}R_1}{V_m} = \frac{\sin\omega t_1 + \sin\omega t_2}{2}. \quad (8)$$

After substituting (8) into (6) and collecting terms, we obtain

$$\left[\frac{\sin\omega t_1}{2}\left(1 + \frac{R_d}{R_1}\right) + \frac{\sin(2\phi - \omega t_1)}{2} + \frac{\sin\omega t_2}{2}\frac{R_d}{R_1}\right]\epsilon^{\omega t_1/\tan\phi} = \left[\frac{\sin\omega t_2}{2}\left(1 + \frac{R_d}{R_1}\right) + \frac{\sin(2\phi - \omega t_2)}{2} + \frac{\sin\omega t_1}{2}\frac{R_d}{R_1}\right]\epsilon^{\omega t_2/\tan\phi}. \quad (9)$$

This equation during charge for steady-state conditions contains only two unknowns, ωt_1 and ωt_2. Substituting (8) into (7) and collecting terms results in the expression

$$\sin(\omega t_1)\left[1 + \frac{1}{2}\frac{1/\omega C_1}{R_1}(2\pi + \omega t_1)\right] + \frac{\sin(\omega t_2)}{2}\frac{1/\omega C_1}{R_1}(\omega t_1) = \sin(\omega t_2)\left[1 + \frac{1}{2}\frac{1/\omega C_1}{R_1}(\omega t_2 - 2\pi)\right] + \frac{\sin(\omega t_1)}{2}\frac{1/\omega C_1}{R_1}(\omega t_2). \quad (10)$$

This equation during discharge for steady-state conditions also contains only two unknowns, ωt_1 and ωt_2.

Since (9) and (10) contain transcendental functions, the variables ωt_1 and ωt_2 cannot be separated and a solution cannot be obtained directly. Eqs. (9) and (10) are solved graphically in the following manner.

For a given $K=R_1/X_{C_1}$ and R_1/R_d, the left-hand side of expression (9) is plotted as a function of ωt, for an assumed value of ωt_2 and the right-hand side is plotted as a function of ωt_2 for an assumed value of ωt_1. Eq. (10) is plotted in a similar manner. From these plotted curves, values of ωt_1 and ωt_2 are obtained which conditionally satisfy (9) and (10). Now if these values ωt_1 and ωt_2 are plotted one against the other, they will provide continuous curves which are in general as shown in Fig. 3.

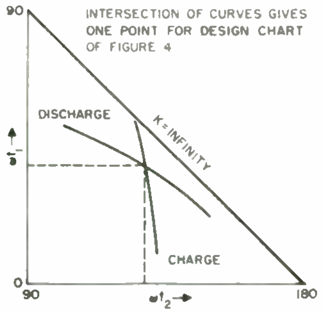


Fig. 3

The point of intersection of these two curves provides the only value of ωt_1 and ωt_2 which satisfies (9) and (10) simultaneously. The value of ωt_2 and ωt_1 at this intersection should agree with the assumed value of ωt_2 for the left-hand side of expressions (9) and (10) and the assumed value of ωt_1 for the right-hand side of expressions (9) and (10) used to obtain the two curves. If they do not agree, the procedure outlined above is repeated assuming a different value of ωt_2 and a different value of ωt_1 until agreement is reached. If this process is repeated for various values of K (K varies with frequency) keeping R_1/R_d fixed, a series of plots are obtained which provide values of ωt_1 and ωt_2 which satisfy steady-state conditions. Such a series of plots are shown in Fig. 4 for the first successive approximation for a few specific R_1/R_d ratios.

Second Successive Approximation

As a second approximation, consider that i_2 consists of a constant dc term and a constant ac term. This is equivalent to assuming that L is finite and that higher order terms in i_2 are ignored:

$$i_2 = i_{20} + i_{2ac} \sin(\omega t + \psi). \tag{11}$$

The expression for the voltage across C_1 during capacitor charge (diode conducting) can now be expressed as

$$V_{C_1}(t) = V_{C_1(dc)} + V_{C_1(ac)} \sin(\omega t + \theta). \tag{12}$$

where $V_{C_1(ac)}$ and θ are obtained from the Fourier analysis of the voltage waveform across C_1 for the first approximation and $\psi=\theta\text{-}\tan^{-1}(\omega L/R_1)$. The circuit equation for $V_{C_1}(t)$ is shown in (2). The expression is defined for $0 \leq t \leq t_2 - t_1$.

The circuit equation for the voltage across C_1 during discharge (diode non-conducting) can be written from inspection of Fig. 2(d).

$$V_{C_1}(t) = \frac{q_0}{C_1} - \frac{1}{C_1}\int i_2 dt. \tag{13}$$

This expression is defined for $t_2 \leq t \leq (t_1 + 2\pi/\omega)$.

The solution of the two circuit equations, for this case, is obtained in a manner similar to that outlined in Appendix I. Solution of (2) results in the following expression:

$$\left[\frac{\sin \omega t_1}{2}\left(1+\frac{R_d}{R_1}\right)+\frac{\sin(2\phi-\omega t_1)}{2}+\frac{\sin \omega t_2}{2}\frac{R_d}{R_1}\right.$$
$$\left.+\frac{i_{2ac}R_1}{V_m}\frac{1/\omega C_1}{R_1}\sin\phi\sin(-\phi+\psi)\right]\epsilon^{\omega t_1/\tan\phi}$$
$$=\left[\frac{\sin \omega t_2}{2}\left(1+\frac{R_d}{R_1}\right)+\frac{\sin(2\phi-\omega t_2)}{2}+\frac{\sin \omega t_1}{2}\frac{R_d}{R_1}\right.$$
$$\left.+\frac{i_{2ac}R_1}{V_m}\frac{1/\omega C_1}{R_1}\sin\phi\sin(\omega t_2-\omega t_1-\phi+\psi)\right]\epsilon^{\omega t_2/\tan\phi}, \tag{14}$$

where

$$i_{2ac} = \frac{V_{C_1(ac)}}{\sqrt{R_1^2+(\omega L)^2}}$$

and $\phi=\tan^{-1}(R_d/X_{C_1})$. This is the equation during charge for steady-state conditions for the second successive approximation. Solution of (13) results in the following expression:

$$\sin(\omega t_1)\left[1+\frac{1/\omega C_1}{2R_1}(2\pi+\omega t_1)\right]+\frac{\sin\omega t_2}{2}\frac{1/\omega C_1}{R_1}\omega t_1$$
$$-\frac{i_{2ac}R_1}{V_m}\frac{1/\omega C_1}{R_1}\cos(\omega t_1+\psi)$$
$$=\sin(\omega t_2)\left[1+\frac{1/\omega C_1}{2R_1}(\omega t_2-2\pi)\right]+\frac{\sin\omega t_1}{2}\frac{1/\omega C_1}{R_1}$$
$$-\frac{i_{2ac}R_1}{V_m}\frac{1/\omega C_1}{R_1}\cos(\omega t_2+\psi). \tag{15}$$

This is the equation during discharge for steady-state conditions for the second successive approximation. Eqs. (14) and (15) are solved graphically in the same manner as expressions (9) and (10). In this case a value of L must be chosen. The manner of choosing a value for L will be discussed later. Values for θ and $V_{C_1(ac)}$ are obtained from the Fourier analysis of the waveform across C_1 from the solution by the first approximation for the same R_1, R_d and C_1 with $\psi=\theta\text{-}\tan^{-1}\omega L/R_1$ as before. As in the first approximation, some repetition in calculations is required to obtain agreement between the assumed values of ωt_1 and ωt_2 and their final values. The results of the solution by the second approximation are plotted as X's on Fig. 1 and Fig. 4 for particular choices of parameters.

Discussion of Results

In Fig. 4 are plotted curves (solid lines) which show the values of ωt_1, and ωt_2 which satisfy steady-state conditions, that is, satisfy (9) and (10) simultaneously, for the case of an assumed infinite inductance (first successive approximation) in the broadband detector circuit. Curves are shown for values of R_1/R_d of 2.5, 5.0. 7.5 and 10.0 and they were obtained by varying $K(R_1/X_{C_1})$ as outlined above. The dashed lines represent curves which satisfy (8) for values of carrier detection efficiency D from 10 per cent to 90 per cent in 10 per cent steps. Eq. (8) is based on the assumption, which is very nearly true, that the average rectified voltage across the detector output capacitor C_1 for the charge period is the same as for the discharge period. Eq. (9) and (10) are also based on the assumption expressed in (8). Therefore their solution for the values of ωt_1 and ωt_2 which satisfy steady-state conditions also provides the D which corresponds to these values of ωt_1 and ωt_2. The plotted curves (solid lines) of Fig. 4 are then plots of carrier detection efficiency D as of function of both the angles ωt_1 and ωt_2 (at which diode conduction starts and stops respectively) and the dimensionless quantity $K = R_1/X_{C_1}$. Two of the curves of Fig. 4 (for values of R_1/R_d of 2.5 and 5.0) are replotted in Fig. 1 as a function of K to present a comparison of the carrier detection efficiency D of the broadband detector output circuit with that of the standard rc detector output circuit.

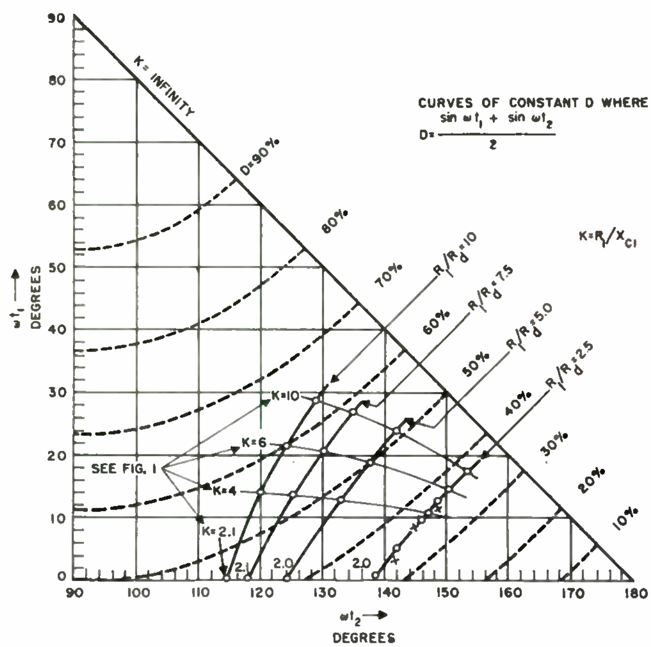


Fig. 4

Because it is not practical to make inductance L infinite, the question of how large L must be, practically speaking, arises. The solution obtained from the second successive approximation (where L is considered finite) provides a method of determining a minimum value for L. Carrier detection efficiency D was calculated for the case of $R_1/R_d = 2.5$ using (14) and (15), second successive approximation, as outlined above. Calculations were made for values of K of 2.59, 3.53 and 4.67 with ωL equal to $3R_1$ at the lowest frequency considered ($K = 2.59$). Calculated values of D are plotted as x's on Fig. 4 and Fig. 1. Comparison of results obtained using the solution from the first successive approximation and the second successive approximation shows only slight difference in D for the same values of K. It appears safe to assume, therefore, that the solution obtained by the first successive approximation represents the behavior of physically realizable broadband detector circuits for the case where ωL is at least $3R_1$ at the lowest carrier frequency considered.

Operation with a Modulated Input

Let us examine the video bandwidth of the broadband detector output circuit with a signal applied to the detector from a constant voltage source. In the following discussion, simplifying assumptions have been made and a detailed analysis of detector operation to a modulated signal has not been attempted. Various authors[6-8] have analyzed the factors which produce distortion of the modulation envelope in the detection process.

Consider the case of a pulsed sinewave applied to the detector to produce a detected video output pulse. From the solution for the first successive approximation (infinite inductance or practically speaking $\omega L \geqq 3R_1$), with the detector diode not conducting, capacitor C_1 discharges through R_1 with a constant current i_{20}. If the voltage across C_1 (the amplitude of the detected pulse) is designated V_{C_1}, then

$$i_{20} = V_{C_1}/R_1. \tag{16}$$

From the relationship

$$V = V_{C_1} - 1/C_1 \int i_{20} dt \tag{17}$$

$$dV/dt = - i_{20}/C_1, \tag{18}$$

we see that the rate of discharge of C_1 is

$$V_{C_1}/R_1C_1. \tag{19}$$

At the end of the pulsed sinewave, the average time required for the voltage across C_1 to drop to zero is

$$t = R_1C_1. \tag{20}$$

This is compared to a rise (or fall) time of 2.2 rc to the 90 per cent (or 10 per cent) point of a step function for an rc circuit. Because of the inductance L, there will be a tendency to continue charging C_1 to an opposite polarity. This will cause the detector diode to conduct and with a zero (dc) impedance signal source driving the detector, the voltage across C_1 will go negative an amount $i_{20}R_d$. The time constant $L/(R_1+R_d)$ will then determine the duration of this effect.

[6] S. N. Van Voorhis, "Microwave Receivers," Radiation Laboratory Series, Vol. 23, McGraw-Hill Book Company, New York; 1948.
[7] F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Company, New York; 1948.
[8] M. J. O. Strutt, "Ultra- and Extreme-Short Wave Reception," D. Van Nostrand Company, Inc. New York; 1947.

Next consider an amplitude modulated signal

$$v(t) = V_m(1 + m \sin \alpha t) \sin \omega t. \tag{21}$$

The function of the second detector is to provide an output that follows the envelope of this modulated signal; that is, its output must follow the expression

$$1 + m \sin \alpha t. \tag{22}$$

In particular, detector capacitor C_1 must discharge at a rate equal to or greater than the maximum time rate of change of this expression to follow the modulation envelope, practically speaking. The time derivation of this expression is

$$\frac{d}{dt}(1 + m \sin \alpha t) = \alpha m \cos \alpha t. \tag{23}$$

Remembering that $\alpha = 2\pi f_{\text{mod}}$ and that the above expression is maximum when $\cos(\alpha t) = 1$, we can write

$$1/R_1C_1 \geqq 2\pi m f_{\text{mod}} \tag{24}$$

or

$$f_{\text{mod}} \leqq 1/2\pi m R_1 C_1. \tag{25}$$

This can also be written as

$$X_{C_1,\text{mod}}/R_1 \geqq m. \tag{26}$$

Appendix I

Circuit equation (2) is solved in the following manner for the first approximation. For $0 \leqq t \leqq (t_2 - t_1)$, during diode conduction,

$$i_1(t) = \frac{v(t) - V_{C1}(t)}{R_d} \tag{27a}$$

and

$$i_2 = i_{20}. \tag{27b}$$

After substituting (3), (27a) and (27b) into (2) and performing the necessary integration, (2) can be written as

$$V_{C1}(t) + \frac{1}{R_dC_1}\int V_{C1}(t)dt = V_m \sin \omega t_1 + \frac{V_m}{R_dC_1\omega}\cos \omega t_1 - \frac{i_{20}}{C_1}t - \frac{V_m}{R_dC_1\omega}\Big[\cos(\omega t_1)\cos(\omega t) - \sin(\omega t_1)\sin(\omega t)\Big]. \tag{28}$$

This expression for the voltage spectrum across C_1 can be written in operational form by applying the LaPlace transform[9,10] to both sides of (28). After rearranging terms, the operational expression becomes

$$V_{C_1}(p) = \frac{1}{p + 1/R_dC_1}\Bigg[\left(1 + \frac{1}{R_dC_1}\frac{p}{p^2 + \omega^2}\right) \cdot (V_m \sin \omega t_1) - \frac{1}{p}\frac{i_{20}}{C_1} + \frac{\omega}{p^2 + \omega^2}\frac{V_m \cos \omega t_1}{R_dC_1}\Bigg]. \tag{29}$$

[9] M. F. Gardner and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc. New York; 1942.
[10] G. A. Campbell and R. M. Foster, "Fourier Integrals for Practical Applications," Bell System Monograph B-584, 1931.

The synthesis of this voltage spectrum into a time function $v_{C_1}(t)$ can be accomplished by applying the inverse LaPlace transform to both sides of the above expression using the relationship

$$L^{-1}v(p) = v(t) = \frac{1}{2\pi j}\int_{-j\infty}^{+j\infty} v(p)\epsilon^{pt}dp. \tag{30}$$

The integral, applied term by term to (29), is evaluated by the Cauchy integral theorem (residue theorem).

After rearranging and collecting terms, the expression

$$\frac{v_{C1(t)}}{V_m} = \sin\phi \cos(\omega t_1 - \phi)\epsilon^{-\omega t/\tan\phi} - \frac{i_{20}R_1}{V_m}\frac{1/\omega_{C1}}{R_1}\tan\phi(1 - \epsilon^{-\omega t/\tan\phi}) + \cos\phi \sin(\omega t + \omega t_1 - \phi) \tag{31}$$

is obtained, where

$$\phi = \tan^{-1}\frac{\omega}{1/R_dC_1}.$$

This expression, which is defined for $0 \leqq t \leqq (t_2 - t_1)$, represents the normalized time expression for the voltage across the detector capacitor during charge.

The circuit equation during capacitor discharge is solved by substituting (5) into (4) and dividing both sides of the equation by V_m to obtain

$$\frac{v_{C1(t)}}{V_m} = \sin \omega t_2 - \frac{i_{20}R_1}{V_m}\frac{1/\omega_{C1}}{R_1}(\omega t - \omega t_2). \tag{32}$$

This expression, which is defined for

$$t_2 \leqq t \leqq \left(\frac{2\pi}{w} + t_1\right),$$

represents the normalized time expression for the voltage across the detector capacitor during discharge.

During steady-state conditions, the voltage at the end of detector capacitor charge is equal to the voltage at the beginning of discharge and vice versa. By setting $t = t_2 - t_1$ in expression (31) for $v_{C_1}(t)/V_m$ during charge and $t = t_1 + 2\pi/\omega$ in expression (32) for $v_{C_1}(t)/V_m$ during discharge, two simultaneous equations are obtained which contain t_2 and t_1.

Substituting $t = t_2 - t_1$ in (31) for $v_{C_1}(t)/V_m$ and rearranging and collecting terms results in (6). Substituting $t = t_1 + 2\pi/\omega$ in (32) for $v_{C_1}(t)/V_m$ and rearranging and collecting terms results in (7).

Acknowledgment

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The Excitation of Surface Waves by a Vertical Antenna*

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***Summary*—The excitation of surface waves by a vertical, linear antenna erected on an infinite, flat, perfectly conducting image plane coated with a thin layer of dielectric is investigated theoretically. Helmholtz-type integral expressions, derived through the use of the results of a previous paper, are utilized in the computation of numerical data. The appendix is devoted to the development of an approximate current distribution for the antenna.**

The far-zone fields for the configuration are composed of two types of waves—the radiated wave, a wave of spherical type, and one or more surface waves, waves of cylindrical TM type which attenuate exponentially with height above the dielectric surface. The theoretical magnitudes of these two are combined to yield field patterns that show the large magnitude of the surface-wave component excited by the vertical antenna.

The experimental approximation to the configuration consisted of a large ground plane enclosed by a microwave dark room. The experimental data, in addition to being presented so as to verify the theory, were replotted in a form that yields convenient design curves for surface wave antennas of this type.

Additional measurements were taken over partially coated surfaces to aid in both the understanding of the phenomenon and its practical application to directive radiators.

Introduction

In recent years much interest has been shown in electromagnetic waves guided by modified, perfectly conducting or reactive surfaces. Initially this interest was concentrated on the cylindrical surface, such as the single-wire line studies of Goubau[1] and Roberts.[2] Attwood,[3] who studied the phenomenon of waves guided by a plane, reactive surface, did so not with the intention of utilizing it as such, but with the idea of relating it to the cylindrical surface and of suggesting its use in waveguides. More recently, other authors have considered the reactive plane surface with the aim of using its properties to control radiation.[4]

While the virtues of the plane-coated surface are being realized in practice, very little work of a rigorous nature has been done on the excitation problem, with the exception of that of Elliott,[5] Lo,[6] and Brick.[7] The vertical dipole, the simplest and one of the more efficient surface-wave launchers, seems to have received little attention apart from the theoretical treatments of a Hertzian dipole in the presence of such a surface.[6,7] This may be due to the fact that the theoretical analysis was in need of extension to the practical problem of the finite cylindrical antenna and of experimental verification, and the lack of adequate design data. The purpose of this paper is to provide these.

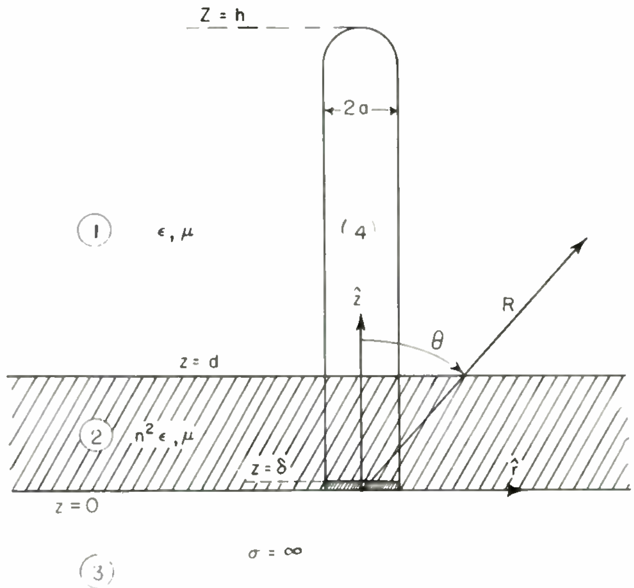

Fig. 1—Theoretical model of the finite transmitting antenna.

The configuration considered is that of a vertical cylindrical antenna erected on a plane conductor which is coated with a thin layer of dielectric, Fig. 1. It will be seen that such an antenna excites two types of waves in the space above the dielectric—those of spherical type, the radiation or compensating field, and those of cylindrical transverse-magnetic type, which attenuate exponentially with distance z above the surface and propagate with velocity less than that characteristic of

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[1] G. Goubau, "Surface waves and their application to transmission lines," *Jour. Appl. Phys.*, vol. 21, pp. 1119–1128; November, 1950; "Single-conductor surface-wave transmission lines," Proc. I.R.E., vol. 39, pp. 619–624; June, 1951; "On the excitation of surface waves," Proc. I.R.E., vol. 40, pp. 865–868; July, 1952.

[2] T. E. Roberts, "Theory of the single-wire transmission line," *Jour. Appl. Phys.*, vol. 24, pp. 57–67; January, 1953.

[3] S. Attwood, "Surface wave propagation over a coated plane conductor," *Jour. Appl. Phys.*, vol. 22, pp. 504–509; April, 1951.

[4] See for instance: Walter Rotman, "The Channel Guide Antenna," and "Metal Clad, Progressive-Phase, Dielectric Antennas," Air Force Cambridge Research Center Reports Nos. E. 5054 and E 5081, Cambridge, Mass.; F. E. Butterfield, "Dielectric sheet radiators," Trans. I.R.E., vol. AP-3, pp. 152–158; October, 1954; Progress and Final Reports on "Ridge and Corrugated Antenna Studies," Stanford Res. Inst., Stanford, Calif.; M. J. Ehrlich and L. Newkirk, "Corrugated Surface Antennas," 1953 IRE Convention Record, Part 2, "Antennas and Communications," pp. 18–33.

[5] R. S. Elliott, "On the theory of corrugated plane surfaces," Trans. I.R.E., vol. AP-2, pp. 71–81; April, 1954.

[6] Y. T. Lo, "Electromagnetic field of a dipole source above a grounded dielectric slab," *Jour. Appl. Phys.*, vol. 25, pp. 733–740; June, 1954.

[7] D. B. Brick, "The Radiation of a Hertzian Dipole Over a Coated Conductor," *Monograph No. 113, The Inst. Elec. Engrs.*, December, 1954, and *Proc. I.E.E.*, Part C, 1955.

region 1. The latter are alternately called surface, guided, slow, or residue waves. The number of waves of this type which may be supported increases with dielectric thickness. All thicknesses considered here are limited to the region of single-mode propagation.

A theoretical formula for the electric field will be derived. It will be of the form of a sum of integrals containing, as integrands, products of the antenna current and the responses to unit sources (the Green's functions) which satisfy the same boundary conditions as the fields. These Green's functions were derived in a previous paper[7] and will not be duplicated here. They are available in closed form only in the far zone of the unit source. The exact current distribution on the antenna is unknown. A method for approximating it is given in the Appendix.

Although the theory is limited strictly to a surface of infinite extent, it will be shown experimentally that the coupling to the surface wave on a large but finite coated surface is closely approximated by the values obtained from the formulas.

In order to conform with the usual field-pattern method of displaying the radiating properties of an antenna, the θ-component of the total far-zone electric field has been numerically computed as a function of θ at a specific distance R (Fig. 1). It is important to note that a specific distance R must be chosen, since the propagation constants and the radial dependences of the two types of waves, which comprise the total field, are different. Another point of departure is that, unlike the radiated wave, the surface waves are of TM type and hence have radial components of the electric field in addition to θ-components. The relationship between the radial- and θ-components of the surface wave has been treated in the literature.[7]

The experimental measurements were performed to verify the theory. This was done in two ways—first, the surface wavelengths were measured and compared with those resulting from the transcendental equation for the determination of the wavelength of the guided mode, and secondly, field patterns were measured and compared with the computed results. These patterns illustrate the large distortions of the dipole fields caused by the presence of the dielectric layer. The large spikes in the patterns near the surface ($\theta \doteq 90°$) are due to the surface waves.

The ratios of the maxima of the θ-components of the E-fields of the surface waves to those of the radiated waves are plotted vs dielectric thickness, with dipole length as a parameter, for all the cases measured. These are intended for use as design curves. In order to give additional physical interpretation to the phenomenon, curves of measurements taken over partially coated surfaces are also shown.

The Theoretical Formulation

In a previous paper[7] the fields due to a vertical Hertzian dipole above or in a dielectric layer coating a flat, perfectly conducting plane were derived. If the dipole moment is set equal to $-j/\omega$, these are just the responses to infinitesimal unit current sources in the z-direction, i.e. $-j\omega\mu$ times, the Green's functions. Due to the linearity of the field equations, superposition prevails and the response to an arbitrary distribution of z-directed current densities is given by the integral of the product of the appropriate unit response function and the current density.

King has shown that for thin, linear antennas the total current may be treated as if it were a current density lying along the axis of the antenna.[8] Hence, if the assumption is made that $\beta a \ll \beta h$ and $\beta a \ll 1$, the electric fields in regions 1 and 2 (Fig. 1) are given by:

Region 1, Fig. 1, $z \geqq d$

$$\vec{E}_1(R,\theta) = \int_d^h I_z(z')\vec{E}_{(A)1}(R,\theta,z')dz' + \int_0^d I_z(z')\vec{E}_{(B)1}(R,\theta,z')dz' \qquad (1a)$$

Region 2, Fig. 1, $0 \leqq z \leqq d$

$$\vec{E}_2(R,\theta) = \int_d^h I_z(z')\vec{E}_{(A)2}(R,\theta,z')dz' + \int_0^d I_z(z')\vec{E}_{(B)2}(R,\theta,z')dz' \qquad (1b)$$

where $I_z(z')$ is the total current on the antenna and $E_{(A)1}$, $E_{(B)1}$, $E_{(A)2}$, $E_{(B)2}$ are the fields due to infinitesimal unit current sources. These are obtained from the corresponding fields in the previous paper[7] by setting the dipole moment p equal to $-j/\omega$. The subscripts (A) and (B) refer to the region in which the unit source is placed, and subscripts 1 and 2 refer to the region of observation. $\beta = \omega\sqrt{\epsilon\mu}$ is the characteristic propagation constant in region 1, and $\omega = 2\pi$ times the frequency.

At large distances from the antenna, $\vec{E}_{(A)1}$ and $\vec{E}_{(B)1}$ are composed of both surface and radiated waves, whereas $\vec{E}_{(A)2}$ and $\vec{E}_{(B)2}$ are composed of surface waves only.[7] Due to the superposition properties of (1a) and (1b), $\vec{E}_1$ is composed of both types of fields, while $\vec{E}_2$ is composed of only surface waves for large R. The propagation constants and functional forms of the surface modes are functions of the dielectric layer only and are hence identical to those due to a Hertzian dipole source.[7] The determination of $I_z(z')$ is discussed in the appendix.

Numerical evaluations of (1a) as a function of θ have been made for $\beta h = \beta d = .10\pi$, 15π, $n\pi/4$ and for $\beta h = \beta d + \pi/4$, $\beta d + \pi/2$, $\beta d + \pi$ with $\beta d = .10\pi$ and $.15\pi$ and $n^2 = 2.54$ (polystyrene) at $\beta R = 170.6$ radians. Examples of these are displayed with the experimental results in Figs. 2–5. The values of the parameters βd and n^2 used for the theoretical curves were chosen to be slightly different from those used for the experimental curves, so that the effects of the slight variations in these parameters could be observed, yet they were close enough so that the experimental results agreed well with theory.

[8] R. W. P. King, "Electromagnetic Engineering," vol. 1, McGraw-Hill, Inc., New York, pp. 242–243; 1945.

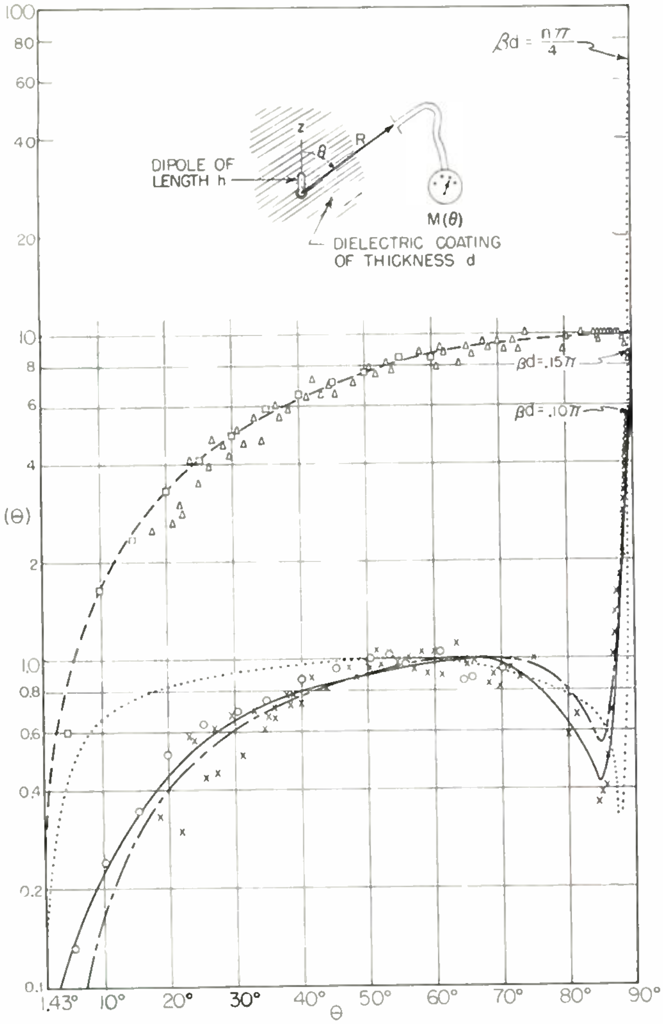

Fig. 2—Experimental and theoretical field patterns at $\beta R = 170.6$ radians.
Experimental:
$\beta h = \beta d = .134\pi$, $n^2 = 2.58$, $\Omega = 3.9$ Polyrod—○ Waveguide—×
$\beta h = .134\pi$, $\beta d = 0$, $\Omega = 3.9$ (10×scale) Polyrod—□ Waveguide—△
Theoretical:
$\beta h = \beta d$, $n^2 = 2.54$, $\Omega = \infty$ $\begin{cases}\beta d = .10\pi \\ \beta d = .15\pi \\ \beta d = n\pi\frac{1}{4}\end{cases}$
$\beta h = .134\pi$, $\beta d = 0$, $\Omega = \infty$ (10×scale)

The quantity $M(\theta)$ that is plotted on a logarithmic scale in Figs. 2–5, is the magnitude of the θ-component of the radiation wave. For Fig. 5, see page 724.

As previously mentioned, it is necessary to choose a value for βR when combining the fields, because the surface wave attenuates radially as $1/\sqrt{r}$ and propagates as a "slow wave," while the radiated wave attenuates as $1/R$ and propagates with velocity $(\mu\epsilon)^{-1/2}$. As βR increases, the ratio of the magnitudes of the surface to radiated waves increases, and the minimum between the main regions occupied by the two waves decreases. $\beta R = 170.6$ radians was chosen so as to agree with the experimental value.

Experimental Equipment

The infinite perfectly conducting plane assumed in theory was approximated in practice by a large (71.9λ ×81.5λ), but finite, Dural image plane. It was equipped with a coaxial feed for the cylindrical transmitting an-

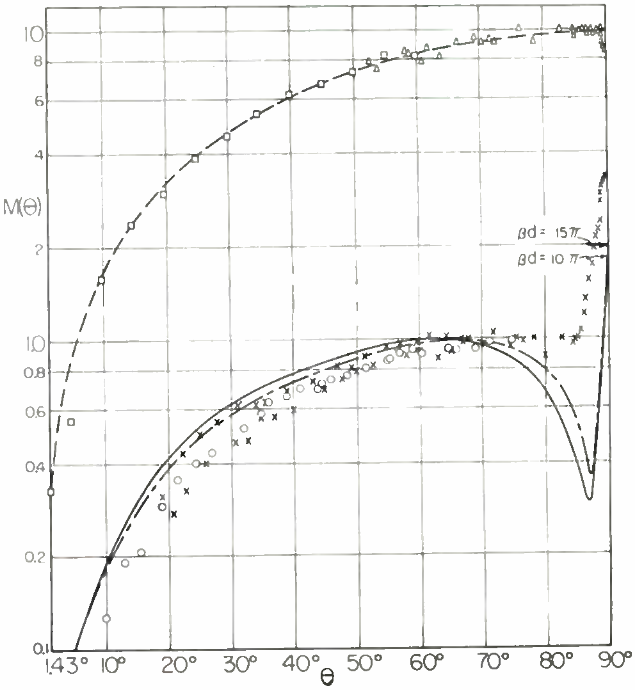

Fig. 3—Experimental and theoretical field patterns at $\beta R = 170.6$ radians.
Experimental:
$\beta h = \pi/4 + \beta d = .320\pi$, $\beta d = .070\pi$, $n^2 = 2.58$, $\Omega = 5.7$ Polyrod—○ Waveguide—×
$\beta h = .320\pi$, $\beta d = 0$, $\Omega = 5.7$ (10×scale) Polyrod—□ Waveguide—△
Theoretical:
$\beta h = \pi/4 + \beta d$, $n^2 = 2.54$, $\Omega = \infty$ $\begin{cases}\beta d = .10\pi \\ \beta d = .15\pi\end{cases}$
$\beta h = .320\pi$, $\beta d = 0$, $\Omega = \infty$ (10×scale)

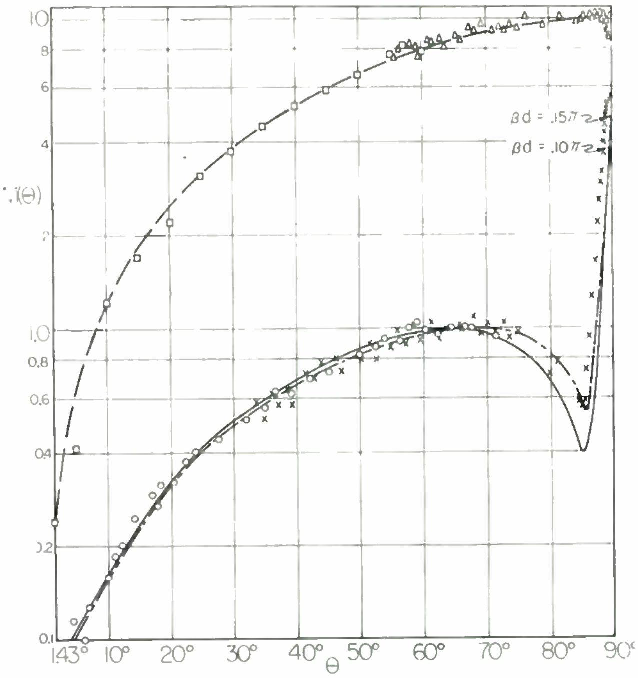

Fig. 4—Experimental and theoretical field patterns at $\beta R = 170.6$ radians.
Experimental:
$\beta h = \pi/2 + \beta d = .603\pi$, $\beta d = .103\pi$, $n^2 = 2.58$, $\Omega = 7.0$ Polyrod—○ Waveguide—×
$\beta h = .603\pi$, $\beta d = 0$, $\Omega = 7.0$ (10×scale) Polyrod—□ Waveguide—△
Theoretical:
$\beta h = \pi/2 + \beta d$, $n^2 = 2.54$, $\Omega = \infty$ $\begin{cases}\beta d = .10\pi \\ \beta d = .15\pi\end{cases}$
$\beta h = .603\pi$, $\beta d = 0$, $\Omega = \infty$ (10×scale)

tenna, a movable surface probe to measure surface wavelength, and a rotating frame to support a field pattern measuring antenna. The equipment was designed for X-band operation. To minimize reflections, the space above the ground screen was enclosed by absorbing material to form an electromagnetic anechoic chamber.

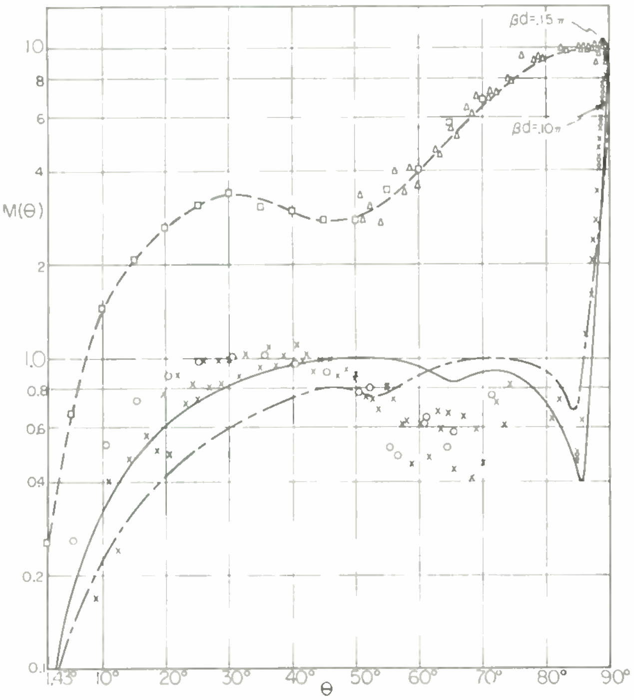

Fig. 5—Experimental and theoretical field patterns at $\beta R = 170.6$ radians.
Experimental:
$\beta h = \pi + \beta d = 1.134\pi$, $\beta d = .134\pi$, $n^2 = 2.58$, $\Omega = 8.3$
Polyrod—○ Waveguide—×
$\beta h = 1.134\pi$, $\beta d = 0$, $\Omega = 8.3$ (10×scale) — — —
Polyrod—□ Waveguide—△
Theoretical:
$\beta h = \pi + \beta d$, $n^2 = 2.54$, $\Omega' = 2ln[2(h-d)/a] = 8$
$\beta d = .10\pi$ ——— – ———
$\beta d = .15\pi$ ————————

Three thicknesses of dielectric were used to coat the ground screen. They consisted of Dow-Styron 475, a modified polystyrene ($n^2 = 2.58$) of thicknesses $d = .044''$, $.065''$ and $.084''$ or of electrical thicknesses, $\beta d = .070\pi$, 0.103π, and 0.134π respectively at 9,450 mcps. The sheets were cemented with electrical varnish and the surface probe travelled through a slot in each dielectric layer.

Although the direction of probe travel was not radially outward from the transmitter, as may be seen in Fig. 6, the measuring apparatus attached to the probe allowed the desired radial separations to be measured. This was accomplished by fixing one end of a rigid steel rule to the travelling probe mount and allowing it to have rotational freedom about the probe axis. The other end of the rule was made to slide in a vernier scale which was mounted beneath the transmitting antenna with rotational freedom about the antenna axis. The rule was well below the probe and antenna feeds so as not to interfere with their operation.

Measurements were taken with two field pattern antennas. For large distances above the ground-screen surface a polyrod antenna was used because if its high gain which discriminated against waves reflected by the absorbing chamber. Near the ground screen surface its use was not feasible due to its strong image coupling, large size precluding a close approach to the surface, and its large aperture which averaged the field over an electrically large region in both the angular and radial directions. Hence an antenna that did not have these

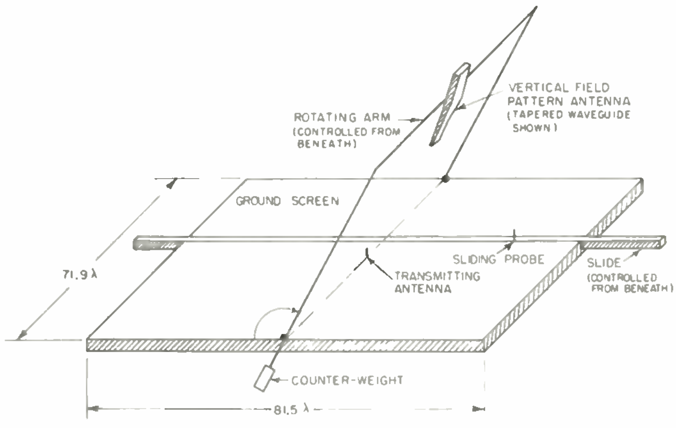

Fig. 6—X-band ground screen and associated equipment.

limitations was needed near the surface. A tapered waveguide antenna, consisting of an open-ended standard size waveguide tapered to one-fifth its narrow dimension at the aperture, was chosen. The external waveguide was covered with absorbing material to minimize external currents and concentrate the effective receiving area at the narrow aperture. A combined experimental and theoretical method was used to calibrate variations of effective receiving length as a function of height for each dielectric thickness.[9] This quantity has strong functional dependence on dielectric thickness.

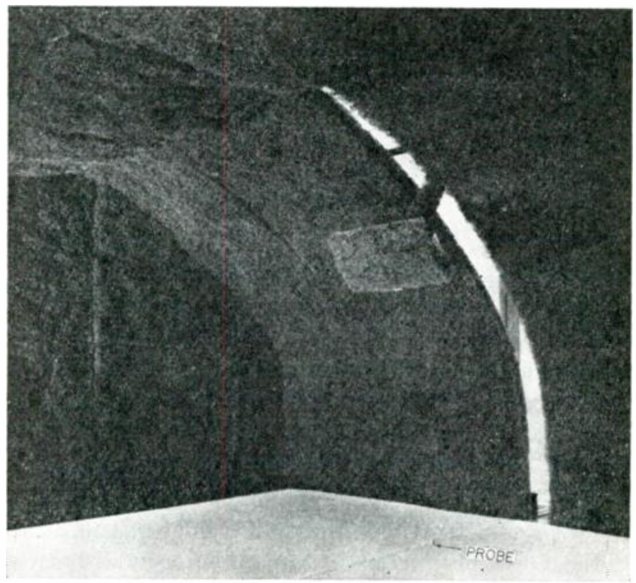

Fig. 7—Microwave anechoic chamber.

Fig. 6 is a schematic diagram of the equipment and Fig. 7 shows the dielectric-coated ground screen, surface probe, anechoic chamber, and tapered-waveguide antenna covered with absorbing material.

[9] The method is described at length by D. B. Brick, "An Experimental Investigation of the Radiation of a Vertical Antenna over a Coated Conductor," Technical Report No. 195, Cruft Laboratory, Harvard University, Cambridge, Mass.; 1954.

Experimental Results

Surface Wavelength Measurements

The surface probe mentioned previously was utilized as a standing wave detector to measure the radial surface wavelength. Although the vswr along the surface was initially very low indicating that the absorbing material at the edge of the screen provided a good match for the surface waves, a higher vswr was induced, for the purpose of wavelength measurements, by placing a small metal sheet at the edge of the slot in the dielectric through which the probe traveled.

The results of these measurements for the three dielectric thicknesses are shown in Table I. N indicates the number of half-surface wavelengths over which the measurements were performed while the error given is rms deviation. The theoretical values are those computed from the transcendental equation:[6,7]

$$n^2\sqrt{\beta_g{}^2 - \beta^2} = \sqrt{n^2\beta^2 - \beta_g{}^2}\tan\left(d\sqrt{n^2\beta^2 - \beta_g{}^2}\right)$$

where β is the characteristic propagation constant in region 1 and β_g is the surface-wave propagation constant. n and d are defined in Fig. 1.

TABLE I

βd	N	Theoretical Surface Wave-length (cm)	Experimental Surface Wavelength (cm)	Theoretical $\frac{\beta_g}{\beta} = \frac{\lambda}{\lambda_g}$	Experimental $\frac{\beta_g}{\beta} = \frac{\lambda}{\lambda_g}$
0.134π	42	3.069	$3.06 \pm .008$	1.036	1.04
0.103π	28	3.115	$3.12 \pm .014$	1.021	1.02
0.070π	30	3.152	$3.14 \pm .015$	1.009	1.01

In order to check the effect of the slots in the dielectric on the wavelength measurements, the wavelength was measured with the slot empty and with the slot filled with Vaseline, a material very similar electrically to the Styron, so that a homogenous layer was more closely approximated. To within the limits of measurement no differences in the average wavelengths could be detected.

Field Pattern Measurements

Experimental measurements of the magnitudes of the θ-components of the electric fields were made for four antenna lengths, $\beta h = \beta d$, $\beta d + \pi/4$, $\beta d + \pi/2$, and $\beta d + \pi$, relative to each of the three dielectric layers with and without the dielectric coatings on the ground screen. The patterns taken over the uncoated screen serve as comparison curves so that the distortions caused by the coatings may be more readily appreciated. Figs. 2–5 contain examples of these patterns plotted on a logarithmic scale. The quantity given is again $M(\theta)$, the θ-component of the electric field normalized to the maximum value of the radiated wave, at $\beta R = 170.6$ radians. The data taken with the tapered-waveguide antenna was corrected by the calibration curve of the variations of its effective length with height mentioned earlier.[9]

The experimental and theoretical data are in good agreement. The major deviations are due to the imperfect absorbing qualities of the chamber.[9] The results verify the existence of a large surface wave that causes a considerable distortion of the original dipole field.

In Fig. 8, the quantity of design interest—the magnitude of the surface wave as a function of dielectric thickness with dipole length as a parameter—has been plotted. It is interesting to note that for thinner dielectric layers short antennas appear to be more efficient surface-wave launchers, while for thicker layers the longer antennas appear to be more efficient. This last relation is limited to the range of dipole lengths measured, because as the electrical lengths of the antennas increase beyond $\beta d + \pi$ the efficiency of surface-wave excitation falls off.

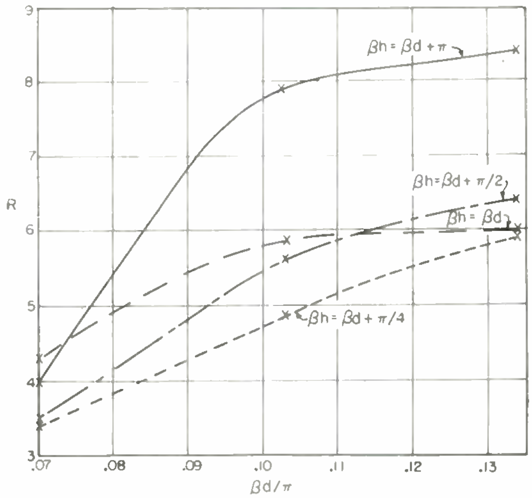

Fig. 8—Experimental ratio, R of the maximum of the θ-component of the electric field of the surface wave to that of the radiation wave for a dipole of length h and a coating of thickness d at $\beta R = 170.6$ radians, $n^2 = 2.58$.

For the dielectric thickness $\beta d = 0.103\pi$, field patterns were also measured with the ground-screen partially coated and partially uncoated, as shown in Figs. 9 and 10 on the next page.

These measurements were taken for several reasons: the first of these is to show that the excitation of surface waves is a phenomenon occurring continuously throughout the radial direction, but that its major contribution arises from the region within twenty wavelengths of the antenna. Hence, the excitation data obtained for an infinite coating is a good approximation for a large finite dielectric layer. One effect of increasing βc in Fig. 9 is to decrease the magnitude of the surface wave while causing only minor perturbations of the shape of the pattern of the radiation wave, thus agreeing with the observations of Ehrlich and Newkirk.[10] They observed that for a surface wave antenna of a different type the radiation pattern is composed essentially of a linear combination of the radiation of the feed alone, and that of the surface. The constants in this combination are determined by the degree of excitation of the surface waves and the geometry involved.

[10] M. J. Ehrlich and L. Newkirk, *loc. cit.*

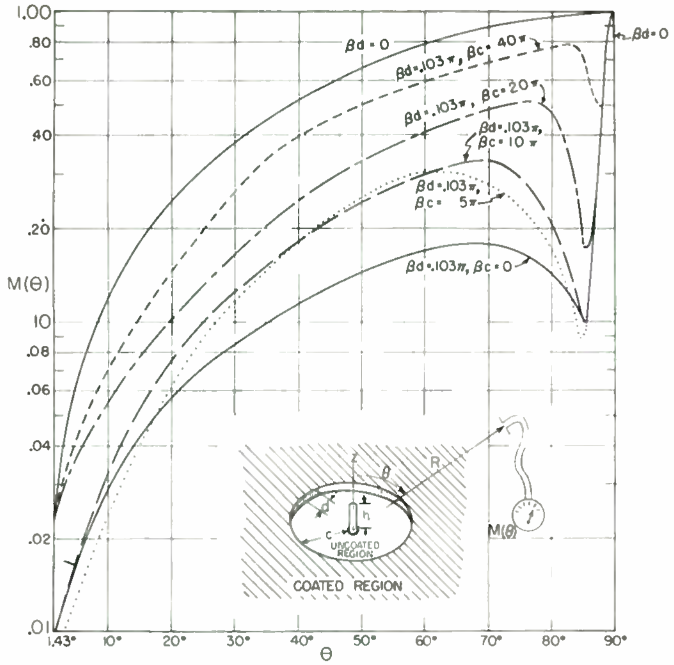

Fig. 9—Measured field patterns over a partially uncoated ground plane, $\beta h=0.603\pi$, $\beta R=170.6$ radians, $n^2=2.58$.

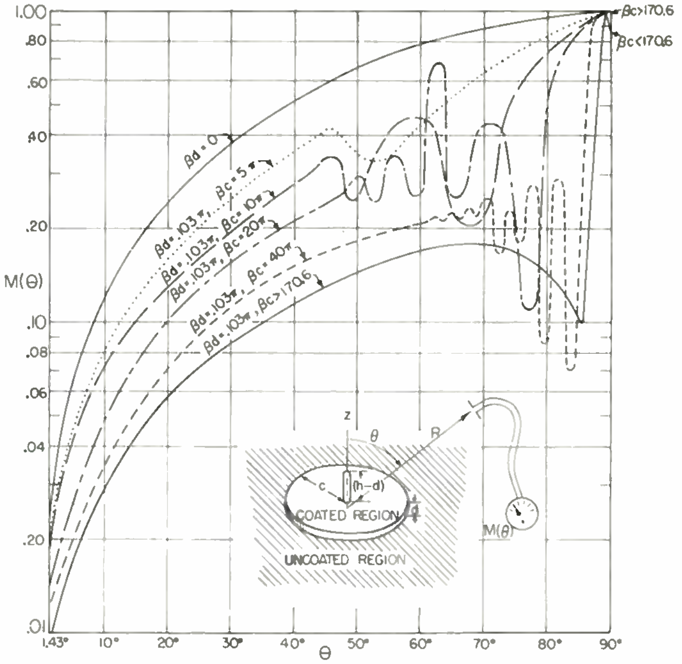

Fig. 10—Measured field patterns over a partially coated ground plane, $\beta h=0.603\pi$, $\beta R=170.6$ radians, $n^2=2.58$.

The measurements of Fig. 10 were taken in order to display the directive effect that occurs beyond the edge of the dielectric. The large oscillations that appear in the curves for larger βc are due to the fact that the measuring antenna was not sufficiently far from the combined dipole-dielectric surface radiator. The configuration of Fig. 10 has many possibilities in field pattern regulation. With proper shaping, choice of radius, and control of the dielectric properties, a highly directive antenna appears to be feasible. One such possibility is a planar analogue of the polyrod antenna. No effort in that direction was made at this time.

Discussion

This research has served to verify the previously derived theories and has demonstrated that a vertical antenna excites a cylindrical-type surface-wave component that is large compared to the radiation field. In doing so, it has also shown that a dielectric layer, whether of finite or infinite extent, introduces a considerable degree of perturbation directed towards broadside radiation. Data and theory have been provided as an aid towards the utilization of such a structure in the excitation of surface waves. Although a theoretical analysis of the layer of finite extent has not been attempted, it has been shown that the degree of excitation on a large surface is well approximated by that for an infinite surface. With this knowledge and the experimental and theoretical data made available, an analysis of the finite surface excited by a vertical antenna should not be difficult.

Appendix

Utilizing the z-components of the electric Hertz vectors for the configuration of Fig. 1 and the Hertz potentials for Hertzian dipole excitation, derived in a previous paper[7] as unit source or Green's functions, it is relatively straightforward to develop, in the usual manner,[11] a Hallén-type integral equation for the antenna of Fig. 1.[12] Although the iteration of this equation to obtain higher-than-zeroth order approximations to the current distribution is possible in theory, the fact that the potentials for Hertzian dipole excitation[6,7] are known only at large distances from the dipole precludes an iteration. The direct result of the integral equation is therefore limited to a zeroth-order distribution which is identical to that on an open-circuited, lossless, transmission line of length βh driven at $z=0$. This distribution is illustrated in Fig. 11 for several antenna lengths and dielectric thicknesses.

Guided by the integral equation and physical considerations of the problem, a better approximation to the current distribution was derived. The need for a better approximation is quite obvious if it is noted that the most important current elements for the excitation of surface waves are in the vicinity of the dielectric layer[7] and that the departures between the zeroth-order and exact current distributions on a dipole[11] are most marked in the region of the driving point. The approximation which has been made assumes a thin dielectric coating and is described as follows:

(1) The current distribution on the section of antenna in region 1, the space above the dielectric surface, is assumed to be essentially the same as if that section were half of a cylindrical transmitting antenna of total length $2(h-d)$ and radius "a" isolated in a space with characteristics, ϵ, μ, and zero conductivity. The King-Middleton approximate second-order distribution[11] was thus assumed to exist on this section, and was graphi-

[11] R. King and D. Middleton, "The cylindrical antenna, current and impedance," *Quart. Appl. Math.*, vol. 3, pp. 302–335; January, 1946.

[12] For the details of this development and the approximate determination of $I_z(z')$, see reference 9.

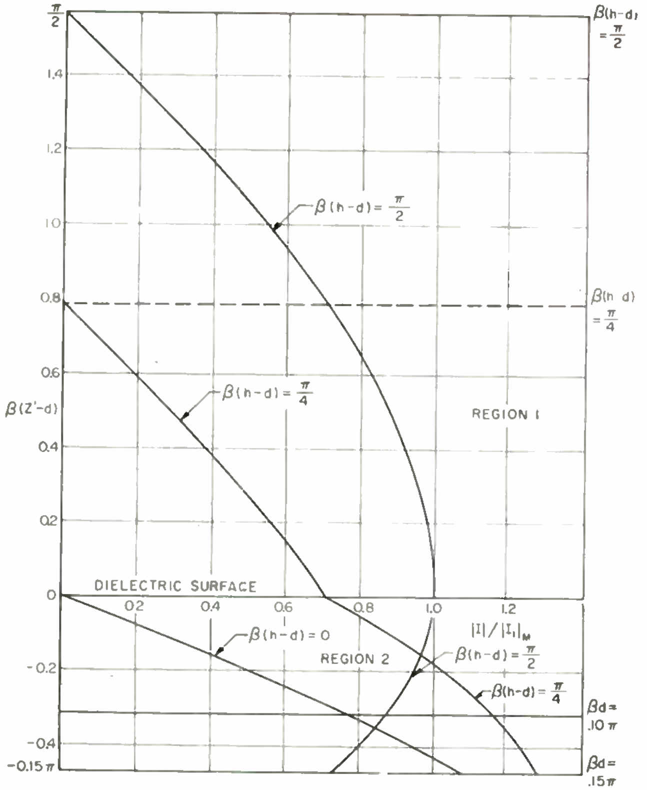

Fig. 11—Zeroth order current distribution on antennas for which $\beta(h-d)=0$, $\pi/4$, and $\pi/2$; and $\beta d=.10\pi$, and $.15\pi$.

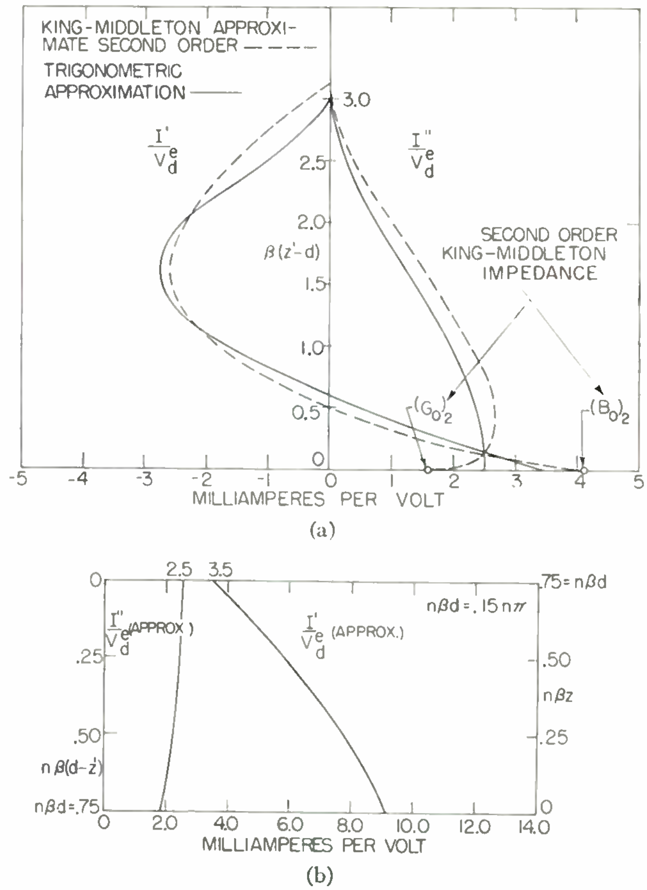

Fig. 12—(a) Assumed current distribution on the section of antenna of length $\beta(h-d)=\pi$, $\Omega^1=2ln[2(h-d)/a]=8$ above the dielectric layer; (b) assumed current distribution on the section in the dielectric layer of thickness d of an antenna of total length $\beta h=\pi+\beta d$, $n^2=2.54$, $n\beta d=.75$.

cally approximated by trigonometric and exponential functions that are easily integrable when substituted in equation (1a).

(2) The section of antenna in region 2 is assumed to be a very short half-section of top-loaded dipole isolated in the dielectric medium of region 2. The current distribution is then described by the following linear combination of trigonometric functions appropriate to the zeroth-order distributions

$$I_z = A\cos\left[n\beta(d-z)\right] + B\sin\left[n\beta(d-z)\right].$$

The approximate distributions in regions 1 and 2 are related at $z=d$ by

$$I_z(z=d+) = I_z(z=d-)$$

$$\frac{n^2\partial I_z(z=d+)}{\partial_z} = \frac{\partial I_z(z=d-)}{\partial_z}.$$

The first relation is due to the equation of continuity and the second is a relation satisfied exactly by the zeroth-order (transmission-line) distribution at the boundary and in good approximation by the exact distribution.[9] These relations determine the constants A and B and hence, these constants are independent of the dielectric thickness.

For short antennas ($\beta h \leqq \beta d+\pi/2$) the distribution reduces to the zeroth-order distribution but for larger antennas it is quite different. An example of this approximate distribution, $I=I''+jI'$, normalized to the driving-point voltage, V_d^e, on a fairly long antenna is shown in Fig. 12. In that figure the antenna length is $\beta h=\beta d+\pi$, $n\beta d=75$, $n^2=2.54$, and the approximate current distributions are given here by

Region 1 $d \leqq z \leqq h$

$$\frac{I''}{V_d^e} = 2.5\left[\frac{\cos\beta(z-d) - \cos\beta(h-d)}{1-\cos\beta(h-d)}\right]$$

$$\frac{I'}{V_d^e} = \left[3.5(1-e^{-2\beta(h-z)}) - 6.1\sin\beta(h-z)\right]$$

Region 2 $0 \leqq z \leqq d$

$$\frac{I''}{V_d^e} = 2.5\cos n\beta(d-z)$$

$$\frac{I'}{V_d^e} = 3.5\cos n\beta(d-z) + (6.1)n\sin n\beta(d-z).$$

The justification for this method of approximation is in the agreement of the field patterns with experiment, whereas for antenna lengths $\beta h > \beta d+\pi/2$, the zeroth-order approximation yields results that only vaguely resemble the measured values.

Acknowledgment

The author wishes to thank Professor R. W. P. King, and Drs. Tetsu Morita and Ronald Row for their advice and encouragement. He also wishes to express his gratitude to Miss Janet Connolly and Mrs. Elaine O'Malley for performing the computations, and to Mr. Charles Sampson, who built the equipment.

Coding for Constant-Data-Rate Systems—Part II Multiple-Error-Correcting Codes*

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Summary—The study of coding for constant-data-rate systems begun in Part I is continued. A simple pulse model of such a system is introduced which is easier to visualize than the correlation detection scheme of Part I. The Wagner code is extended to two multiple-error-correcting schemes. The efficiency of the Wagner code is evaluated by comparing it with Shannon's ideal coding. It is shown that the Wagner code is just the minimum probability of error detector for parity-checked sequences of independent binary digits.

I. Introduction

IN PART I of this paper[1] we introduced the Wagner code, a new means of correcting single errors in sequences of binary digits (words). The Wagner code differs from the Hamming code[2] by being *likely* rather than *certain* to correct single errors, but may be preferable in certain communication applications because of its use of only one check digit. In Part I it is assumed that the time-bandwidth product (TW) of the electrical signals representing the digits is large (so that there is a negligible increase in bandwidth when more check digits are accommodated in the same time interval), and that detection is by correlation. However, we might just as well represent the digits by positive and negative pulses of equal amplitude and use threshold detection. We must then assume that the system has extra bandwidth available to accommodate more check digits. It is shown in Appendix A that the analysis of Part I carries over intact for the simpler case of pulses.

Two extensions of the Wagner code are given in this paper. In Section II the Wagner principle is applied to the Hamming single-error-correcting, double-error-detecting code[2] by using the extra parity check digit to correct double errors. In Section III we show the advantage of dividing a long word into separately Wagner-coded subwords.

The constant-data-rate comparison presented in this paper (except in Section V of Part II) came about in answer to the practical question of whether to encode for reduced error rate (and if so, how), when the data rate (information rate at the channel input) and transmitter power are fixed. We consider only codes such that the receiver computes its own replacements for detected errors. There are codes in use which have the receiver ask the transmitter for a replacement for a detected error. See, for instance, the work of J. B. Moore.[3] A comparison of codes on an efficiency basis rather than in terms of per-word probability of error at constant-data-rate is possible, and is more pertinent from the standpoint of information theory. Section V is devoted to this complementary efficiency comparison.

Though the Wagner code is more properly described as a detection scheme to be used with parity-checked sequences of binary digits than as a "code," we feel that the phrase "Wagner code" is justified by its linguistic convenience.[4] All statements made in this paper as to the relative merits of different codes (except in Section V of Part II) apply only to a situation where the digits are converted into electrical signals and transmitted at constant-data-rate over a continuous channel perturbed by the addition of white Gaussian noise. The authors wish to avoid implying that the Wagner code is superior to the Hamming code in any over-all sense.

II. The Hamming-Wagner Code[5]

To extend the principle of the Wagner code to the construction of a double-error-correcting code, we add further check digits to the Wagner-coded word with the purpose of revealing double errors as well as single errors. If a double error is detected, we change the *two* digits of the stored word which are most in doubt; if a single error is detected, we change only the most doubtful digit.

The success of this scheme requires a system of check digits which indicates both single and double errors, and furthermore distinguishes between them. The usual geometrical model of message space is well suited for examining the possibility of setting up such check digits. In this model the various coded messages of n binary digits are represented by the vertexes of a cube in n dimensions, and the distance between two vertexes is defined as the number of binary digits in which the corresponding messages differ. The reader will easily see that a distance of at least four between the possible coded messages is required if both single and double errors are to be detected *and distinguished.*

In a Hamming single-error-correcting, double-error-detecting code, all transmitted messages are separated

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† Lincoln Lab., Mass. Inst. Tech., Lexington, Mass.

[1] R. A. Silverman and M. Balser, "Coding for constant-data-rate systems—part I. A new error-correcting code," Proc. IRE, vol. 42, pp. 1428–1435; September, 1954.

[2] R. W. Hamming, "Error detecting and error correcting codes," *Bell Sys. Tech. Jour.*, vol. 29, pp. 147–160; April 1950.

[3] J. B. Moore, "Accuracy and speed on short-wave teleprinter services," *Proc. Nat. Electronic Conf.*, vol. 9, pp. 927–934; February, 1954.

[4] The Wagner code is equivalent to the minimum probability of error detector. (See Appendix B.)

[5] The same reservation applies in calling this scheme a "code" as in calling the Wagner scheme a "code." (See Section I.)

by at least distance four. This is just the distance required for successful operation of a Wagner code that corrects both single and double errors. Thus the number of check digits needed to correct *all* single errors before applying the Wagner procedure to double errors is the same as the number required to apply the Wagner procedure to both single and double errors. This suggests a "Hamming-Wagner code" which obviously corrects more errors than the corresponding "Wagner-Wagner code."

In this way we arrive at a code which is like the Hamming single-error-correcting, double-error-detecting code, except that if the code indicates a double error, we change the *two* most doubtful digits in the stored word. The analysis of this Hamming-Wagner code is completely analogous to that of the simple Wagner code. As we saw in Part I, the probability of error per Wagner-coded word of m message digits is given by

$$P_W = 1 - q^{m+1}(\alpha) - \Pi_{m+1}(\alpha), \quad \text{(I-29)[6]}$$

where

$$\alpha = \frac{c_1 - c_2}{\sqrt{2(\sigma_1^2 + \sigma_2^2)}}, \quad (1)$$

if detection is by correlators with Gaussian outputs of means c_1 and c_2 $(c_1 > c_2)$ and standard deviations σ_1 and σ_2; or

$$\alpha = \frac{V}{\sqrt{2N_0 W}}, \quad (2)$$

if pulses of amplitude $\pm V$ and bandwidth W are received in the presence of white Gaussian noise of power N_0 per unit bandwidth. The definition of $q(\alpha)$ is

$$q(\alpha) = \tfrac{1}{2}(1 + \text{erf } \alpha). \quad \text{(I-14)}$$

where

$$\text{erf } x = \frac{2}{\sqrt{\pi}} \int_0^x \exp(-t^2)dt.$$

The quantity $\Pi_n(\alpha)$ is the integral

$$\Pi_n(\alpha) = \frac{n}{(\sqrt{\pi})^n} \int_0^\infty \exp[-(x+\alpha)^2] \cdot \left\{ \int_x^\infty \exp[-(y-\alpha)^2]dy \right\}^{n-1} dx$$

$$= \frac{2}{\sqrt{\pi}} \frac{n}{2^n} \int_0^\infty \exp[-(x+\alpha)^2] \cdot [1 - \text{erf}(x-\alpha)]^{n-1}dx. \quad (3)$$

Expanding the term in brackets by the binomial theorem, we get[7]

$$\Pi_n(\alpha) = \frac{n}{2^n} \sum_{i=1}^{n} \binom{n-1}{i-1} (-1)^{i-1} I_i(\alpha), \quad (4)$$

where

$$I_i(\alpha) = \frac{2}{\sqrt{\pi}} \int_0^\infty \exp[-(x+\alpha)^2][\text{erf}(x-\alpha)]^{i-1}dx. \quad \text{(I-23)}$$

In an entirely analogous fashion we find that the probability of error per Hamming-Wagner coded word is

$$P_{HW} = 1 - q^{m+k+1}(\alpha) - (m+k+1)q^{m+k}(\alpha)p(\alpha) - {}^2\Pi_{m+k+1}(\alpha), \quad (5)$$

where k is the number of check digits required by the Hamming single-error-correcting code.[2] The quantity ${}^2\Pi_n(\alpha)$ is the integral

$${}^2\Pi_n(\alpha) = \frac{n(n-1)}{(\sqrt{\pi})^n} \int_0^\infty \exp[-(x+\alpha)^2] \cdot \int_0^x \exp[-(y+\alpha)^2]dy \cdot \left\{ \int_x^\infty \exp[-(z-\alpha)^2]dz \right\}^{n-2} dx$$

$$= \frac{2}{\sqrt{\pi}} \frac{n(n-1)}{2^n} \int_0^\infty \exp[-(x+\alpha)]^2 \cdot [\text{erf}(x+\alpha) - \text{erf } \alpha][1 - \text{erf}(x-\alpha)]^{n-2}dx. \quad (6)$$

Expanding the term in brackets by the binomial theorem, we get

$${}^2\Pi_n(\alpha) = \frac{n(n-1)}{2^n} \sum_{i=2}^{n} \binom{n-2}{i-2} (-1)^i \, {}^2I_i(\alpha), \quad (7)$$

where

$${}^2I_i(\alpha) = \frac{2}{\sqrt{\pi}} \int_0^\infty \exp[-(x+\alpha)^2][\text{erf}(x+\alpha) - \text{erf } \alpha] \cdot [\text{erf}(x-\alpha)]^{i-2}dx. \quad (8)$$

$P_{HW}(1.35)$ and $P_{HW}(1.80)$ are tabulated in Table I (next page) for various values of m, also corresponding probabilities of error for uncoded, Hamming-coded, and Wagner-coded words. The values of α used in computing P_U, P_H, and P_W are chosen so that all words (message digits plus check digits) have the same duration, as required in a constant-data-rate system. Thus

$$P_U(\alpha_U) = 1 - q^m(\alpha_U), \quad \alpha_U = \sqrt{\frac{m+k+1}{m}}\,\alpha \quad (9)$$

$$P_H(\alpha_H) = 1 - q^{m+k}(\alpha_H) - (m+k)q^{m+k-1}(\alpha_H)p(\alpha_H), \quad \alpha_H = \sqrt{\frac{m+k+1}{m+k}}\,\alpha \quad (10)$$

$$P_W(\alpha_W) = 1 - q^{m+1}(\alpha_W) - \Pi_{m+1}(\alpha_W), \quad \alpha_W = \sqrt{\frac{m+k+1}{m+1}}\,\alpha. \quad (11)$$

[6] Equation numbers preceded by the Roman numeral I refer to correspondingly numbered equations in Part I.
[7] Eqs. (3) and (4) are to be preferred for their simplicity to (I-12) and (I-22). They are, of course, entirely equivalent.

Table I shows that P_H eventually becomes less than P_W as m increases. This happens for two reasons: (1) the ratio k/m decreases with increasing m, so that the corresponding values of α for the two codes become more nearly equal, which dissipates the advantage of the check-digit economy of the Wagner code; (2) the probability that the Wagner code corrects single errors decreases as m increases. Table I below also shows that P_{HW} is less than P_W for m greater than about 15.

III. The Syllabified Wagner Code

If the probability of error per digit is held constant, the proportion of multiple errors increases as the length of the word increases. Consequently, any single error-correcting code loses its effectiveness as m increases. One way to combat this effect without introducing a true multiple-error-correcting code is to divide each word into subwords or *syllables* separately coded for single error correction. In this way multiple errors can be corrected if no syllable contains an error in more than one digit.

Suppose a word with m message digits is divided into j syllables, each containing $n_i = m_i + k_i$ digits, where

$$m = \sum_{i=1}^{j} m_i$$

and k_i is the number of check digits required in the ith syllable by the single-error-correcting code in question. Let $Q(m_i)$ denote the probability that the ith syllable be correct after decoding. The probability of error for the syllabified word is then

$$P_S(m_1, m_2, \cdots, m_j) = 1 - Q(m_1)Q(m_2) \cdots Q(m_j). \quad (12)$$

For a given number of syllables and a given probability of error per digit it can be shown that P_S is smallest when the syllables have equal length (or as nearly equal as possible).

Under the assumption of constant-data-rate, syllabification is worthwhile only if the probability of error per digit is not increased excessively by the introduction of too many check digits. Because of its economy in the use of check digits (one per syllable) the Wagner code is well suited for syllabification. For Wagner-coded syllables,

$$Q(m) = q^{m+1}(\alpha) + \Pi_{m+1}(\alpha). \quad (13)$$

The optimum number of Wagner-coded syllables is a compromise between lowering α excessively by introducing too many check digits and having too small a probability of correction because of excessive syllable length. The optimum number of syllables is not necessarily critical or for that matter the same for all α. A syllable length of seven to ten digits seems to be the best, as illustrated in Table II.

TABLE I

Comparison of Hamming, Wagner, and Hamming-Wagner Codes

(a) $\alpha = 1.35$

m	P_U	P_H	P_W	P_{HW}
10	0.093	0.044	0.030	0.037
11	0.111	0.050	0.038	0.043
12	0.110	0.065	0.038	0.057
13	0.128	0.073	0.047	0.064
14	0.146	0.081	0.056	0.071
15	0.165	0.090	0.067	0.079
16	0.183	0.098	0.079	0.087
17	0.202	0.107	0.092	0.096
18	0.220	0.116	0.105	0.104
19	0.239	0.126	0.119	0.113
20	0.257	0.135	0.134	0.122
21	0.275	0.145	0.149	0.132

(b) $\alpha = 1.80$

m	P_U	P_H	P_W	P_{HW}
10	0.0091	0.0016	0.00067	0.00063
11	0.0117	0.0018	0.00095	0.00076
12	0.0109	0.0025	0.00081	0.00107
13	0.0135	0.0029	0.00111	0.00124
14	0.0163	0.0033	0.0015	0.0014
15	0.0193	0.0037	0.0019	0.0016
16	0.0224	0.0042	0.0024	0.0019
17	0.0258	0.0046	0.0030	0.0021
18	0.0292	0.0051	0.0037	0.0024
19	0.0328	0.0057	0.0044	0.0026
20	0.0364	0.0062	0.0052	0.0029
21	0.0401	0.0068	0.0061	0.0032
22	0.0439	0.0074	0.0070	0.0036
23	0.0478	0.0080	0.0080	0.0039
24	0.0518	0.0086	0.0091	0.0043

Values of α are for a Hamming-Wagner code.
m = number of message digits.

TABLE II

Comparison of the Hamming-Wagner and Syllabified Wagner Codes

$\alpha_{HW} = 1.80$

m	P_{HW}	C_{HW}	j	P_{SW}
12	0.00107	0.77	1	0.0081
			2	0.0087
14	0.00143	0.75	1	0.00148
			2	0.00144
16	0.00186	0.73	2	0.00228
18	0.00236	0.72	2	0.00322
			3	0.00342
20	0.00292	0.70	2	0.00438
			3	0.00449
22	0.00356	0.68	2	0.00577
			3	0.00570
24	0.00426	0.67	3	0.00700
			4	0.00735
30	0.00730	0.62	3	0.00981
			4	0.00975
			5	0.01006
42	0.0146	0.55	5	0.0200
			6	0.0201
54	0.0244	0.50	6	0.0318
			7	0.0317
72	0.0448	0.43	8	0.0468
90	0.0688	0.38	10	0.0659

j = number of syllables.

Table II presents a comparison of P_{HW} and P_{SW} (the per-word error probabilities for the Hamming-Wagner and syllabified Wagner codes) for $\alpha_{HW}=1.80$. P_{SW} is calculated from (12) and (13), using

$$\alpha_{SW} = \sqrt{\frac{m+k+1}{m+j}}\,\alpha_{HW}. \tag{14}$$

At about $m=80$, P_{SW} becomes less than P_{HW}. This is because C_{HW}, the probability that the Hamming-Wagner code corrects a detected double error, decreases as m increases, as shown in Table II. C_{HW} is given by

$$C_{HW}(\alpha) = \frac{2^2\Pi_n(\alpha)}{n(n-1)q^{n-2}(\alpha)p^2(\alpha)}, \tag{15}$$

where $n=m+k+1$ and $p(\alpha)=1-q(\alpha)$.

IV. Other Multiple-Error-Correcting Codes

In addition to the Wagner-type multiple-error-correcting "codes" described in Sections II and III, there exist algebraic multiple-error-correcting codes.[8–10] We have made no systematic study of the operation of such codes in constant-data-rate systems. We have, however, made a few calculations using a Muller-Reed triple-error-correcting code which show that despite the use of a large number of check digits, this code has a lower per-word probability of error than any of the codes we have considered when the number of message digits is sufficiently large.[11] This suggests that further studies of such multiple-error-correcting codes would be profitable. It would also be of interest to calculate the quantity

$$P(n, k, \alpha) = 1 - \sum_{i=0}^{k}\binom{n}{i}p^i(\alpha)q^{n-i}(\alpha), \tag{16}$$

the residual error in sequences of n binary digits after all errors up to order k have been corrected. This could be used to determine the limiting number of check digits that can be profitably used (i.e., in comparison with a known code) at constant data-rate with a k-error-correcting code.

V. The Efficiency of the Wagner Code

Until now we have compared the performance of the Wagner code with other codes only on a constant-data-rate (fixed source entropy) basis. From the standpoint of information theory, a natural measure of the efficiency of a code is the amount by which the signal-to-noise ratio needed to achieve its signaling rate exceeds the signal-to-noise ratio needed to achieve the same rate with ideal coding. We now calculate this excess signal to noise ratio for the Wagner code and compare it with the corresponding quantities for other codes as presented in a paper by Gilbert.[12] We use Gilbert's notation in the discussion that follows.

Suppose the transmitted alphabet comprises K words (waveforms) determined by D sample values $1/(2W)$ seconds apart. The information rate per unit bandwidth, assuming negligible equivocation, is

$$\frac{R}{W} = \frac{2\log K}{D}. \tag{17}$$

The signaling rate for Wagner-coded sequences of D binary pulses is, by (17),

$$\frac{R}{W} = 2\,\frac{D-1}{D}, \tag{18}$$

since one of the D pulses is a check digit. Gilbert's criterion for negligible equivocation is that the average probability of error per word is

$$P = 10^{-5}\log_{10} K. \tag{19}$$

(This convention corresponds to an average of one error per ten thousand ten-decimal-digit "telephone numbers.") The value of α [see (2)] required to achieve this error rate using the Wagner code is, by (I-29) and (19), the solution of

$$10^{-5}(D-1)\log_{10} 2 = 1 - q^D(\alpha) - \Pi_D(\alpha). \tag{20}$$

The corresponding value of signal-to-noise ratio is just

$$Y = \frac{V^2}{N_0W} = 2\alpha^2. \tag{21}$$

The signaling rate for ideal coding is, by a well-known theorem of Shannon,

$$\frac{R}{W} = \log(1+Y). \tag{22}$$

We find the value of Y for ideal coding at the signaling rate given by (18) and find ΔY by subtracting it from the value of Y for the Wagner code, obtained from (20) and (21). (All values of Y are in decibels.) This ΔY is the measure of code efficiency referred to above. Table III, on the following page, shows values of ΔY for Wagner-coded sequences of various lengths.

The closest approach to ideal coding occurs at about $D=25$, where the code requires 6.56 db more power than the ideal for the same information rate. Moreover, the Wagner code is almost as efficient over a wide range of values of D. This is remarkable in the light of Gilbert's conclusion that "one cannot signal over a channel with signal to noise level much less than 7 db

[8] D. E. Muller, "Metric Properties of Boolean Algebra and Their Application to Switching Circuits," Report No. 46, Digital Computer Lab., Univ. of Illinois; April, 1953.
[9] I. S. Reed, "A class of multiple-error-correcting codes and the decoding scheme," *Trans. IRE*, PGIT-4; pp. 38–49; September, 1954.
[10] M. Plotkin, "Binary Codes with Specified Minimum Distance," Research Division Report 51-20, Moore School of Elec. Eng., Univ. of Penna., January, 1951.
[11] R. A. Silverman and M. Balser, "Coding for constant-data-rate systems," *Trans. IRE*, PGIT-4, pp. 50–57; September, 1954.
[12] E. N. Gilbert, "A comparison of signaling alphabets," Bell Sys. Tech. Jour., vol. 31, pp. 504–522; May, 1952.

TABLE III

Efficiency of Various Wagner Codes

Length of coded sequence, D	Information rate, R/W(bits/unit bandwidth)	Signal-to-noise ratio, Y (db)	Signal-to-noise ratio in excess of ideal coding, ΔY (db)
5	1.60	10.44	7.36
11	1.82	10.73	6.70
19	1.89	10.92	6.58
25	1.92	11.01	6.56
33	1.94	11.10	6.58
∞	2.00	13.11	8.34

above the ideal level of (22) without using an unbelievably complicated alphabet."[12] The only code considered by Gilbert which is more efficient than the Wagner code is a ternary code devised by Slepian which requires 6.23 db over the ideal.

It is of interest to note that the point $D=\infty$ corresponds to the case considered by Oliver, Pierce, and Shannon,[13] i.e., "uncoded" PCM. The fact that they obtained a value of $\Delta Y=9.2$ db (8.3 times the power) rather than 8.3 db is due to a slightly different criterion for negligible equivocation. The Wagner code improvement of about 1.8 db over the uncoded case is the same using their criterion.

VI. Conclusion

In this paper we consider further aspects of the Wagner code introduced in Part I. Though the Wagner code is not a code in the usual sense, we regard the phrase "Wagner code" as a convenient abbreviation. We show in Appendix B that the Wagner code is just the minimum probability of error detector for parity checked sequences of independent binary digits.

The performance of the Wagner code is evaluated in two ways. First, in Part I, we compare Wagner-coded sequences (words) with uncoded and Hamming-coded sequences of the same number of message digits, under the assumption of constant-data-rate (fixed information rate at the channel input regardless of the coding adopted). The fact that the probability of error of short Wagner-coded words is on this basis less than the probability of error of Hamming-coded words with the same number of message digits is due to its economy in the use of check digits (only one parity check). Obviously, if the constraint of constant-data-rate is removed, the Hamming code corrects more errors (all single errors) than the Wagner code (only some single errors, albeit most of them in low noise). The second way we evaluate the performance of the Wagner code is by comparing it with ideal coding, as Gilbert has done for a variety of other codes. This comparison is made in Section V, where it is found that the Wagner code is almost as efficient as the best of far more complicated codes. This supports the intuitive notion that by making use of the magnitude of the difference between the *a posteriori* probabilities, the Wagner code recovers information that would otherwise be lost.

[13] B. M. Oliver, J. R. Pierce, and C. E. Shannon, "The philosophy of PCM," Proc. I.R.E., vol. 36, p. 1324–1331; November, 1948.

In Sections II and III the principle of the Wagner code is used in two multiple-error-correcting schemes. The first uses the double-error-detecting feature of the Hamming single-error-correcting, double-error-detecting code to initiate a search for the two most doubtful digits when a double error is indicated. The second consists of dividing long sequences into separately Wagner-coded subsequences. In this way, many multiple errors (those in which only single errors occur in the subsequences) can be corrected. Of course, such a subdivision into separately coded subsequences can be set up using any error-correcting code, but the Wagner code, because of its economy in the use of check digits, is especially suited for such subdivision.

In Section IV we refer to sample calculations which suggest strongly that algebraic multiple-error-correcting codes of the type recently developed by several authors are most suitable at constant-data-rate if the number of message digits is rather large.

In Appendix A we present a much simpler model of a constant-data-rate system than the correlation detection scheme of Part I, namely, the model of binary pulses of bandwidth W.

Appendix A

A Simple Pulse Model

It is asserted in the introduction that there is a complete correspondence between the correlation-detection scheme of Part I and the much simpler model of binary pulses of bandwidth W. We now exhibit this correspondence.

Consider two pulses centered at $t=0$:

$$x_1(t) = V\frac{\sin 2\pi Wt}{2\pi Wt} \quad \text{and} \quad x_2(t) = -V\frac{\sin 2\pi Wt}{2\pi Wt}. \qquad (23)$$

Let the channel noise be additive white Gaussian noise with power N_0 per unit bandwidth, i.e., the probability density of the noise is

$$W(n) = \frac{1}{\sqrt{2\pi N_0 W}}\, e^{-n^2/2N_0W}. \qquad (24)$$

If x_1 is sent, the received signal $y=V+n$ is a Gaussian variable with mean V and variance N_0W. An error occurs if $y<0$. Thus there is a formal analogy between the correlation and pulse cases if for Δz, Δc and σ we substitute y, V and $\sqrt{N_0W}$, respectively. In particular, the per digit probability of error is

$$\int_V^\infty W(n)dn = \tfrac{1}{2}(1-\operatorname{erf}\alpha), \qquad (25)$$

where now α is given by (2). (In the correlation detection case, Δz is the difference between two Gaussian correlator outputs and a proof is required to show that Δz is itself Gaussian (see Appendix of Part I); in the pulse case, the fact that y is Gaussian follows directly from the fact that the noise is Gaussian).

If $m+k$ pulses are compressed into an interval originally containing only m pulses, the bandwidth needed to accommodate these $m+k$ pulses is $(m+k)m^{-1}W$. The total noise power is now $(m+k)m^{-1}N_0W$, so that the new value of α is

$$\sqrt{\frac{m}{m+k}}\,\frac{V}{\sqrt{2N_0W}}.$$

The first factor is just the one introduced in the correlation-detection case to account for shortening the digits. Thus the analogy between the two cases is complete.

Appendix B

The Wagner Code as a Detection Scheme

Consider all possible sequences, X_i, of $m+1$ binary digits of which 2^m (half) are allowed messages. Let the sequences be transmitted over a noisy channel. Upon reception of a perturbed sequence Y the probability $P(X_i/Y)$[14] may be computed for each sequence X_i if the *a priori* probabilities $P(X_i)$ and the noise statistics are known. Let us assume that we have a receiver which computes these *a posteriori* probabilities and picks out the allowed message with the largest *a posteriori* probability. By definition, this receiver is a minimum probability of error detector (or ideal observer), since the choice of any other message would result in a larger probability of error. It is easily seen that the Wagner code is a minimum probability of error detector when the allowed messages are sequences of independent digits, all with the same parity. For, since the digits are independent, the total probability is the product of the probabilities of the individual digits, so that either the sequence of individually more probable digits is an allowed message or, if not, the sequence with the digit changed which has the least difference between its *a posteriori* probabilities is an allowed message.

If the receiver computes instead $P(Y/X_i)$ for each sequence (it may, for instance, not know the *a priori* probabilities $P(X_i)$), and picks out the maximum of these, we call it a maximum likelihood detector. We could have defined the Wagner code to operate in this way, but we wanted a code with a low probability of error and these two detectors are not, in general, equivalent. For the only case we computed, the *a priori* probabilities are equal, and the two detectors become equivalent.

It should be noted that a minimum probability of error detector (and, of course, a maximum likelihood detector) can be defined for any set of 2^m allowed sequences of the 2^{m+1} possible sequences. It can be shown that the symmetry given by our choice of sequences with the same parity results in the lowest minimum probability of error.

In defining the operation of the Wagner code, we need make no restrictions as to either the representation of the binary digits or the nature of the perturbing noise, provided that the receiver can compute the *a posteriori* probabilities $P(X_i/Y)$. In this paper we have concerned ourselves exclusively with the case in which the binary digits are low-pass electrical signals of bandwidth W and the perturbing noise is Gaussian with constant spectral density over the band. Another application of the Wagner code is to the transmission of binary digits with unequal *a priori* probabilities over a discrete channel, with probability p_{01} of changing a 0 to a 1 and p_{10} of changing a 1 to a 0.

[14] By $P(X_i/Y)$ we mean the probability that if Y is received, X_i was sent.

Acknowledgment

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Correction

V. H. Rumsey, author of the paper, "Part I—Transmission Between Elliptically Polarized Antennas," which appeared on pages 535–540 of the May, 1951 issue of the Proceedings of the IRE, has brought the following correction to the attention of the editors.

(β) should be replaced by ($-\beta$). The direction of the x-axis in Figs. 2 and 3 should be reversed.

The Excitation of Circular Polarization in Microwave Cavities*

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Summary—**The usefulness of exciting a single circularly-polarized mode in a microwave cavity is indicated. A matrix method is presented then for the analysis of the operation of various components, such as transition pieces and differential phase shifters on waves propagating in waveguides with two degenerate orthogonal modes. This method is applied to describe three distinct systems for generating the desired circular polarization in a cavity that is coupled directly to the side wall of a waveguide. The final system, which involves a minimum of critical adjustments, is described in more detail, and its performance is indicated. It is shown that purely circular radiation can be set up along the axis of the cavity. However, the largest preponderance of one rotating mode over the other which can be set up, if one averages over the entire cavity volume, is 11.5 to 1. This value is obtained with the TE_{11n} modes.**

Introduction

A NUMBER OF microwave experiments and applications require, or would at least be facilitated by, the excitation of a single circularly-polarized mode in a resonant cavity. Examples from existing work are (1) determination of ΔM_J selection rules in microwave paramagnetic resonance transitions,[1] and (2) measurement of the complex permeability tensor of ferrite materials as a function of the static magnetic field.[2] A possible future application would be in an oscillator magnetically tuned with a ferrite-loaded cavity. In all such applications, it is necessary to apply a magnetic field of the order of kilogauss along the axis of symmetry of the cavity. Since this requires that the cavity be placed between the polepieces of a magnet, a simple end-on feed to the cavity, from a waveguide in which a circularly-polarized wave is propagating, is impossible unless a hole is bored along the axis of the polepiece, or unless the gap is made wide enough for bends which do not destroy the circularity of incident radiation. These considerations indicate the desirability of a system for exciting a single circularly-polarized mode by coupling directly from the side of a piece of waveguide which is inserted in the gap. Several such systems are described in this paper.

The essential problem is that of establishing a purely rotating microwave H field at a point on the wall of the waveguide. If this is done, a single rotating mode may be excited in a cylindrical cavity with two degenerate modes in the following manner. The cavity is attached to the wall of the waveguide in such a position that the center of an end wall lies at the point where the microwave field is rotating in the guide, and coupling is accomplished by means of a circular hole in the wall at that point. Our theoretical treatment is to first order, that is, we neglect effect of aperture on waves in the guide. In this approximation electric field must be normal to the wall, and hence it could not be used to provide a rotating electric field in the plane of the hole.

To demonstrate the operation of the systems to be described and to show how other such systems may be studied, we first present a matrix algebraic technique for concisely analyzing the behavior of microwaves in a (square or round) waveguide with two degenerate and orthogonal modes of propagation. It is well-known that waves that are circularly polarized about the axis of the guide may be set up in such a waveguide by simply retarding the phase of one component of a wave by $\pi/2$ with respect to an equal orthogonal component.[3] Our problem of establishing at the wall a field circularly polarized about an axis normal to the wall is sufficiently complex to justify setting up this formal technique for quantitative analysis of various configurations suggested by more qualitative considerations.

Matrix Method for Bimodal Waveguide

In this method we follow the scattering matrix approach.[4] The fields are expanded in terms of waves traveling along the guide (z direction) and polarized along x and y, respectively. Suppressing the universal transverse spatial dependence and the $e^{i\omega t}$ time dependence, we have

$$\vec{E}_{\rm I} = e^{-i\beta z}\vec{u}_y$$
$$\vec{E}_{\rm II} = e^{-i\beta z}\vec{u}_x, \qquad (1)$$

where the $\vec{u}$'s are unit vectors, β is the propagation constant $2\pi/\lambda_g$, and z is the distance along the guide from any convenient reference plane. The waves propagating

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[1] M. Tinkham and M. W. P. Strandberg, "Theory of the fine structure of the molecular oxygen ground state and the measurement of its paramagnetic spectrum," *Phys. Rev.* (in press). More details are given in the Ph.D. Dissertation of M. Tinkham, Dept. of Physics, MIT; 1954.

[2] J. O. Artman and P. E. Tannenwald, "Measurement of permeability tensor in ferrites," *Phys. Rev.*, vol. 91, p. 1014; 1953. H. G. Beljers and J. L. Snoek, "Gyromagnetic phenomena occurring with ferrites," *Phillips Tech. Rev.*, vol. 11, pp. 313–322; 1950.

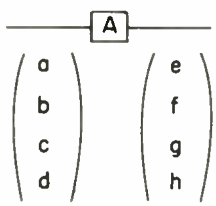
[3] J. R. Eshbach and M. W. P. Strandberg, "Apparatus for Zeeman effect measurements on microwave spectra," *Rev. Sci. Instr.*, vol. 23, p. 623; 1952.

[4] C. G. Montgomery, R. H. Dicke, and E. M. Purcell, "Principles of Microwave Circuits," MIT Radiation Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., especially ch. 5 and 10; 1948. As discussed further, our conventions in setting up the matrixes differ from those used in this reference.

in the reverse (left) direction would then be given by $\vec{E}_{I}^{*} = \vec{E}_{III}$ and $\vec{E}_{II}^{*} = \vec{E}_{VI}$. Thus the field in any region of undistorted guide may be expressed by four complex numbers giving the amplitude and phase of these four waves. These numbers are conveniently collected in a column vector, with their order defined as follows:

$$\begin{pmatrix} c_{I} \\ c_{II} \\ c_{III} \\ c_{IV} \end{pmatrix}$$

Vectors in various regions of guide will be related by matrix operators that represent various types of elements introduced. These matrixes are defined to have the property they yield the coefficients of the four outgoing waves when matrix is multiplied into a column vector composed of coefficients of incoming waves from both sides. To illustrate, if the situation is



the matrix of A must satisfy the matrix equation

$$A \begin{pmatrix} a \\ b \\ g \\ h \end{pmatrix} = \begin{pmatrix} e \\ f \\ c \\ d \end{pmatrix}$$

The matrixes of various devices are readily tabulated by considering simple cases. To handle a complicated system, one then sets up a group of these equations, eliminates the undesired unknowns which describe the waves inside the unit, and one is left with a new matrix which describes the more complex system.

Let us now list a few typical examples. If we introduce an element $\boxed{\Phi}$ which produces a phase shift ϕ and a transmission factor α (depending on the mode), but no reflection or mode conversion, the matrix is simply

$$\Phi = \begin{pmatrix} \alpha_{I}e^{i\phi_{I}} & 0 & 0 & 0 \\ 0 & \alpha_{II}e^{i\phi_{II}} & 0 & 0 \\ 0 & 0 & \alpha_{I}e^{i\phi_{I}} & 0 \\ 0 & 0 & 0 & \alpha_{II}e^{i\phi_{II}} \end{pmatrix}$$

We note that this matrix reduces properly to the unit matrix in the limit of no attenuation or phase shift.[5] Such a device $\boxed{\Phi}$ may be made by inserting dielectric, conductive, or lossy vanes into the guide in one or the other of the planes of polarization. The quarter-wave pipe described[4] for generating a circularly polarized traveling wave is an example of such an element with $\phi_{I} - \phi_{II} = \pi/2$.

A short $\boxed{S}$ converts waves propagating to the right to ones propagating to the left with a phase dependent on the location of the short. Since the two polarizations may be shorted independently by orthogonal vanes at z_{I} and z_{II}, respectively, the general matrix is

$$S = \begin{pmatrix} 0 & 0 & -e^{i2\beta z_{I}} & 0 \\ 0 & 0 & 0 & -e^{i2\beta z_{II}} \\ -e^{i2\beta z_{I}} & 0 & 0 & 0 \\ 0 & -e^{i2\beta z_{II}} & 0 & 0 \end{pmatrix}.$$

The factor β is here the same for both modes, since we are considering the case with degenerate modes.

If we introduce a "Babinet compensator" $\boxed{B}$ (in the form of a section of guide squeezed along the diagonal,[6] we may consider the wave expanded in terms of ξ and η components along the diagonals at 45 degrees to the x, y waves. These waves now propagate at different speeds because of the distortion that lifts the degeneracy. At the end of the distorted section having a length l, we re-expand in the original basic waves. Neglecting reflection and the mean phase shift, exp $[i\frac{1}{2}(\beta_{\eta}+\beta_{\xi})l]$, the matrix is

$$B = \begin{pmatrix} \cos\delta & i\sin\delta & 0 & 0 \\ i\sin\delta & \cos\delta & 0 & 0 \\ 0 & 0 & \cos\delta & i\sin\delta \\ 0 & 0 & i\sin\delta & \cos\delta \end{pmatrix},$$

where δ is half the differential phase shift $\frac{1}{2}(\beta_{\eta}-\beta_{xi})l$; e.g., $\delta = \pi/4$ for a quarter-wave plate.

If we consider an ideal transition piece $\boxed{T}$ which couples between two pieces of guide with axes rotated by an angle θ, the matrix is easily seen to be

$$T = \begin{pmatrix} \cos\theta & \sin\theta & 0 & 0 \\ -\sin\theta & \cos\theta & 0 & 0 \\ 0 & 0 & \cos\theta & -\sin\theta \\ 0 & 0 & \sin\theta & \cos\theta \end{pmatrix}$$

where again we neglect reflection. The turnstile coupler discussed in a later section is a special form of such a device, with $\theta = +45$ degrees and -45 degrees for the two side arms. This matrix also represents the effect of rotating the reference axes in circular waveguide through an angle θ.

Application of the Method to Circular Polarization

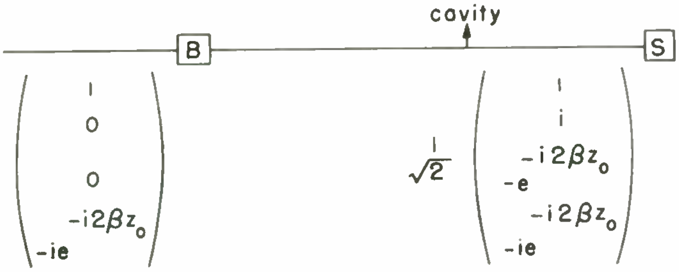
In this section we shall give three arrangements in which the components described above may be used to provide the proper phase and amplitude relations to

[5] This desirable reduction does not hold with the conventions used by Montgomery, et al. (reference 4), since their matrixes differ from ours by an exchange of the upper and lower two rows. A more important advantage of our convention is that, by using wave coefficients rather than terminal voltages as the basis for the matrixes, we obtain a cleaner representation in the guide of the fields which are actually coupling into the cavity.

[6] For example, Eshbach and Strandberg, *loc. cit.*

excite circularly-polarized radiation in a cavity coupled through a hole in the sidewall of the waveguide.

The first arrangement one might try in analogy to the problem of generating circular polarization in waveguides[3] would be a Babinet compensator set as a quarter-wave plate with a movable short at the end of the line to aid in matching into the cavity. Assuming that purely vertically-polarized (I) radiation is incident, and that the two polarizations are shorted at the same point, $z_I = z_{II} = z_0$, our matrix method yields the fields indicated by the column vectors:



We note that purely horizontally-polarized (IV) radiation is reflected. From the vector at the right, we may then construct the following z-dependence of the electric field in the region where the cavity is coupled:

$$\vec{E} = \sqrt{2}e^{-i\beta z_0} \sin \beta(z - z_0)(\vec{u}_x - i\vec{u}_y).$$

From this expression, the z-dependence of H at the wall is easily shown to be proportional to

$$\vec{H} = e^{-i\beta z_0}[\zeta \sin \beta(z - z_0)\vec{u}_z - i \cos \beta(z - z_0)\vec{u}_x], \quad (2)$$

where $\zeta = (H_z)_{\max}/(H_x)_{\max}$. For this to be purely circular, we then must choose the position of the short (z_0) with respect to the coupling hole (z) so that $\cot \beta(z - z_0) = \pm\zeta$.

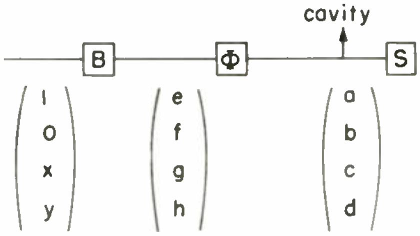
Reference to the forms of the waveguide modes[7] shows that $\zeta = \lambda_g/\lambda_c$ for square waveguide, and $1.841\ \lambda_g/\lambda_c$ for circular waveguide (operating in the lowest mode in both cases). It is also useful to note that

$$\lambda_g/\lambda_c = [(\lambda_c/\lambda)^2 - 1]^{-1/2},$$

where the cut-off wavelength λ_c is given by $2a$ for rectangular waveguide (a is the width) and by $3.41R$ for circular waveguide (R is the radius). Pertinent examples are that $\zeta = 1$ for 0.90×0.90 inch (I.D.) square guide at 9,300 mc and $\zeta = 2.38$ for 0.94 inch (I.D.) circular guide at the same frequency. Thus for these typical dimensions, the short should be distant by an odd multiple of $(\lambda_g/8)$ from the hole with the square guide, and displaced approximately $(\lambda_g/16)$ from these points for the circular guide. Although this method was successfully used in an early model, it suffers from the disadvantage that the coupling fields at the hole have only $1/\sqrt{2}$ their maximum values and are therefore changing rapidly with position. This makes the positioning for exact circularity critical.

[7] H. R. L. Lamont, "Wave Guides," Methuen and Co., Ltd., London, Eng.; 1949.

One would expect to obtain less critical performance with the addition of a differential phase shifter $\boxed{\Phi}$ to achieve independent phase and amplitude control. This arrangement is



For a quantitative investigation, we apply the matrix method, assuming vertically polarized incident radiation but leaving all other parameters general. After algebraic elimination of the unwanted variables $efgh$, we find that the field in the region of the cavity is given by

$$\begin{pmatrix} a \\ b \\ c \\ d \end{pmatrix} = \begin{pmatrix} e^{i\phi_I} & \cos\delta \\ ie^{i\phi_{II}} & \sin\delta \\ -e^{i(\phi_I - 2\beta z_I)} & \cos\delta \\ -ie^{i(\phi_{II} - 2\beta z_{II})} & \sin\delta \end{pmatrix}.$$

From these coefficients we may construct the E and H fields as in the previous example. The result for H is

$$\vec{H} = \zeta \sin\delta \sin\beta(z - z_{II})e^{i(\phi_{II} - \beta z_{II})}\vec{u}_z - i \cos\delta \cos\beta(z - z_I)e^{i(\phi_I - \beta z_I)}\vec{u}_x. \quad (3)$$

The strength of the matrix method is the ease with which it yields these general expressions which allow the full capabilities of an arrangement to be seen at once. Inspecting (3) we see that the conditions for circularity are

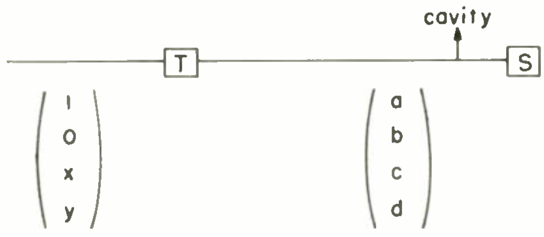
$$\phi_{II} - \beta z_{II} = \phi_I = \beta z_I$$

and

$$\zeta \sin\delta \sin\beta(z - z_{II}) = \cos\delta \cos\beta(z - z_I).$$

These may be satisfied by choosing $\phi_I - \phi_{II} = \pm\pi/2$, $z_I - z_{II} = \pm\lambda_g/4$, and $\delta = \pm\cot^{-1}\zeta$. This choice allows the coupling window to be at the maxima of both coupling fields. The positioning is noncritical and a smaller coupling hole may be used. To carry out this choice of parameters, the two modes I and II are shorted a quarter-wave apart by orthogonal vanes, and the two varied together to the point of strongest coupling. The degree of phase shift required in the squeeze section $|B|$ depends on the ratio ζ. For our example of the square guide with $\zeta = 1$, $\delta = \pm\pi/4$, and the device acts as a quarter-wave plate. For the circular guide example, however, $\delta = \pm 22.8$ degrees. In either case, the major fraction of the reflected power, $y^2 = (2\zeta/1 + \zeta^2)^2$, is in the IV mode. An apparatus of this type also was successfully tested, but the problem of coupling signal out of IV mode and adjustment of $\boxed{B}$ and $\boxed{\Phi}$ are troublesome.

With the experience now obtained it will be realized that, essentially, one desires independent amplitude and phase control of the two modes. Simple amplitude control is obtained from a rotation section $\boxed{T}$. The orthogonal shorting vanes used in the second case (above) provide a simple method of phase control. These considerations lead to the final model, which may be symbolized as shown:



Following the same procedure as above, we find

$$\vec{H} = \zeta \sin\theta \sin\beta(z - z_{II})e^{-i\beta z_{II}}\vec{u}_z - \cos\theta\cos\beta(z - z_I)e^{-i\beta z_I}\vec{u}_x. \quad (4)$$

The conditions for circularity are $z_I - z_{II} = \pm\lambda_g/4$, and $\theta = \cot^{-1}\zeta$. The shorts are located so that $z - z_I = z - z_{II} \pm\lambda_g/4 = n\lambda_g/2$ at the hole, giving maximum and noncritical coupling. For these settings, the values of $|x|$ and $|y|$ are $(1-\zeta^2)/(1+\zeta^2)$ and $2\zeta/(1+\zeta^2)$, respectively. Thus for ζ near unity, the greater share of the reflected signal is again in the IV mode. This is no problem, however, with the turnstile transition coupler of the sort to be described. Also, the settings of θ and $z_I - z_{II}$ are simple mechanical properties, and the more subtle and troublesome squeeze and phase shift setting of the previously described method are eliminated.

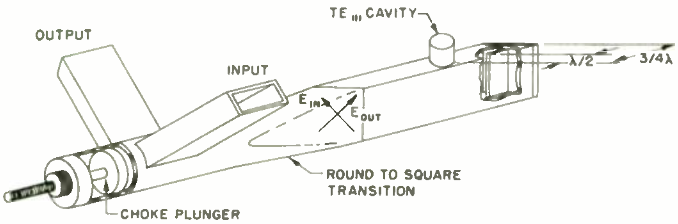


Fig. 1—Schematic diagram of transition coupler for circular polarization.

The Final Device

In the previous section we have used a matrix formalism to indicate several methods for providing circular excitation to a cavity. We now describe the final device, shown in Figs. 1 and 2, in a more concrete manner and indicate its operation in nonmathematical terms for the simple case of a square guide with $\zeta = 1$.

The incident wave propagates down the square guide, polarized along a diagonal. At the far end, the vertical component is shorted by a conducting vane a quarter-wave in front of the end plate, which shorts the horizontal component. The reflected components combine to give a wave polarized along the other diagonal which leaves through the other arm. This output arm is oriented at exactly 90 degrees to the input arm to avoid direct cross-coupling. (The choke plunger is adjusted for optimum matching of the input and output arms. By symmetry, the same setting is best for both.) The incident and reflected waves combine to set up a standing-wave pattern in which the vertically- and horizontally-polarized waves are 90 degrees out-of-phase (in space and time) because of the $\lambda_g/4$ difference in path lengths to the effective shorting position. The round coupling hole to the cavity is located on the center line of the wall of the square guide at such a distance from the end that it is at the maximum of both the longitudinal H of the horizontally-polarized standing wave and the transverse H of the vertically-polarized standing wave. Since ζ is assumed equal to 1, the two components are of equal intensity and 90 degrees out-of-phase. Thus we succeeded in producing a circularly-polarized radiation field at the window that couples into the cavity.

The circular excitation will not give circular radiation in the cavity, though, unless the cavity has two degenerate orthogonal modes differing only by a 90-degree rotation about the axis of the cavity. This will be the case with circular or square TE_{lmn} or TM_{lmn} modes. The two modes then may be considered to be excited independently and 90-degrees out-of-phase, with a circularly-polarized radiation field as the result. Along the axis, the field will rotate purely in one direction. However, averaged over the cavity, this is not true. In fact if we use TM modes, the energy, on the average, is shared equally between the two senses of rotation. That this is so may be seen qualitatively by noting that the lines of H are closed. In a plane (TM) field, this implies

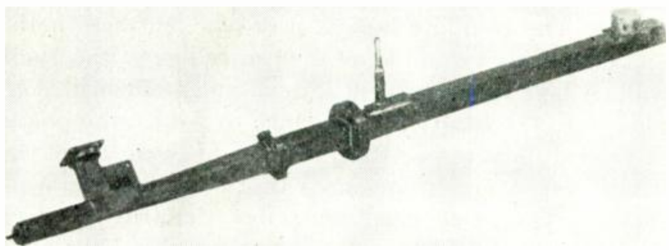


Fig. 2—Photograph of transition coupler for circular polarization.

that there is as much rotation in one sense as in the other. If we go to a TE mode, however, longitudinal H is allowed, and the loops of H can close in the axial direction. This enables us to get a net circular polarization over the cavity at the expense of acquiring some axial fields. If one is dealing with a magnetic material confined near the axis of the cavity the fields are purely circular over the sample, and these averages are of little relevance. However, if the material is a gas which fills the cavity, they are important.

To render this discussion more quantitative, we may expand field in cavity in terms of rotating unit vectors,

$$\vec{H} = H_+\left(\frac{\vec{u}_x + i\vec{u}_y}{\sqrt{2}}\right) + H_-\left(\frac{\vec{u}_x - i\vec{u}_y}{\sqrt{2}}\right) + H_z\vec{u}_z$$

and then, for example, define f_+, the fraction of the energy stores in the H_+ rotating fields, by

$$f_+ = \frac{\int H_+^2 d\tau}{\int H^2 d\tau}.$$

For TM modes, $f_+ = f_- = \frac{1}{2}$, $f_z = 0$. For TE_{lmn} modes, the fractions depend on m, only $m=1$ giving a large preponderance of one sense of circular polarization over the other. For the TE_{11n} modes, evaluation of the required integrals shows that

$$f_+ = 11.5 f_- = 0.92(1 - f_z)$$

$$f_z = 0.086\left(\frac{c}{a\nu}\right)^2 = \frac{1}{1 + 2.91(an/l)^2},$$

where c is the velocity of light, a is the radius of the cavity, l is its length, and ν is the resonant frequency. With the dimensions of the cavity actually used, $f_z = 0.26$, $f_+ = 0.68$, and $f_- = 0.06$. The ratio of 11.5 between f_+ and f_- is quite adequate to allow unambiguous results, even in this case of a cavity filled with the material under study.

A photograph of the actual apparatus is given in Fig. 2. Note the tuning screws in the cavity. They are required to balance out the effect of coupling into the waveguide, which is not equivalent in the two orthogonal directions, and to compensate for any other imperfections which destroy the exact degeneracy of the two modes. The coupling hole is precisely centered in the end of the cavity and is on the center line of the guide wall. A hole diameter of 0.25 inch gives sufficient coupling into the cavity at resonance to reduce the power at the exit arm to one-third of its value away from the cavity resonance frequency. In mounting the $\lambda_g/4$ shorting vane, one must remember that the effective shorting plane lies several millimeters behind the edge of the vane.[8] An O-ring mica seal is used to isolate the gaseous samples studied with the apparatus. To reverse the sense of rotation, the assembly is unscrewed and turned through 180 degrees at the square-flange joint visible in Fig. 2.

In the measurement arrangement for which it was actually designed, the device is used as follows. A chart recording is made of the damping of the cavity resonator by resonant absorption of the sample gas as a function of the static magnetic field. The dominant sense of rotation is then reversed and another chart is made. By contrasting the two charts one can readily distinguish which absorptions are caused by H_+, H_-, and H_z, since the strengths of the H_+ and H_- absorptions change by factors of 11.5 in opposite directions, while the H_z absorptions are unchanged. It should probably be repeated here that, for a sample which is localized on the axis of the cavity, perfect selection of H_+ or H_- is theoretically possible, and is nearly realizable in practice.

As an indication of the degree to which our cavity circular polarizer approaches ideal behavior, we quote the following results. The vswr of the input or output with optimum adjustment of the choke plunger is 1.2. The cross-coupled power is down by 20 db. The full theoretical ratio of $f_+/f_- \approx 12$ was observed with gaseous samples. Thus this simple apparatus gives quite usable performance. Reduction of the cross-coupling could probably be obtained by tapering the input arms into narrower apertures in the cylindrical section, since this would reduce the distortion of the modes of the cylinder, leaving more complete orthogonality.

If one wishes to operate at other frequencies, or if one uses circular rather than square guide, one must cope with $\zeta \neq 1$. In that case, the permanent rigid construction shown in Fig. 2 would have to be modified by the insertion of a rotatable joint in the cylindrical section to allow the angle θ in (4) to be adjusted properly. As this is a simple mechanical adjustment, it should introduce no uncertainty in operation.

[8] N. Marcuvitz, "Wave Guide Handbook," MIT Radiation Lab. Ser., McGraw-Hill Book Co., Inc., New York, N. Y., ch. 4, p. 172; 1951.

CORRECTION

A. H. Zemanian, author of the paper, "Bounds Existing in the Time and Frequency Responses of Various Types of Networks," which appeared on pages 835–839 of the May, 1954 issue of the PROCEEDINGS OF THE IRE, has brought the following correction to the attention of the editors.

The definition of settling time appearing on page 839 should be changed to read: "The settling time to ϵ, $\tau_{s\epsilon}$, is the least time beyond which the step response remains greater than $r(1-\epsilon)$ and the impulse response remains within the bounds $\pm\epsilon/C$, where ϵ is a positive quantity less than unity, r is the final value of the step response and $1/C$ is the initial value of the impulse response. This assumes that the input functions are impressed at $t=0$."

The Ultra-Bandwidth Finline Coupler*

S. D. ROBERTSON†, SENIOR MEMBER, IRE

Summary—**The finline coupler is a recently developed microwave circuit element with which it has been possible to assemble hybrid junctions, directional couplers, and polarization-selective couplers capable of operating over bandwidths of at least three to one in frequency. Constructional details and experimental results are given.**

THE ACCELERATED development in the past few years of modern communication technology has been characterized by two readily discernible patterns, the progression to higher and higher frequencies in the spectrum, and the corollary demand for greater bandwidths. The requirements of the latter have been met in part by the development of traveling-wave amplifiers and backward-wave oscillators capable of operating over enormous bandwidths of the order of two to one in frequency. It would appear that a point has been reached where further development in the direction of increasing bandwidths is being inhibited by the lack of sufficiently wide-band microwave circuit components, such as directional couplers, hybrid junctions, and waveguide bends. It is the purpose of the present paper to describe the finline coupler, a new microwave circuit element, in a form, evolved jointly by H. T. Friis and the author, with which it has been possible to design hybrids, directional couplers, and polarization-selective couplers capable of operating over bandwidths of at least three to one in frequency.

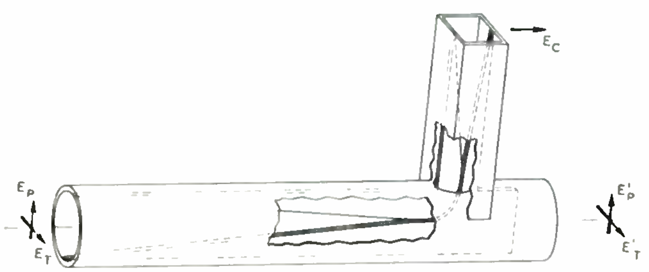


Fig. 1—Basic finline coupler.

The basic finline coupler is shown in Fig. 1. The device consists, in this particular case, of a length of circular waveguide fitted with a pair of diametrically opposite, thin fins that taper in from the outer wall of the guide until their opposing edges are separated by a small gap at the center. Thus, substantially all of the energy associated with the electric field E_p (where the subscript p denotes that the vector is parallel to the plane of the fins) is concentrated from the dominant mode of propagation in the circular waveguide to a finline mode in which the energy is largely confined to the gap and its immediate vicinity. This energy may then be removed from the circular guide by curving the finline and bringing it out through a small hole in the side wall. It may then be launched into another waveguide as shown.

On the other hand, a wave characterized by the transverse field E_T will pass on through the guide, relatively undisturbed by the presence of the fins, and will emerge as E_T'.

One can readily see that the finline coupler offers a means whereby it is possible to separate two waves perpendicularly polarized to one another. The fact that this is done with smooth tapers several wavelengths long suggests that the coupler ought to work over very wide bands. Such has been found to be the case.[1]

If only one wave is present in the guide, it is found that one may abstract any desired proportion of its energy by simply rotating the coupler about its axis so that the fins are inclined with respect to the plane of polarization of the wave. The abstracted field will then be proportional to the cosine of the angle.

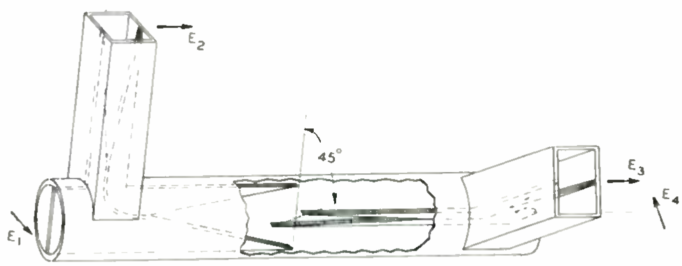


Fig. 2—Ultra-broad-band hybrid junction using two finline couplers.

Fig. 2 shows how two finline couplers may be arranged to form a hybrid junction. When the planes of the fins are inclined at an angle of 45 degrees, as shown in the figure, one obtains a 3 db hybrid. A wave entering at E_1 passes through the left-hand coupler without modification. Upon entering the second coupler, however, it is split into two equal components, one of which emerges at E_3, and the other emerges at E_4. Likewise, a wave entering E_2 travels through the first coupler and is

* Original manuscript received, March 16, 1955.
† Bell Telephone Labs., Holmdel, New Jersey.

[1] It may be noted that the finline coupler bears a superficial resemblence to the ridged waveguide-to-coaxial transitions described by S. B. Cohn, "Design of simple broad-band waveguide-to-coaxial-line junctions," PROC. IRE, vol. 35, p. 926; September, 1947. Attention is called, however, to the fact that the particular geometry of the finline coupler permits it to be used as an eight-terminal circuit element whose four pairs of terminals are defined by E_p, E_c, E_T, and E_T'. The latter two pairs are not permitted by the geometry of Cohn's transducer. In order to obtain the additional two pairs, it is necessary to use very thin fins and to select a waveguide cross section the cutoff frequency of which will allow the transmission of a wave polarized in the plane of E_T.

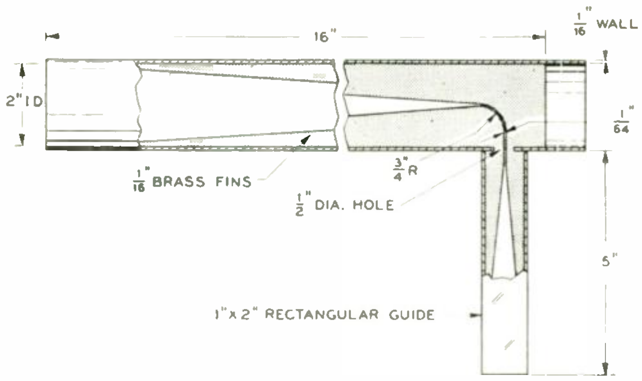


Fig. 3—Dimensions of experimental finline coupler.

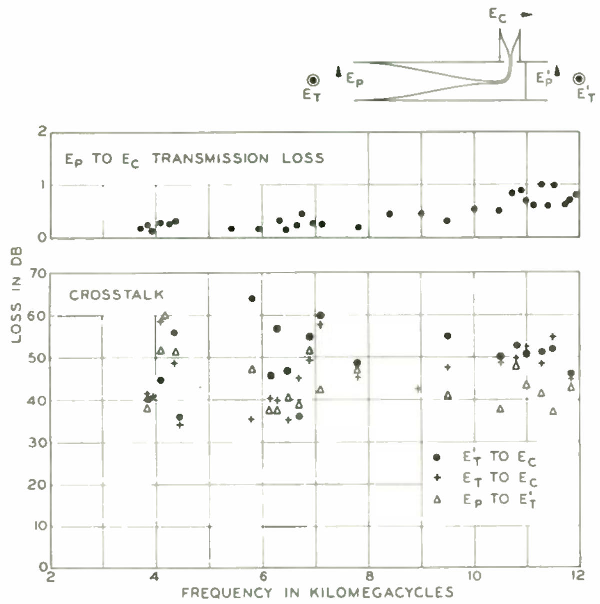


Fig. 4—Transmission properties of a finline coupler.

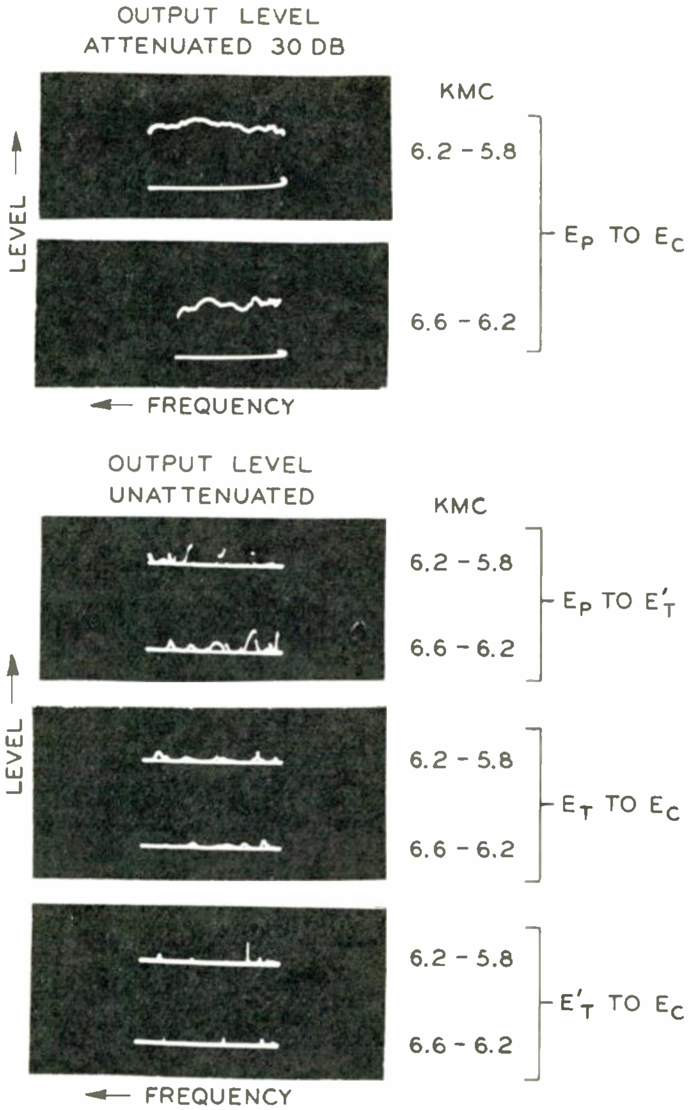


Fig. 5—Oscillographic traces of the transmission characteristics of the coupler in the 6 kmc band.

split by the second coupler into two equal components. Degrees of coupling other than 3 db may be obtained by inclining the fins at angles other than 45 degrees.

In order to learn something of the performance of the coupler, an experimental model was built having the dimensions shown in Fig. 3. The dimensions selected were obtained largely by guesswork, and the successful results obtained with this coupler only serve to point out the fact that a truly broad-band circuit element must, by its very nature, be relatively uncritical in its dimensions. Other couplers have been made having a 1/32-inch gap between the fins and a $\frac{7}{8}$-inch hole in the side wall which performed substantially as well as the one illustrated. It may be said in general that the required hole size is proportional to the gap spacing.

No detailed study has yet been made to determine the optimum length and proportions of the fin tapers. Longer tapers would no doubt give lower standing wave ratios. The tapers are linear except at those points where they join the walls of the waveguides and where they meet the circular arcs at the turn-off. They were curved at these points to provide a more uniform change of finline impedance. The input standing-wave ratios for these tapers were less than 3 db at all measured points in the band extending from 3.75 to 12.0 kilomegacycles.

Fig. 4 is a plot of the measured transmission losses between the various terminals of the coupler. Losses measured from E_p to E_c are shown in the upper part of the figure. It will be noted that they are less than one db over the entire range of frequencies from 3.75 to 12.0 kilomegacycles. Over most of the range the losses are only a few tenths of a db. The rise in loss at the higher frequencies is believed to be due in part to higher ohmic losses (perhaps 0.2 db) and to mode conversion losses. The latter have been reduced in several cases by giving careful attention to the measuring equipment circuit components, which suggests that not all of the mode conversion losses may be due to the coupler itself, but rather to auxiliary circuit elements used in the measurement. Since one is dealing with waveguide dimensions that will support higher order modes at these frequencies, it is necessary to consider the coupler in relation to its environment of associated equipment.

Losses of the transverse wave through the coupler from E_T to E_T' were measured and found to be of the order of 0.1 db or less and are not plotted in the figure.

The points plotted in the lower parts of Fig. 4 represent three of the six possible "crosstalks" which are believed to be of particular interest. It will be noted that they are all in excess of 34 db. Other crosstalks such as E_p to E_p', E_c to E_p', and E_T to E_p' were also measured at a few frequencies and were found to be of the same order of magnitude as those plotted here. It is believed that the crosstalk discrimination can be increased substantially by maintaining closer dimensional tolerances on the coupler and by lengthening the fin tapers.

Fig. 5 (facing page) presents some oscillographic trace of the transmission characteristics given in Fig. 4, but confined to the frequency band from 5.8 to 6.6 kmc. These traces were made with a broad-band sweeper for this band developed by D. A. Alsberg.[2] They are important for showing that there are no serious "holes" in the loss measurements in this particular frequency band, which may have been missed in the earlier point-by-point measurements. To be sure, several sharp peaks are noted in the transmission levels of some of the crosstalk components, but they are still 35 db or more below the input signal. It should be pointed out that the E_p to E_c traces were obtained with an additional 30 db of rf attenuation in the circuit above that used in obtaining the crosstalk traces.

[2] This is discussed in a paper to be published in *IRE, Trans.* PGI.

Conclusion

It appears that the finline coupler may be of considerable utility as a basic ultra-bandwidth circuit element that may be used as a variable coupler, a hybrid junction, or as a polarization selector. It is likely that other applications will be found.

Acknowledgment

The author wishes to express his appreciation to his colleagues at Bell Telephone Laboratories for much helpful advice and encouragement. C. F. Chapman was of great assistance in the experimental part of the work, and H. E. Heskett, A. P. King, and R. W. Dawson very kindly made their microwave measuring equipment available for obtaining the necessary experimental data.

Correction

Mr. D. J. Nelson of the Bell Telephone Laboratories has pointed out to J. R. Macdonald, co-author of the paper, "The Charging and Discharging of Nonlinear Capacitors," which appeared on pages 71–78 of the January, 1955 issue of the Proceedings of the IRE, that the following correction should be brought to the attention of the readers.

The integral $J(\eta, \xi)$ defined in eq. 19 can be expressed in terms of exponential integrals. Equations (18), (22), and (25) can therefore be rewritten in the simplified forms

$$\tau = 1 - e^{-W} + e^{-\eta}[Ei(\eta) - Ei(x)], \tag{18'}$$

$$\tau = [\eta/2 \sinh \eta][e^{\eta}\{Ei(-\eta) - Ei(-W)\} + e^{-\eta}\{Ei(\eta) - Ei(W)\}], \tag{22'}$$

$$\tau = [\eta/2 \sinh \eta][Ei(-\eta) - Ei(-x) + Ei(\eta) - Ei(x)]. \tag{25'}$$

Tables of the exponential integral of positive argument are given in footnote 13 of the above paper.

IRE Standards on Television: Definitions of Color Terms, 1955*

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Achromatic Locus (Achromatic Region). Chromaticities which may be acceptable reference standards under circumstances of common occurrence are represented in a chromaticity diagram by points in a region which may be called the "achromatic locus."

Note—The boundaries of the achromatic locus are indefinite, depending on the tolerances in any specific application. Acceptable reference standards of illumination (commonly referred to as "white light") are usually represented by points close to the locus of Planckian radiators having temperatures higher than about 2,000°K. While any point in the achromatic locus may be chosen as the reference point for the determination of dominant wavelength, complementary wavelength and purity for specification of object colors, it is usually advisable to adopt the point representing the chroma-

* This Standard, 55 IRE 22.S1, is an expansion and revision of an earlier Standard, 53 IRE 22.S1, entitled "Standards on Television: Definition of Color Terms, Part I, 1953," which appeared in the March, 1953, issue of the Proceedings of the IRE. Reprints of 55 IRE 22.S1 are available and may be purchased while available from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y., at $0.60 per copy. A 20 per cent discount will be allowed for 100 or more copies mailed to one address.

† Deceased.

ticity of the luminator. Mixed qualities of illumination, and luminators with chromaticities represented very far from the Planckian locus, require special consideration. Having selected a suitable reference point, dominant wavelength may be determined by noting the wavelength corresponding to the intersection of the spectrum locus with the straight line drawn from the reference point through the point representing the sample. When the reference point lies between the sample point and the intersection, the intersection indicates the complementary wavelength. Any point within the achromatic locus, chosen as a reference point, may be called an "achromatic point." Such points have also been called "white points."

Black and White. See *Monochrome.*

Brightness. The attribute of visual perception in accordance with which an area appears to emit more or less light.

Note: Luminance is recommended for the photometric quantity which has been called "brightness." Luminance is a purely photometric quantity. Use of this name permits "brightness" to be used entirely with reference to the sensory response. The photometric quantity has been often confused with the sensation merely because of the use of one name for two distinct ideas. Brightness will continue to be used, properly, in nonquantitative statements, especially with reference to sensations and perceptions of light. Thus, it is correct to refer to a brightness match, even in the field of a photometer, because the sensations are matched and only by inference are the photometric quantities (luminances) equal. Likewise, a photometer in which such matches are made will continue to be called an "equality-of-brightness" photometer.

A photo-electric instrument, calibrated in foot-lamberts, should not be called a "brightness meter." If correctly calibrated, it is a "luminance meter." A troublesome paradox is eliminated by the proposed distinction of nomenclature. The luminance of a surface may be doubled, yet it will be permissible to say that the brightness is not doubled, since the sensation which is called "brightness" is generally judged to be not doubled.

Brightness Signal—Deprecated. See *Luminance Signal.*

Candle. The unit of luminous intensity. One candle is defined as the luminous intensity of 1/60th square centimeter of a blackbody radiator operating at the temperature of solidification of platinum. Values for standards having other spectral distributions are derived by the use of accepted luminosity factors.

Candlepower. Luminous intensity expressed in *Candles.*

Chroma (Munsell Chroma). The dimension of the Munsell system of color which corresponds most closely to *Saturation.*

Note: Chroma is frequently used, particularly in English works, as the equivalent of *Saturation* (*q.v.*).

Chromaticity. The color quality of light definable by its *Chromaticity Co-ordinates*, or by its *dominant* (or *complementary*) *wavelength* and its *purity* taken together.

Chromaticity Co-ordinate. The ratio of any one of the *Tristimulus Values* of a sample to the sum of the three *Tristimulus Values.*

Chromaticity Diagram. A plane diagram formed by plotting one of the three *Chromaticity Co-ordinates* against another.

Note: The most common *Chromaticity Diagram* at present is the *CIE* (x, y) diagram plotted in rectangular co-ordinates (see Fig. 1).

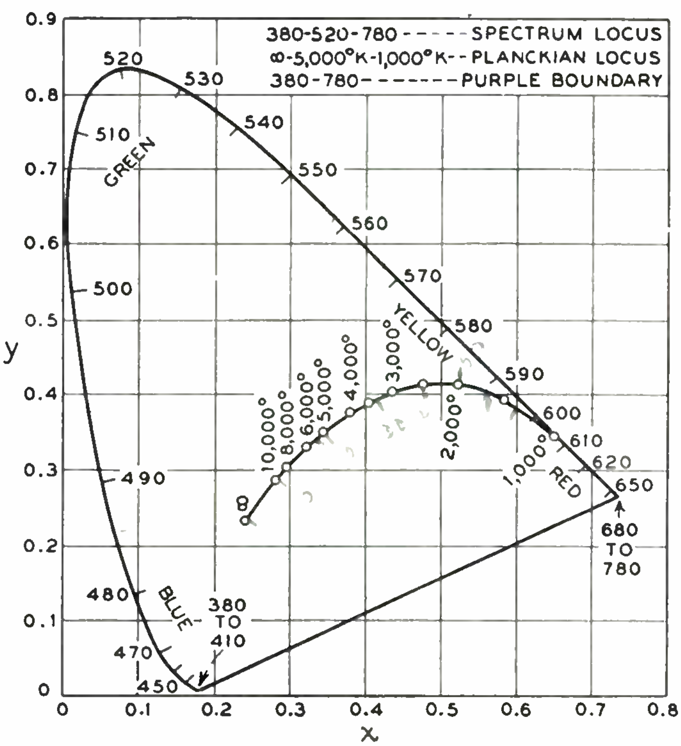

Fig. 1—Chromaticity diagram.

Chromaticity Flicker. That *Flicker* which results from fluctuation of *Chromaticity* only.

Chrominance. The colorimetric difference between any *color* and a reference *color* of equal *Luminance*, the reference *color* having a specified *Chromaticity.*

Note 1: In three-dimensional color space, *Chrominance* is a vector which lies in a plane of constant *luminance.* In that plane it may be resolved into components, called chrominance components.

Note 2: In color television transmission, for example, the *Chromaticity* of the reference color may be that of a specified *white.*

Chrominance Components. See *Chrominance.*

Chrominance Demodulator. A demodulator used in color television reception for deriving video frequency *Chrominance Components* from the *Chrominance Signal* and a sine wave of *Chrominance Subcarrier* frequency.

Chrominance Modulator. A modulator used in color television transmission for generating the *Chrominance Signal* from the video frequency *Chrominance Components* and the *Chrominance Subcarrier.*

Chrominance Primary. A *Transmission Primary* which is one of two whose amounts determine the *Chrominance* of a *Color.*

Note: Chrominance Primaries have zero *Luminance* and are nonphysical.

Chrominance Signal (Carrier Chrominance Signal). The sidebands of the modulated *Chrominance Subcarrier* which are added to the *Monochrome Signal* to convey color information.

Chrominance Subcarrier. The carrier whose modulation sidebands are added to the *Monochrome Signal* to convey color information.

CIE. Abbreviation for "Commission Internationale de l'Eclairage."

Note: These are the initials of the official French name of the "International Commission on Illumination." This translated name is approved for usage in English-speaking countries, but at its 1951 meeting the Commission recommended that only the initials of the French name be used. The initials "ICI" which have been used commonly in this country are deprecated because they conflict with an important trademark registered in England and because the initials of the name translated into other languages are different.

Coarse Chrominance Primary. In the color television system at present standardized for broadcasting in the United States, that one of the two *Chrominance Primaries* which is associated with the lesser transmission bandwidth.

Color. The characteristics of light other than spatial and temporal inhomogeneities.

Note 1: The measure of color is three dimensional. One of the many ways of measuring color is in terms of *Luminance, Dominant Wavelength,* and *Purity.*

Note: Inhomogeneities, for example, particular distributions and variations of light, and characteristics of objects which are revealed by variations such as gloss, lustre, sheen, texture, sparkle, opalescence, and transparency, are not included among the color characteristics of objects.

Color Breakup. Any fleeting and partial separation of a color picture into its *display primary* components caused by a rapid change in the condition of viewing.

Note: Illustrations of rapid changes in the condition of viewing are: (1) fast movement of the head, (2) fast interruption of the line of sight, (3) blinking of the eyes.

Color Burst. That portion of the *Composite Color Signal,* comprising a few cycles of a sine wave of *chrominance subcarrier* frequency, which is used to establish a reference for demodulating the *Chrominance Signal.*

Color Carrier. See the preferred term *Chrominance Subcarrier.*

Color Coder. In color television transmission, an apparatus for generating the *Color Picture Signal* (and possibly the *Color Burst*) from camera signals and the *Chrominance Subcarrier.*

Color Contamination. An error of color rendition due to incomplete separation of paths carrying different color components of the picture.

Note: Such errors can arise in the optical, electronic, or mechanical portions of a color television system as well as in the electrical portions.

Color Co-ordinate Transformation. Computation of the *Tristimulus Values* of *colors* in terms of one set of primaries from the *Tristimulus Values* of the same *colors* in another set of primaries.

Note: This computation may be performed electrically in a color television system.

Color Decoder. In color television, an apparatus for deriving the signals for the color display device from the *color picture signal* and the *Color Burst.*

Color-Difference Signal. An electrical signal which when added to the *Monochrome Signal* produces a signal representative of one of the *Tristimulus Values* (with respect to a stated set of *primaries*) of the transmitted *color.*

Color Encoder. See *Color Coder.*

Color Flicker. That *Flicker* which results from fluctuation of both *Chromaticity* and *Luminance.*

Color Fringing. Spurious *Chromaticity* at boundaries of objects in the picture.

Note: Color Fringing can be caused by the change in relative position of the televised object from field to field or by misregistration and, in the case of small objects, may even cause them to appear separated into different colors.

Color Match. The condition in which the two halves of a structureless photometric field are judged by the observer to have exactly the same appearance.

Note: A Color Match for the *Standard Observer* may be calculated.

Color Mixture. *Color* produced by the combination of *light* of different *colors.*

Note 1: The combination may be accomplished by successive presentation of the components, provided the rate of alternation is sufficiently high, or the combination may be accomplished by simultaneous presentation, either in the same area or on adjacent areas, provided they are small enough and close enough together to eliminate pattern effects.

Note 2: A Color Mixture as here defined is sometimes denoted as an Additive Color Mixture, to distinguish it from combinations of dyes, pigments, and other absorbing substances. Such mixtures of substances are sometimes called Subtractive Color Mixtures, but might more appropriately be called Colorant Mixtures.

Color-Mixture Data. See *Tristimulus Values*, the preferred term.

Color Picture Signal. The electrical signal which represents complete color picture information, excluding all synchronizing signals.

Note: One form of color picture signal consists of a monochrome component plus a subcarrier modulated with chrominance information.

Color Signal. Any signal at any point in a color television system, for wholly or partially controlling the *chromaticity* values of a color television picture.

Note: This is a general term which encompasses many specific connotations, such as are conveyed by the words, *Color Picture Signal*, *Chrominance Signal*, Carrier Color Signal, *Monochrome Signal* (in color television), etc.

Color Temperature. Temperature of the complete (blackbody or Planckian) radiator required to produce the same *Chromaticity* as the light under consideration.

Color Transmission. In television, the transmission of a signal wave for controlling both the *luminance* values and the *chromaticity* values in a picture.

Color Triangle. A triangle drawn on a *chromaticity diagram*, representing the entire range of chromaticities obtainable as additive mixtures of three prescribed *primaries*, represented by the corners of the triangle.

Colorimetry. The techniques for the measurement of *color* and for the interpretation of the results of such measurements.

Compatibility. That property of a color television system which permits substantially normal monochrome reception of the transmitted signal by typical unaltered monochrome receivers.

Complementary Wavelength. The wavelength of light of a single frequency, which matches the reference standard light when combined with a sample color in suitable proportions.

Note 1: The wide variety of purples which have no dominant wavelengths, including nonspectral violet, purple, magenta, and nonspectral red colors, are specified by use of their complementary wavelengths.

Note 2: Refer to *Dominant Wavelength.*

Composite Color Signal. The color picture signal plus blanking and all synchronizing signals.

Composite Color Sync. The signal comprising all the sync signals necessary for proper operation of a color receiver. This includes the deflection sync signals to which the color sync signal is added in the proper time relationship.

Constant Luminance Transmission. That type of transmission in which the *transmission primaries* are a *Luminance Primary* and two *Chrominance Primaries.*

Display Primaries (Receiver Primaries). The *colors* of constant *chromaticity* and variable *luminance* produced by the receiver or any other display device which, when mixed in proper proportions, are used to produce other *colors.*

Note: Usually three primaries are used: red, green and blue.

Distribution Coefficients. The *Tristimulus Values* of *monochromatic* radiations of equal power.

Note: Generally represented by overscored, lowercase letters, such as $\bar{x}$, $\bar{y}$, $\bar{z}$ in the *CIE* system.

Dominant Wavelength. The wavelength of light of a single frequency, which matches a color when combined in suitable proportions with a reference standard light.

Note: Light of a single frequency is approximated in practice by the use of a range of wavelengths within which there is no noticeable difference of color. Although this practice is ambiguous in principle, the dominant wavelength is usually taken as the average wavelength of the band used in the mixture with the reference standard matching the sample. Many different qualities of light are used as reference standards under various circumstances. Usually the quality of the prevailing illumination is acceptable as the reference standard in the determination of the dominant wavelength of the colors of objects.

Dot-Sequential. Pertaining to the association of the several primary colors in sequence with successive picture elements.

Examples: Dot-sequential pickup, dot-sequential display, dot-sequential system, dot-sequential transmission.

Equal-Energy Source. A light source for which the time rate of emission of energy per unit of wavelength is constant throughout the visible spectrum.

Excitation Purity (Purity). The ratio of the distance from the reference point to the point representing the sample, to the distance along the same straight line from the reference point to the *spectrum locus* or to the *purple boundary*, both distances being measured (in the same direction from the reference point) on the *CIE chromaticity diagram.*

Note: The reference points is the point in the *chromaticity diagram* which represents the reference standard light mentioned in the definition of *Dominant Wavelength.*

Field-Sequential. Pertaining to the association of individual primary colors with successive fields.

Examples: Field-sequential pickup, field-sequential display, field-sequential system, field-sequential transmission.

Fine Chrominance Primary. In the color television system at present standardized for broadcasting in the United States, that one of the two *Chrominance Primaries* which is associated with the greater transmission bandwidth.

Flicker. In television, a fluttering sensation which results from the periodic fluctuation of *light.*

Note: Flicker frequencies usually lie in the range from a few cycles per second to a few tens of cycles per second.

Footcandle. A unit of *illuminance* when the foot is taken as the unit of length. It is the *illuminance* on a surface one square foot in area on which there is a uniformly distributed flux of one *lumen*, or the *illuminance* at a surface all points of which are at a distance of one foot from a uniform source of one *candle.*

Footlambert. A unit of *Luminance* equal to $1/\pi$ *Candle* per square foot, or to the uniform *Luminance* of a perfectly diffusing surface emitting or reflecting light at the rate of one *lumen* per square foot.

Note: A *Footcandle* is a unit of incident light and a *Footlambert* is a unit of emitted or reflected light. For a perfectly reflecting and perfectly diffusing surface, the number of *Footcandles* is equal to the number of *Footlamberts.*

Frequency Interlace. In television, the relationship of intermeshing between the frequency spectrum of an essentially periodic interfering signal and the spectrum of harmonics of the scanning frequencies, which relationship minimizes the visibility of the interfering pattern by altering its appearance on successive scans.

Gamma. In television, the exponent of that power law which is used to approximate the curve of output magnitude vs input magnitude over the region of interest.

Note: For quantitative evaluation it is customary to plot the log of the output magnitude (ordinate) versus the log of the input magnitude (abscissa), as measured from a point corresponding to some reference black level, and select a straight line which approximates this plot over the region of interest and take its slope. If the plot departs seriously from linearity it cannot be adequately described by a single value of gamma. Even when the plot is reasonably linear the procedure for determining the approximation should be prescribed.

Gamma Correction. The introduction of a nonlinear output-input characteristic for the purpose of changing the effective value of *Gamma.*

Hue. The attribute of color perception that determines whether it is red, yellow, green, blue, purple, or the like.

Note 1: This is a subjective term corresponding to the psychophysical term *Dominant* (or *Complementary*) *Wavelength.*

Note 2: White, black, and gray are not considered as being hues.

ICI—Deprecated. See *CIE.*

Illuminance (Illumination). The density of the luminous flux on a surface; it is the quotient of the flux by the area of the surface when the latter is uniformly illuminated.

Lambert. A unit of *Luminance* equal to $1/\pi$ *Candle* per square centimeter, and, therefore, equal to the uniform *Luminance* of a perfectly diffusing surface emitting or reflecting light at the rate of one *lumen* per square centimeter.

Light. The aspect of radiant energy of which a human observer is aware through the visual sensations that arise from the stimulation of the retina of the eye. For the purposes of engineering, light is visually evaluated radiant energy.

Note 1: Light is psychophysical, neither purely physical nor purely psychological. Light is not synonymous with radiant energy, however restricted, nor is it merely sensation.

Note 2: The present basis for the engineering evaluation of light consists of the color-mixture data $\bar{x}$, $\bar{y}$, $\bar{z}$ adopted in 1931 by the International Commission on Illumination.

Lumen. The unit of luminous flux. It is equal to the flux through a unit solid angle (steradian) from a uniform point source of one *Candle*, or to the flux on a unit surface all points of which are at unit distance from a uniform point source of one *Candle.*

Luminance. The luminous intensity of any surface in a given direction per unit of projected area of the surface as viewed from that direction.

Note: See Note under *Brightness.*

Luminance Flicker. That *Flicker* which results from fluctuation of *Luminance* only.

Luminance Primary. That one of a set of three *Transmission Primaries* whose amount determines the *Luminance* of a *color.*

Luminance Signal. A signal wave which is intended to have exclusive control of the *Luminance* of the picture.

Luminosity. Ratio of *luminous flux* to the corresponding *radiant flux* at a particular wavelength. It is expressed in lumens per watt.

Luminosity Coefficients. The constant multipliers for the respective *Tristimulus Values* of any *color*, such that the sum of the three products is the *Luminance* of the *Color.*

Luminous Efficiency. The ratio of the *luminous flux* to the *Radiant Flux*.

Note: Luminous efficiency is usually expressed in lumens per watt of radiant flux. It should not be confused with the term "efficiency" as applied to a practical source of light, since the latter is based upon the power supplied to the source instead of the radiant flux from the source. For energy radiated at a single wavelength, luminous efficiency is synonymous with *luminosity*.

Luminous Flux. The time rate of flow of light.

Luminous Intensity (in any direction). The ratio of the *luminous flux* emitted by a source or by an element of a source, in an infinitesimal solid angle containing this direction, to the solid angle.

Note: Mathematically, a solid angle must have a point at its apex; the definition of Luminous Intensity, therefore, applies strictly only to a point source. In practice, however, Light emanating from a source whose dimensions are negligible in comparison with the distance from which it is observed may be considered as coming from a point.

Matrix (noun). In color television, an array of coefficients symbolic of a *color co-ordinate transformation*.

Note: This definition is consistent with mathematical usage.

Matrix (verb). In color television, to perform a *color co-ordinate transformation* by computation or by electrical, optical, or other means.

Matrix Unit (Matrix Circuit). A device which performs a color co-ordinate transformation by electrical, optical, or other means.

Mixed Highs. Those high-frequency components of the picture signal which are intended to be reproduced achromatically in a color picture.

Moire. In television, the spurious pattern in the reproduced picture resulting from interference beats between two sets of periodic structures in the image.

Note: Moires may be produced, for example, by interference between regular patterns in the original subject and the target grid in an image orthicon, between patterns in the subject and the line pattern and the pattern of phosphor dots of a three-color kinescope, and between any of these patterns and the pattern produced by the *Chrominance Signal*.

Monochromatic. Referring to a negligibly small region of the spectrum.

Monochrome. Having only one *Chromaticity*, usually achromatic.

Monochrome Signal. (1) In monochrome television, a signal wave for controlling the *luminance* values in the picture. (2) In color television, that part of the signal wave which has major control of the *luminance* values of the picture, whether displayed in *color* or in *Monochrome*.

Monochrome Transmission. In television, the transmission of a signal wave for controlling the *luminance* values in the picture, but not the *chromaticity* values.

Narrow-Band Axis. In phasor representation of the *Chrominance Signal*, the direction of the phasor representing the *coarse Chrominance Primary*.

Nonphysical Primary. A *primary* represented by a point outside the area of the *chromaticity diagram* enclosed by the *Spectrum Locus* and the *Purple Boundary*.

Note: Nonphysical primaries cannot be produced because they require negative power at some wavelengths. However, they have properties which facilitate colorimetric calculation. *Tristimulus Values* based upon them are derived from *Tristimulus Values* based upon physical primaries.

Photometry. The techniques for the measurement of *Luminous Flux* and related quantities.

Note: Such related quantities are *Luminous Intensity*, *Illuminance*, *Luminance*, *Luminosity*, etc.

Pickup Spectral Characteristic. The set of spectral responses of the device, including the optical parts, which converts radiation to electric signals, as measured at the output terminals of the pickup tubes.

Note: Because of nonlinearity, the Spectral Characteristics of some kinds of pickup tubes depend upon the magnitude of *radiance* used in the measurement.

Planckian Locus. The locus of *chromaticities* of Planckian (blackbody) radiators having various temperatures (see Fig. 1).

Primaries. The *colors* of constant *chromaticity* and variable amount, which, when mixed in proper proportions, are used to produce or specify other *colors*.

Note: Primaries need not be physically realizable.

Purity (Excitation Purity). The ratio of the distance from the reference point to the point representing the sample, to the distance along the same straight line from the reference point to the *Spectrum Locus* or to the *Purple Boundary*, both distances being measured (in the same direction from the reference point) on the *CIE chromaticity diagram*.

Note: The reference point is the point in the *chromaticity diagram* which represents the reference standard light mentioned in the definition of *Dominant Wavelength*.

Purple Boundary. The straight line drawn between the ends of the *Spectrum Locus* (see Fig. 1).

Radiance. The *Radiant Flux* per unit solid angle per unit of projected area of the source.

Note: The usual unit is the watt per steradian per square meter. This is the radiant analog of *Luminance*.

Radiant Flux. The time rate of flow of radiant energy.

Radiant Intensity. The energy emitted per unit time, per unit solid angle about the direction considered; for example, watts per steradian.

Receiver Primaries. See *Display Primaries.*

Reference White. The light from a nonselective diffuse reflector which is lighted by the normal illumination of the scene.

Note 1: Normal illumination is not intended to include lighting for special effects.

Note 2: In the reproduction of recorded material, the word scene refers to the original scene.

Relative Luminosity. The ratio of the value of the *luminosity* at a particular wavelength to the value at the wavelength of maximum *luminosity.*

Saturation. The attribute of any color perception possessing a hue that determines the degree of its difference from the achromatic color perception most resembling it.

Note 1: This is a subjective term corresponding to the psychophysical term *Purity.*

Note 2: The description of saturation is not commonly undertaken beyond the use of rather vague terms, such as vivid, strong, and weak. The terms brilliant, pastel, pale, and deep, which are sometimes used as descriptive of saturation, have connotations descriptive also of *brightness.*

Spectral Characteristic. The set of spectral responses of the color separation channels with respect to wavelength.

Note 1: The channel terminals at which the characteristics apply must be specified and an appropriate modifier may be added to the term, such as pickup spectral characteristic or studio spectral characteristic.

Note 2: Because of nonlinearity, some Spectral Characteristics depend upon the magnitude of *radiance* used in the measurement.

Note 3: Nonlinearizing and matrixing operations may be performed within the channels.

Spectrum Locus. The locus of points representing the *Chromaticities* of spectrally pure stimuli in a *chromaticity diagram* (see Fig. 1).

Standard Observer. A hypothetical observer who requires standard amounts of *Primaries* in a *color mixture* to match every *color.*

Note: Standard amounts of a particular set of primaries used by this observer can be computed, by established methods, from standard amounts of the standard (and usually nonphysical) primaries. The present standard primaries and standard amounts of them required to match various wavelengths of the spectrum were established in 1931 by the International Commission on Illumination.

Transmission Primaries. The set of three *primaries,* either physical or nonphysical, so chosen that each corresponds in amount to one of the three independent signals contained in the *Color Picture Signal.*

Note: The *Chromaticities* of two possible sets of Transmission Primaries are: (1) Those of the *Display Primaries* (Receiver primaries), (2) Those of a specified *Luminance Primary* and two *Chrominance Primaries.*

Tristimulus Values. The amounts of the *Primaries* that must be combined to establish a match with the sample.

Value, Munsell. The dimension of the Munsell system of object-color specification which indicates the apparent luminous transmittance or reflectance of the object on a scale having approximately equal perceptual steps under the usual conditions of observation.

White.

Note: In color television, the term White is used most commonly in the nontechnical sense. More specific usage is covered by the term *Achromatic Locus,* and this usage is explained in the *Note* under the term *Achromatic Locus.*

White Object. An object which reflects all wavelengths of light with substantially equal high efficiencies and with considerable diffusion.

Wide-Band Axis. In phasor representation of the *Chrominance Signal,* the direction of the phasor representing the *fine Chrominance Primary.*

Zero-Subcarrier Chromaticity. The *chromaticity* which is intended to be displayed when the subcarrier amplitude is zero.

Correspondence

Diffusion Capacitances and High-Injection Level Operation of Junction Transistor*

It has been observed that the emitter- and collector-junction diffusion capacitances of a junction transistor do not vary with the emitter current as the theory for the low-injection level operation predicts.[1,2] In this note we shall calculate the diffusion capacitances, using the results already obtained and following Hall-Johnson's reasoning.[1,3]

The diffusion capacitance is due to the charge of holes[4] stored in the base region and is expressed by the ratio of the increment of the charge to that of the voltage across the junction which caused the former.

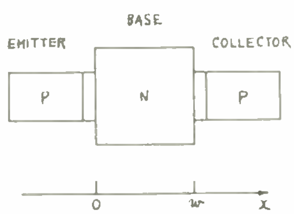

Fig. 1—*P-n-p* junction transistor and co-ordinate system.

When it is assumed that, in the base region, recombination of holes is small and the electron current is so in comparison with the hole current, which are valid in ordinary transistors, the hole distribution in the base region is given by Rittner's Eqs. (45) and (46)[5] in the co-ordinate system in Fig. 1:

$$x = (qD_p n_{BO}/I_p)[2(P_1 - P) + \ln(1 + P) - \ln(1 + P_1)], \quad (1)$$

$$I_p = (qD_p n_{BO}/w)[2P_1 - \ln(1 + P_1)], \quad (2)$$

where n_{BO} is the equilibrium electron density in the base region, P the hole density divided by n_{BO}, I_p the hole-current density, the subscript 1 implies the emitter end of the base region and the hole density at the collector end is equated to zero, corresponding to the negative bias applied to the collector. The number of holes per unit area of the emitter junction, N_p, in the base is obtained by integrating (1):

$$N_p = n_{BO}\int_0^{P_1} x(P)dP = (qD_p n_{BO}^2/I_p)[P_1^2 - P_1 + \ln(1+P_1)]; \quad (3)$$

therefore the charge of holes, Q, at given emitter and collector biases, V_E, V_C, is expressed by

$$Q = qN_p = (q^2D_p n_{BO}^2/I_p)[P_1^2 - P_1 + ln(1 + P_1)] \quad (4)$$

and (2), where P_1 is a function of V_E and w of V_C.

Using the following relation between P_1 and V_E[6]

$$P_1^2 + P_1 = (p_{BO}/n_{BO}) \exp (qV_E/kT), \quad (5)$$

where p_{BO} is the equilibrium hole density in the base region, which is an extension of Shockley's one for high-injection level, and noting I_p is a function of P_1 by (2), we obtain as an emitter-junction diffusion capacitance

$$C_{dE} = dQ/dV_E = qn_{BO}w(q/kT)(P_1/Z)(1 + P_1) \cdot [1 - (P_1/Z)], \quad (6)$$

where

$$Z = I_p w/qD_p n_{BO} = 2P_1 - \ln(1 + P_1) \quad (7)$$

by (2). Eqs. (6) and (7) are plotted in Fig. 2

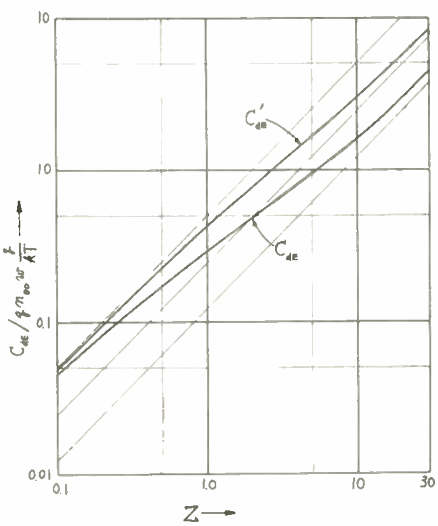

Fig. 2—Dependence of the emitter-junction diffusion capacitance on the emitter current. Z is a parameter nearly proportional to the emitter current: $Z = wI_{Ep}/(D_p n_{BO} q)$. The capacitance is per unit area of the emitter junction. Straight lines are asymptotes at high or low injection levels. C'_{dE} for Shockley's and C_{dE} for the extended boundary condition.

together with the capacitance C_{dE}' derived by using Shockley's relation ($P_1 = (p_{BO}n_{BO})$ exp (qV_E/kT)):

$$C_{dE}' = qn_{BO}w(q/kT)(P_1/Z)(1 + 2P_1) \cdot [1 - (P_1/Z)]. \quad (8)$$

Both C_{dE} and $C_{dE'}$ approach Johnson's result[3]

$$C_{dEO} = qn_{BO}w(q/kT)(Z/2) \quad (9)$$

when P_1 is small, but their high injection level values are $C_{dEO}/4$ and $C_{dEO}/2$, respectively. C_{dE}/C_{dEO} and C_{dE}'/C_{dEO} are plotted in Fig. 3 as a function of Z.

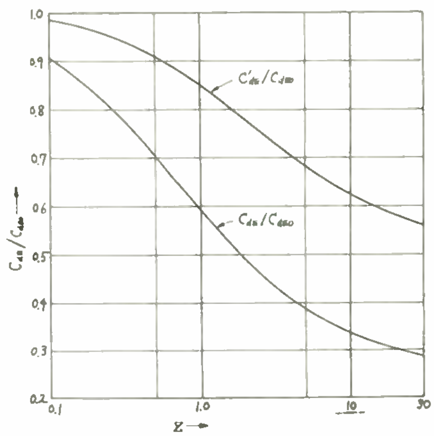

Fig. 3—"Fall-off" factor of the emitter-junction diffusion capacitances vs emitter current.

The collector-junction diffusion capacitance is obtained by the similar method, noting P_1 is dependent on w by (2):

$$C_{dC} = n_{BO}q(dw/dV_c)P_1. \quad (10)$$

This is approximated by Pritchard's one[1]

$$C_{dCO} = n_{BO}q(dw/dV_c)Z \quad (11)$$

for small P_1, and $C_{dCO}/2$ when $P_1 \rightarrow \infty$. C_{dC} and C_{dC}/C_{dCO} are plotted in Figs. 4 and 5.

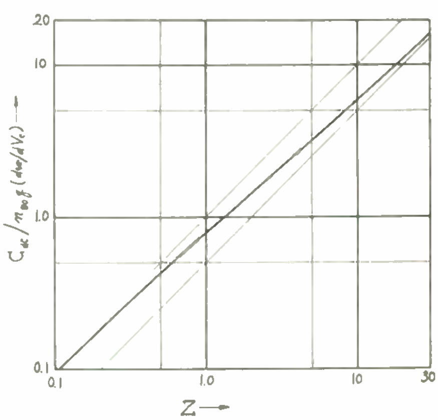

Fig. 4—Collector-junction diffusion capacitance vs emitter current.

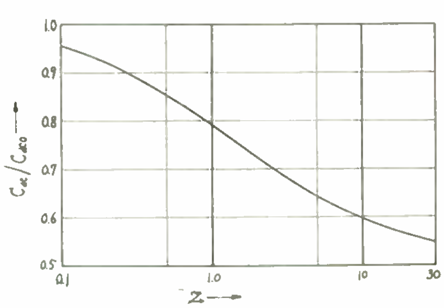

Fig. 5—"Fall-off" factor of the collector-junction diffusion capacitance vs emitter current.

* Received by the IRE, January 24, 1955.

[1] R. L. Pritchard, "Frequency variations of junction-transistor parameters," Proc. I.R.E., vol. 42, pp. 786–799; May, 1954.

[2] W. M. Webster, "On the variation of junction-transistor current-amplification factor with emitter current," Proc. I.R.E., vol. 42, pp. 914–920; June, 1954.

[3] C. W. Mueller and J. I. Pankove, "A *p-n-p* triode alloy-junction transistor for radio-freuqency amplification," Proc. I.R.E., vol. 42, pp. 386–391; February, 1954.

[4] Speaking of a *p-n-p* junction transistor.

[5] E. S. Rittner, "Extension of the theory of the junction transistor," *Phys. Rev.*, vol. 94, pp. 1161–1171; June 1, 1954.

[6] T. Misawa, *Jour. Phys. Soc* (Japan), to be published.

Even at high-injection level the ratio of C_{dc} to the depletion layer capacitance C_C serves as a measure of injection level,

$$C_{dc}/C_C = P_1. \qquad (12)$$

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Discussion on "Transient Response in FM," by I. Gumowski*

M. Nadler:[1] As Mr. Gumowski was kind enough to call attention to my work in his article, I feel called upon to explain the method I have used, which is somewhat at variance with his. My article[2] treated only phase jumps, corresponding here to the Dirac frequency modulation impulse considered by Gumowski. I must admit that I have not studied the use of divergent integrals in the sense of the distribution theory considered by Gumowski, but the results obtained seem to be equivalent to my results, attained by a more pedestrian means. In particular, Fig. 3 of Gumowski's paper agrees rather closely with Fig. 3 of mine.

The method I adopted was operational calculus based on the Laplace-Wagner transform, and involves a simple trigonometrical substitution, after which the problem simply falls apart.

Consider a signal $E \sin \omega t$ which is subjected to a sudden change of phase at the moment when the instantaneous phase has any arbitrary value. The resultant signal is then expressed by the following pair of equations:

$$E(t) = \begin{cases} E \sin(\omega t + \phi) & t \leq 0 \\ E \sin(\omega t + \theta) & t \geq 0 \end{cases} \qquad [\theta - \phi = \alpha].$$

This pair of equations, however, can be replaced by a single equation

$$E(t) = E\{\sin(\omega t + \phi) + H(t)[\sin(\omega t + \theta) - \sin(\omega t + \phi)],$$

where $H(t)$ is the unit step function. Upon performing a well known trigonometric substitution, we finally obtain

$$E(t) = E\left\{\sin(\omega t + \phi) + 2 \sin \frac{\alpha}{2} H(t) \cos\left(\omega t + \frac{\theta + \phi}{2}\right)\right\}.$$

Now, the first term exists for all t; therefore the response to this term is obtained from steady state theory, whereas the second term must be solved by some method of transient analysis. However, since it is a simple trigonometric function of time, any method will do, and will not be elaborated on here. The response to the second term, added to that of the first term for $t \geq 0$ obviously gives the response of the system to the phase jump at $t=0$. The result may be manipulated to yield the phase and amplitude modulation directly, while the equivalent frequency modulation is obtained by derivation of the phase function, as in equation 15 of Gumowski's paper.

The extension to the unit step-frequency modulation is obvious; the new frequency appears in one of the factors of $H(t)$, the old frequency in the other. Again a suitable trigonometric substitution yields us a manageable form.

On the basis of the above substitutions, the analysis of an FM or PM system is reduced to well known and rigorously established classical forms, as far as the unit impulse and the unit step of frequency is concerned. If we accept (or do we have to prove it?) that the FM network is a black box with a certain modulation at the input and another at the output, without regard to the mode (FM vs. AM), we need not go any further, for the response to these two forms can be used to infer the response to other functions, which are not so easily handled. The fact is that sinusoidal frequency modulation yields the infinite series known as a Bessel function no matter how you slice it. It is only Gumowski's good fortune that the impulse and step functions yield a closed form; I dare to predict that the divergent integrals in the sense of the distribution theory would not work on sinusoidal modulation. I know that the subterfuge I have used here will not help in the latter case.

I. Gumowski:[3] The method proposed by Mr. Nadler is well known. It merely supposes that the response, $e_0(t)$, to the signal

$$e_i(t) = e^{j\omega_0 t} \cdot e^{j\theta(t)} \qquad (a)$$

can be obtained from the differential equation of the network. Let $H(D)$ be a linear differential operator of degree, n, in D; then this equation is

$$H(D)e_0(t) = e^{j\omega_0 t} \cdot e^{j\theta(t)}. \qquad (b)$$

If $g(t)=0$ for $t<0$, and $g(t)\neq 0$ for $t>0$, then the general solution of (b) is, of course, given by

$$e_0(t) = \sum_{k=1}^{n} C_k \phi_k(t) + \frac{1}{H(j\omega_0)} e^{j\omega_0 t} + G(t), \quad (c)$$

where the first term represents the general solution of the homogeneous equation involving n arbitrary constants $[e_i(t)=0]$, the second term represents the particular integral contributed by the term $e_i(t)=e^{j\omega_0 t}$ (the steady state response), and $G(t)$ represents the particular integral contributed by the term, $e_i(t)=Y(t)e^{j\omega_0 t}[e^{j\theta(t)}-1]$, where $Y(t)$ is the unit-step function. To obtain the particular solution describing a completely defined circuit it is necessary to have some additional information in the form of initial or boundary conditions. In fact, n independent additional conditions are required to determine the n constants of integration, and thus assure a unique solution.

The usefulness of the Fourier transformation lies by no means in the fact that it renders possible the solution of problems which cannot be solved through the direct method illustrated by equations (a), (b) and (c), as Mr. Nadler seems to think. It is a well-known fact that the Fourier transformation fails in numerous cases where the direct method can be applied without any difficulty of principle. Quite to the contrary, the usefulness of Fourier's transformation lies mainly in the fact that it permits one to arrive directly at a particular solution, without having to evaluate any constants of integration. A set of particular boundary conditions is implied in the property that a transformable function, $f(x)$, must be defined for all values in $-\infty < x < +\infty$. The above reasoning is summarized in equation (10) of my paper, which expresses the response to a general signal, $F(t)$, in terms of the impulse response. Of course, equation (10) is only a particular form of equation (c), taking into account the particular set of boundary conditions mentioned above. Fourier's transformation is also useful, because it permits one to deal with problems in which the differential operator, $H(D)$, is either unknown or cannot be expressed in an analytic form; for example, when the transfer function, $A(\omega)$, is given graphically or as a set of measured values.

Suppose now that a network is given by specifying only its transfer function, $A(\omega) = H(j\omega)$. In order to calculate the network's response, $e_0(t)$, to a given signal, $e_i(t)$, by means of the operational calculus based on the Laplace-Wagner transformation, it is necessary to set $p=j\omega$. This substitution is not always valid In fact, according to classical theory, it may only be used if, and only if,

$$\left\{\begin{array}{l} \int_0^\infty e^{-pt} e_i(t) dt \\ \int_0^\infty e^{-pt} e_0'(t) dt \end{array}\right\} \text{ and } \left\{\begin{array}{l} \int_0^\infty e^{-j\omega t} e_i(t) dt \\ \int_0^\infty e^{-j\omega t} e_0(t) dt \end{array}\right\} \qquad (d)$$

exist in the Riemann, or at least in the Lebesgue sense, i.e. if, and only if, $e_i(t)$ and $e_0(t)$ are both Laplace *and* Fourier transformable. The above statement means simply that a function defined by an integral must be invariant with respect to a transformation of variables performed on that integral. In particular this means that no transformation may be used if it renders divergent an originally convergent integral, or vice versa. If the substitution, $p=j\omega$, is used in cases where (d) is not satisfied, then this amounts simply to leaving the framework of classical theory without mentioning this fact. If distributions, or the equivalent thereof, are not used, there is no guarantee whatsoever that the results will be correct.

If the modulating voltage, $g(t)$, is a sine wave (defined for $-\infty < t < +\infty$), then the Fourier transform of $e_i(t)=e^{j\omega_0 t} \cdot e^{j\theta(t)}$ reduces simply to the well-known Fourier series involving Bessel functions. The distribution theory neither helps nor hinders the process of obtaining this particular transform.

The impulse and step responses of my paper occur in closed form because the circuit of the numerical example was chosen with precisely this purpose in mind (note approximation used to pass from eq. (11a) to (11b). Good fortune, to use Nadler's words, had nothing to do with it.

* I. Gumowski, "Transient response in FM," PROC. IRE, vol. 42, pp. 819–822; May, 1954.

[1] Institute of Computation, Czechoslovak Academy of Science, Prague, Czechoslovakia.

[2] M. Nadler, "Transient frequency and amplitude modulation due to a phase jump" (in Czech), *Slaboproudý Obzor*, vol. 13, No. 1, p. 20; 1951.

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M. Nadler:[1] In general, my only comment to Mr. Gumowski's reply is to refer again to my previous letter. Aside from the trigonometric manipulation which enables the phase jump to be handled by the operational calculus based on the Laplace-Wagner (-Carson) transform, there is nothing new in my method, nor do I make such claims. I still have not studied distribution theory, although I have made some inquiries about it. My conclusions for the time are that we can still solve a large number of problems without it, and not get into difficulty with the mathematicians, although I reserve a final judgement for the time when I shall have made closer acquaintance with the subject.

As for the main assertion of Mr. Gumowski's letter, that the substitution $p=j\omega$ is invalid unless the time functions involved "are both Laplace *and* Fourier transformable," this *could* be true for active networks; but, for passive networks, no matter how defined, my own opinion is that the assertion is not true. Passive networks have analytic network functions in the right half-plane. Thus, whether we consider the Fourier transform, which must yield the inverse transform by integration on the axis $-j\infty - +j\infty$, or on an axis to the right, $c-j\infty$, $c+j\infty$, i.e. the Laplace transform, this cannot affect the result. Active networks, so long as they are stable, also must obey certain laws, and it can be shown that the same applies to them.

Before I close, however, I wish to mention that the last three sentences of my original letter, on rereading, are not clear. What I wished to point out was that in using my trigonometric substitution, expressions are avoided which are very difficult to integrate, without some form of development in a series or the like. I was at that time under the impression that the distribution theory was aimed at this question also. Since then I have found out that the theory is only concerned with a mathematical justification of the Dirac function, so that my remark is rather pointless.

As for Mr. Gumowski's "good fortune," the approximation used, to pass from (11a) to (11b), *also* has nothing to do with it. I performed the approximation *after* integration, and still had a closed form, namely, the full expression, which I then simplified by means of the given approximation, to obtain a simplified expression for high Q.

I. Gumowski:[3] Since Mr. Nadler contests my statement that the substitution $p=jw$ is invalid, unless the time functions involved are both Laplace *and* Fourier transformable, I feel that he misunderstands the nature of the problem in hand.

In fact, consider an $f(t)$, defined for all t in $-\infty < t < +\infty$, and its associated Fourier transform

$$g(j\omega) = \int_{-\infty}^{+\infty} e^{-j\omega t} f(t) dt. \qquad (1)$$

It is well known that the "frequency spectrum" $g(j\omega)$ defines in its turn uniquely

$$(t) = \frac{1}{2\pi} \int_{-\infty}^{+\infty} e^{j\omega t} g(i\omega) d\omega \qquad (2)$$

provided (1) and (2) exist in the Riemann or, at least, in the Lebesgue sense. Eq. (2) is useful in linear transient problems because it represents the function $f(t)$ as the superposition of sine waves, i.e. it permits to express the desired transient response $f(t)$ in terms of the easily obtainable "steady-state" sine-wave response. Unfortunately, (1) and (2) are difficult to use in practice, because, if the integrals are to converge, $f(t)$ must satisfy some very stringent conditions. In fact, most popular time-functions (the unit-step function, for instance) do not meet these conditions. The result is that the integral (1) defining their frequency spectrum $g(j\omega)$ diverges. By definition, this is interpreted to mean that such time-functions cannot be represented by the superposition of sine waves. It is obvious that the divergence of (1) depends only on the form of $f(t)$. Whether such an $f(t)$ represents a signal in an active or in a passive circuit, is rather irrelevant.

Consider now a function $f(t)$, defined for all t in $-\infty < t < +\infty$ but set identically equal to zero for all t in $-\infty < t < 0$, and its associated Laplace transform

$$\phi(p) = \int_0^{\infty} e^{-pt} f(t) dt. \qquad (3)$$

If the integral (3) converges, then one has the well-known inverse relation

$$\frac{1}{2\pi j} \int_{c-j\infty}^{c+j\infty} e^{zt} \phi(z) dz = \begin{Bmatrix} 0 & \text{for } t < 0 \\ f(t) & \text{for } t > 0 \end{Bmatrix}. \qquad (4)$$

Let us now determine under what conditions the substitution $p=j\omega$ implies the relation

$$g(j\omega) = \phi(p). \qquad (5)$$

Obviously, for (5) to hold, it is first necessary that $f(t)$ be identically zero for all t in $-\infty < t < 0$, and, second, it is necessary that *both* the integral (1) and the integral (3) be convergent. The second condition means, simply, that $f(t)$ must be both Laplace *and* Fourier transformable. It is difficult to take Mr. Nadler's remarks literally when he contests such an evident statement. If the $f(t)$ is such that (3) converges and (1) diverges (the unit-step function, for instance), then, by definition $g(j\omega)$ does not exist. It becomes, therefore, meaningless to affirm that the nonexisting $g(j\omega)$ is related to an existing $\phi(p)$ by the substitution $p=j\omega$, or for that matter, by any other substitution.

Let us now indicate briefly where the distribution theory enters the above argument. Since the integrals of the type (1) to (4) diverge for many potentially useful functions, it might be interesting to assign to such concepts as the integral a wider sense than that given to it by either Riemann or Lebesgue. There are, of course, many ways to carry out such an extension. A particular extension, which promises to be extremely useful in circuit problems, was elaborated in detail by the French mathematician Schwartz.[4] But, as it is the accepted custom, Schwartz's theory in no way invalidates any previously established laws. It merely extends some of these laws to two types of "generalized" functions, called, respectively, "measures" and "distributions." The best known of those "generalized" functions is the Dirac *measure* $\delta(t)$, which is called "unit-impulse" in engineering literature.

[4] L. Schwartz, "La Théorie des Distributions," Hermann et Cie, Paris, France.

M. Nadler:[1] It seems we have been talking about two different things. The substitution $p=i\omega$ which I have been considering is that one necessary to adapt the *network function* for use with operational calculus, for the purpose of obtaining the *time response* to the given input. Mr. Gumowski is interested in obtaining the *frequency spectrum* of the output signal, i.e., the Fourier transform of the output signal. Naturally, if the Laplace transform is used to obtain a Fourier transform it must itself be a Fourier transform except for the trivial substitution $p=i\omega$. In this case Mr. Gumowski's remarks are perfectly acceptable.

On the Performance of Reflex Klystrons*

Generally the operation of reflex klystrons is described by means of graphs giving frequency and output power as a function of the reflector voltage. More complete information might be obtained by means of graphs giving families of constant frequency and constant output power curves in a reflector voltage/beam voltage plane. For example, Figs. 1, 2 and 3 show such graphs respectively for the first, second and third mode of operation of the 723 A/B reflex klystron.[1] In these diagrams the curve $\Delta f=0$ refers to the frequency 9369 mc, and the frequency difference between two adjacent curves is 5 mc. The output power values are normalized with respect to the reference 30 mw, which is the power occurring at $V_a=300$v on the curve, $\Delta f=0$ for the first mode.

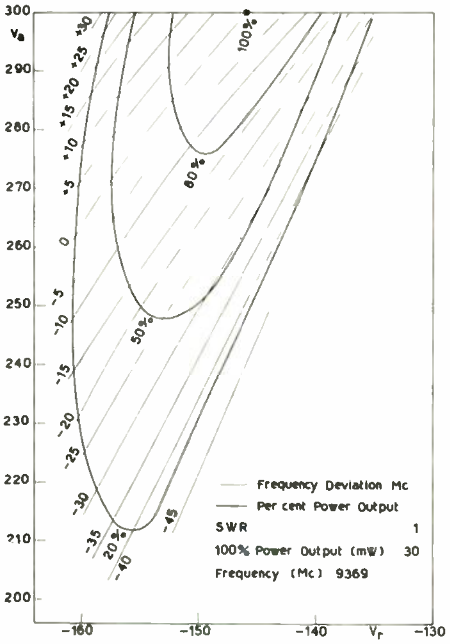

Fig. 1—Performance diagram, first mode.

These families of curves are very useful

* Received by the IRE, November 22, 1954.

[1] G. Zito, "723 A/B reflex klystron performance," *Zeit. Ang. Phys. und Math.*, vol. V, no. 3; 1954.

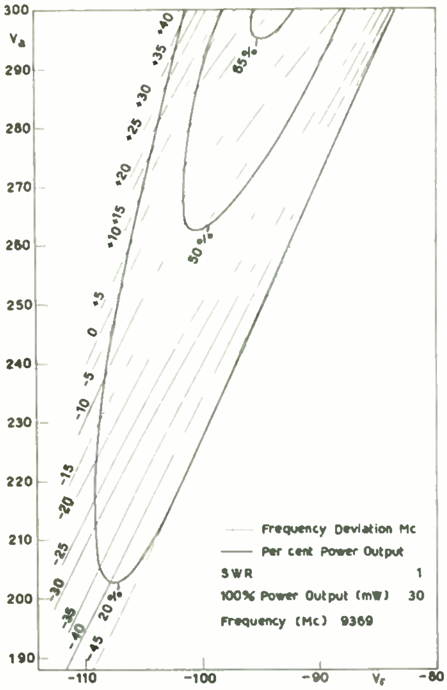

Fig. 2—Performance diagram, second mode.

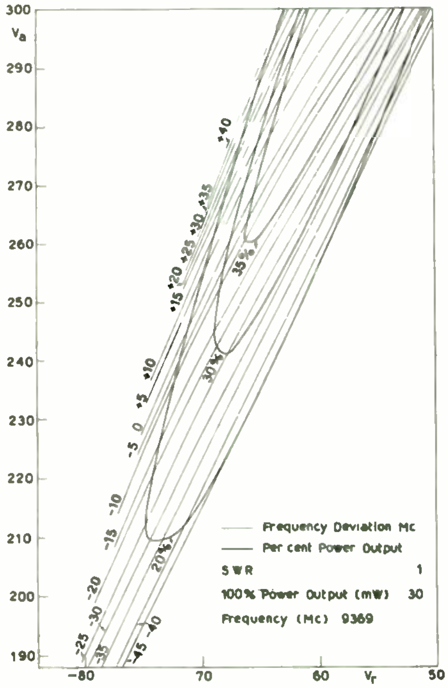

Fig. 3—Performance diagram, third mode.

for the determination of operations with pure amplitude or with pure frequency modulation. For this purpose reflector and beam voltage should be varied simultaneously so that the operating point remains on the prescribed curve of constant frequency or of constant output power. Another interesting application consists in the determination of optimum conditions of stable operation with respect to supply voltage variations. If the operating point is chosen in a region in which constant frequency and constant power curves are almost parallel, simultaneous variations of V_a and V_r, such that the ratio $\Delta V_a/\Delta V_r$ remains constant, do not affect the stability of operation.

The previous curves have been obtained under static operation; slight variations might be expected under dynamic operation due mostly to temperature effects.

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Analysis of Propagating Modes in Dielectric Sheets*

In the introduction to his paper[1] Mr. Hatkin states that, although it has been generally recognized that the propagating wave can be broken up into two crisscrossing wave components, this fact has not been utilized previously in solving for the propagating modes. Furthermore it is explained that the solution he derived is obtained by a combination of geometric optics and transmission line analogies.

In a paper[2] published in March, 1954 (manuscript received by the publisher, May 17, 1953), I have already described a very similar method using the crisscrossing wave components and a combination of geometric optics and transmission line analogies.

In this paper the method has been applied to the modes in waveguides, to a type of transmission line partially filled with dielectrics and to a type of dielectric clad conductor. All of these meet some of the suggestions contained in the conclusions of Mr. Hatkin's paper.

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* Received by the IRE, December 13, 1954.
[1] L. Hatkin, "Analysis of propagating modes in dielectric sheets," Proc. I.R.E., vol. 42, pp. 1565–1568; October, 1954.
[2] A. Guerbilsky, "Estudo sintético dos Guias de Onda,"—Publicação no. 7, Instituto de Eletrotécnica da U.B., Rio de Janeiro, Brazil; 1954.

A Review of VHF Ionospheric Propagation*

Following the appearance of my paper on this subject,[1] I received a communication from E. P. McGrath, Divisional Engineer of the Postmaster-General's Department, 42 Franklin Street, Adelaide, South Australia, citing the reception at Whyalla, South Australia, of 39.1 mc police transmissions from North America in October-December, 1947 and, with less intensity, during the same period in 1948. The signals were present from about 0800 hours to 1200 hours local mean time and a period of two or three days of particularly strong signals was thought to recur at 27-day intervals. Whyalla is at an elevation of 300 feet above sea level on the west side of Spencer's Gulf, and the signals from North America were received there on a 32-element vertically-polarized array with reflectors, aimed at Crystal Brook on the east side of the Gulf, from which FM telephone traffic (±15 kc deviation) is received. On occasions the interfering signals were sufficiently strong to take control of the circuit, notwithstanding the fact that Crystal Brook is but 45 miles away, is at an elevation of 600 feet, and was using 10 watts into a similar 32-element array.

The texts copied suggested that the transmissions were originating with the South Dakota police radio communications system, and Mr. McGrath wrote to the Commissioner of Police in Pierre in November, 1947 and again in August, 1948 to confirm this fact. Neither letter was acknowledged and his purpose in communicating with me was to see if I might succeed in obtaining verification. I wrote to the commanding officer of the state highway patrol in October, 1953 and again the letter was not acknowledged. In November, 1953 I called upon E. W. Allen, Jr., Chief Engineer of the FCC and a member of the IRE Wave Propagation Committee, to ask if his field organization would assist in obtaining verification of Mr. McGrath's log. In August, 1954 he was able to report that about half of the entries had been positively identified as originating with the South Dakota police radio communications system. The texts copied suggest that the remaining entries may have originated in nearby states or Canadian provinces. The FCC's investigation also revealed that police in South Dakota were reported as having heard Australian stations during the same period although this point was not verified.

The period of Mr. McGrath's log corresponds exactly to the period cited in my article for the highest muf's over the North Atlantic path during the last sunspot maximum, that is, the northern autumn of 1947. Further, the local mean time of reception in Australia corresponds to the interval 1130 hours to 1530 hours at the midpoint of the South Dakota-South Australia path (near Fanning Island) which is consistent with the regular diurnal behavior of F2 ionization.

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* Received by the IRE, February 2, 1955.
[1] Proc. IRE, vol. 41, pp. 582–587; May, 1953.

The Power Spectrum of a Carrier Frequency Modulated by Gaussian Noise*

Mr. J. L. Stewart's analysis[1] would appear to show that while phase modulation by a band of white noise produces, when the rms frequency deviation is small, a spectrum of finite extent, frequency modulation by the same noise also for the case of small deviation produces a continuous spectrum unbounded in frequency. This seems physically rather unlikely and comes about, I think, because of a rather too drastic limiting process in dealing with the frequency modulation case.

Frequency modulation by flat noise corresponds to phase modulation by noise whose amplitude is inversely proportional to frequency. The rf spectrum depends directly on the phase modulation: in fact, it can be developed from a power series in the phase modulation. Now, if the spectrum of the frequency modulating noise extends

* Received by the IRE, January 3, 1955.
[1] Proc. I.R.E., vol. 32, pp. 1539–1542; October, 1954.

down to very low frequencies, the corresponding phase modulation contains low-frequency terms of very high (in the limit infinite) amplitudes. Thus it would seem not unduly cautious if one first derives the rf spectrum when the minimum modulating frequency is not zero, and then proceeds to the limit, if it then turns out to be possible.

Mr. Stewart suggests that when the minimum modulating frequency is very small compared with the maximum modulating frequency, the continuous part of the small deviation rf spectrum is not very different from the form that he derives for zero minimum modulating frequency. In fact, it is hoped to show below that for finite minimum modulating frequency the spectrum is bounded, as in the phase modulation case.

Let B_m and B_0 be the maximum and minimum modulating frequencies (radians/sec), so that Stewart's expression (9) (retaining his symbols) becomes

$$W_F(\Delta\omega) = \frac{A_0^2}{2\pi}\int_0 \cos \Delta\omega\tau \cdot \exp\left\{\frac{-D^2\tau}{B_m-B_0}\int_{B_0\tau/2}^{B_m\tau/2}\left[\frac{\sin y}{y}\right]^2 dy\right\} d\tau$$

$$= \frac{A_0^2}{2\pi}\int_0^\infty \cos \Delta\omega\tau \cdot \exp\left\{\frac{-D^2\tau}{B_m-B_0}\left[Si\,(B_m\tau)-Si\,(B_0\tau) - \frac{\left(\sin\frac{B_m\tau}{2}\right)^2}{B_m\tau/2} + \frac{\left(\sin\frac{B_0\tau}{2}\right)^2}{B_0\tau/2}\right]\right\} d\tau, \quad (1)$$

where Si is the sine integral. For very large τ, the exponent becomes

$$\frac{-D^2}{B_mB_0} - \frac{D^2}{(B_m-B_0)\tau}\left[\frac{1}{B_0^2}\sin\,(B_0\tau) - \frac{1}{B_m^2}\sin\,(B_m\tau)\right]. \quad (2)$$

To make the integral converge we must separate off the constant term, which is done, following Stewart's procedure, by rewriting the power spectrum as

$$W_F(\Delta\omega) = \frac{A_0^2}{2\pi}\exp\left(\frac{-D^2}{B_mB_0}\right)\int_0^\infty \cos \Delta\omega\tau d\tau + \frac{A_0^2}{2\pi}\int_0^\infty \cos \Delta\omega\tau \cdot \left[\exp\left\{\frac{-D^2\tau}{B_m-B_0}\left[Si\,(B_m\tau)-Si\,(B_0\tau) - \frac{\left(\sin\frac{B_m\tau}{2}\right)^2}{B_m\tau/2} + \frac{\left(\sin\frac{B_0\tau}{2}\right)^2}{B_0\tau/2}\right]\right\} - \exp\left(\frac{-D^2}{B_mB_0}\right)\right] d\tau.$$

The first integral, expressed in terms of the Dirac delta function, is

$$\frac{A_0^2}{2}\exp\left(\frac{-D^2}{B_mB_0}\right)\delta(\Delta\omega),$$

which gives the energy remaining in the carrier. Stewart (p. 1541) writes down (without derivation) the delta function for this case, but his formula has an extra factor 2 in the numerator of the exponent.

If D is sufficiently small, the two exponentials in the second integral may be replaced by the first two terms of their power series developments, the spectrum becomes

$$W_F(\Delta\omega) = \frac{A_0^2}{2}\exp\left(\frac{-D^2}{B_mB_0}\right)\delta(\Delta\omega) + \frac{A_0^2}{2\pi}\int_0^\infty \cos \Delta\omega\tau \cdot \left\{\frac{-D^2\tau}{B_m-B_0}\left[Si\,(B_m\tau)-Si\,(B_0\tau) - \frac{\left(\sin\frac{B_m\tau}{2}\right)^2}{B_m\tau/2} + \frac{\left(\sin\frac{B_0\tau}{2}\right)^2}{B_0\tau/2}\right] + \frac{D^2}{B_mB_0}\right\} d\tau. \quad (3)$$

The second term, which leads to the continuous part of the spectrum, may be dealt with by integration by parts. After some rather tedious algebra the final result is

$$\frac{A_0^2}{2\pi}\left[\left\{\frac{-D^2}{B_m-B_0}\left[Si\,(B_m\tau)-Si\,(B_0\tau) - \frac{\left(\sin\frac{B_m\tau}{2}\right)^2}{B_m\tau/2} + \frac{\left(\sin\frac{B_0\tau}{2}\right)^2}{B_0\tau/2}\right] + \frac{D^2}{\tau B_mB_0}\right\} \cdot \left\{\frac{\tau}{\Delta\omega}\sin\,(\Delta\omega\tau) - \frac{1}{(\Delta\omega)^2}\cos\,(\Delta\omega\tau)\right\} - \frac{D^2}{B_m-B_0}\left\{-\frac{\cos\,(B_m\tau)\cos\,(\Delta\omega\tau)}{(\Delta\omega)^2B_m\tau} + \frac{\cos\,(B_0\tau)\cos\,(\Delta\omega\tau)}{(\Delta\omega)^2B_0\tau}\right\}\right]_0^\infty$$

$$+ \frac{A_0^2}{2\pi}\,\frac{D^2}{B_m-B_0}\int_0^\infty\left[\frac{\sin\,(B_m\tau)\cos\,(\Delta\omega\tau)}{(\Delta\omega)^2\tau} - \frac{\sin\,(B_0\tau)\cos\,(\Delta\omega\tau)}{(\Delta\omega)^2\tau}\right] d\tau.$$

The first expression vanishes at both upper and lower limits, leaving only the integral. This may be written as

$$\frac{A_0^2}{4\pi(\Delta\omega)^2}\cdot\frac{D^2}{(B_m-B_0)}\int_0^\infty\left[\frac{\sin\,(\overline{B_m-\Delta\omega}\tau)}{\tau} + \frac{\sin\,(\overline{B_m+\Delta\omega}\tau)}{\tau} - \frac{\sin\,(\overline{B_0-\Delta\omega}\tau)}{\tau} + \frac{\sin\,(\overline{B_0+\Delta\omega}\tau)}{\tau}\right] d\tau,$$

and evidently leads to a bounded spectrum. We have, finally,

$$W_F(\Delta\omega) = \frac{A_0^2}{2}\exp\left(\frac{-D^2}{B_mB_0}\right)\delta(\Delta\omega) + F(\Delta\omega) \quad (4)$$

where

$$F(\Delta\omega) = 0 \text{ for } \Delta\omega < B_0 \text{ or } > B_m$$

$$= \frac{A_0^2}{2}\,\frac{D^2/2(B_m-B_0)}{(\Delta\omega)^2} \text{ for } B_0 < \Delta\omega < B_m.$$

If B_0 is now made to approach zero, (4) will clearly not approach Stewart's expression (15). Stewart's spectrum, in fact, is derived by putting B_0 identically zero in (1) above, so the quantity in square bracket becomes

$$Si\,(B_m\tau) - \frac{\left(\sin\frac{B_m\tau}{2}\right)}{B_m\tau/2},$$

which, when $\tau\to\infty$, approaches $\pi/2$. But for any finite B_0, however small, the exact expression cannot approach $\pi/2$ with increasing τ because a cancelling $\pi/2$ will appear from the term $Si(B_0\tau)$. This is the root of the discrepancy between Stewart's result and that derived above.

A further point worth noting is that when $B_0\to 0$ there is difficulty involved in the condition that D is to be "small," since the validity of (4) depends on the function D^2/B_mB_0 being small.

It now becomes apparent that (4) above, as well as Stewart's (29) (the spectrum due to phase modulation by a flat noise band when the deviation is small) can be derived by strictly elementary methods, as can be seen by writing the frequency and phase modulation cases respectively in the forms

$$A_0 \cos\left\{\omega_0 t + \alpha\sum_{B=B_0}^{B_m}\frac{1}{B}\cos\,(Bt+\phi_B)\right\}$$

and

$$A_0 \cos\left\{\omega_0 t + \alpha\sum_{B=B_0}^{B}\cos\,(Bt+\phi_B)\right\},$$

where B is supposed to increase in steps of 1 radian/sec, ϕ_B is a random phase angle and α involves the deviation. Rewriting each expression as a sum of two products of cosines, and expanding the cosine and sine of the modulation as power series, the continuous parts of (4) and Stewart's (29) follow immediately. The power in the carrier is a little more awkward to extract, since it involves picking out the constant term from all even powers of the modulation, but the difficulty is not prohibitively great. This procedure yields higher orders in the spectrum quite easily and naturally. To obtain such higher orders by the autocorrelation approach [at least in the FM case, where it would be necessary to consider higher powers of the expression in square brackets in (3)] would seem excessively tedious.

The limiting case of the Gaussian spectrum for large deviations also follows immediately from a quasi-stationary argument,[2] in both PM and FM cases.

Thus it would seem that in the limiting cases considered by Mr. Stewart nothing new comes out of the autocorrelation approach. The derivation of the carrier power seems easier than before, but the continuous part of the spectrum emerges only after much more arduous labour. Perhaps the value of this method would be apparent for cases of intermediate deviation, which do not seem amenable to elementary analysis. Successful application of the method would presumably depend on whether one could approximate such terms as the exponential in (1) above by some simpler function giving a close fit over the whole range of τ, as Middleton did for a noise-band modulation having a Gaussian amplitude spectrum.

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The General Electric Co., Ltd.
Wembley, England

[2] A. S. Gladwin, "Energy distribution in the spectrum of a frequency modulated wave—Part II," *Phil. Mag.*, Ser. 7, XXXVIII, pp. 229–251; April, 1947.

Rebuttal[3]

Mr. Medhurst's comments regarding my paper are of interest and value, as he has solved for one of the limiting cases that I did not treat; that is, the shape of the power spectrum for $D^2 \ll B_1B$. The differences between Mr. Medhurst's results and my own stem from the nature of the limiting process and apply to *different* situations—my limiting expression assumes $D^2 \gg B_1B$. However, in extending his results to vanishing B_1, Mr. Medhurst is not correct unless the deviation D also vanishes—the result of such a limit where D must be zero when B_1 is zero is meaningless. Because of this, the "simpler" methods for determining the power spectrum for small B_1 as suggested by Mr. Medhurst are not applicable.

My eq. (9), evaluated for lower and upper modulator frequencies B_1 and B, respectively, is

$$W_F(\Delta\omega)=\frac{A_0^2}{2\pi}\int_0^\infty \cos\Delta\omega\tau \cdot \exp\left\{-\frac{D^2\tau}{B-B_1}\int_{B_1\tau}^{B\tau}\frac{1-\cos y}{y^2}\,dy\right\}d\tau. \quad (1)$$

The integral in the exponent can be expressed as the sum of two integrals. The integral of $(\cos y)/y^2$ vanishes for large τ and that of $1/y^2$ does not. This latter integral gives the delta function carrier simply as

$$W_F(\Delta\omega)=\frac{A_0^2}{2}\exp\left(-\frac{D^2}{B_1B}\right)\delta(\Delta\omega)+\frac{A_0^2}{2\pi}\int_0^\infty \cos\Delta\omega\tau \cdot\left\{\exp\left[-\frac{D^2\tau}{B-B_1}\int_{B_1\tau/2}^{B\tau/2}\frac{\sin^2 y}{y^2}\,dy\right]-\exp\left(-\frac{D^2}{B_1B}\right)\right\}d\tau. \quad (2)$$

If $D^2 \ll B_1B$, the exponential can be expanded with only its linear terms retained. In this case, Mr. Medhurst's results appear valid. If one should then go to the limit $B_1 \to 0$, it will also be required that $D \to 0$ in order to maintain D^2/B_1B small. However, a limiting expression based on such a calculation would not appear to be valid for any finite (albeit small) D: with $D \to 0$, it is meaningless. That Mr. Medhurst's limiting results are not reasonable can be demonstrated by finding the area under his eq. (4) (the power of the modulated signal) which increases without limit as $B_1 \to 0$.

By inspecting (2) above, it would appear that a unique expression for finite D with $B_1 \to 0$ can be obtained only by assuming the lower limit of the integral in the exponential $B_1\tau/2$ to be arbitrarily small for all τ. Then, $D^2/B_1B \to \infty$ and the delta function carrier component tends to zero. Then, the second exponential in the second term of (2) can be neglected and, setting $B_1\tau/2=0$, my previous results follow (except for the erroneous factor of two as pointed out by Mr. Medhurst). Mr. Medhurst is correct in stating that $W_F=0$ for $0<\Delta\omega<B_1$. However, that $W_F=0$ for $\Delta\omega>B$ depends upon whether D^2/B_1B is large or small—if it is large as is *always* the case when B_1 tends to zero with *finite* D, then W_F will not be zero at frequencies above B. This can also be seen from elementary arguments.

The power spectrum of a carrier frequency modulated at a sine wave frequency ω_m with a deviation D_m is

$$W(\Delta\omega)=\frac{A_0^2}{2}\sum_{n=0}^{\infty}J_n^2(D_m/\omega_m)\delta(\Delta\omega-n\omega_m). \quad (3)$$

Approximations valid for D_m/ω_m small and large are

$$W(\Delta\omega)\simeq\frac{A_0^2}{2}\sum_{n=0}^{\infty}\frac{(D_m/2\omega_m)^2}{(n!)^2}\delta(\Delta\omega-n\omega_m),\quad D_m\,\omega\ll_m \quad (4)$$

$$W(\Delta\omega)\simeq\frac{A_0^2}{2}\sum_{n=0}^{\infty}\frac{\cos^2(D_m/\omega_m-n\pi/2-\pi/4)}{D_m/\omega_m}\cdot\delta(\Delta\omega-n\omega_m),\quad D_m\gg\omega_m.$$

One can extend the reasoning using these equations to a continuous modulating power spectrum from B_1 to B by applying the equations to each frequency component in the frequency element $\delta\omega_m$ bracketing the frequency ω_m. If D_m/ω_m is small for *all* modulating frequencies, then in any frequency element, the first-order component due to the modulating frequency ω_m will be far more important than higher-order components arising from modulating frequencies less than ω_m that fall in the frequency element $\delta\omega_m$. In this case, D^2/B_1B is small and the power spectrum has the form $1/\Delta\omega^2$ for $B_1<\Delta\omega<B$ and is zero elsewhere. This is essentially the complete derivation of the spectrum shape by simple methods as commented upon by Mr. Medhurst.

On the other hand, if the deviation is not negligible compared to B_1, then there will exist at least some modulating frequency components that bring about large modulation indices—as $B_1 \to 0$ for any finite D_m, the various components in the modulating signal will result in indexes ranging from a small value to infinity; in which case, the higher order components of the lower modulating frequencies become very important, D^2/B_1B becomes very large, and simple methods cannot be used to derive the power spectrum.

It might be added that the differences between Mr. Medhurst's power spectrum and my own within the region $B_1<\Delta\omega<B$ are of second order—neither of us considered second-order terms. It would appear that the power spectrum is zero for $\Delta\omega<B_1$. The power spectrum may or may not be zero for $\Delta\omega>B$ depending upon the value of D^2/B_1B. In the limiting case of vanishing B_1 and finite D, my results are valid—that is, the power spectrum is smooth and continuous for all $\Delta\omega>B_1$.

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[3] Received by the IRE, January 20, 1955.

Cathode-Follower-Coupled Phase-Shift Oscillator*

As Professor Reich indicates,[1] the periodical literature appears to contain no mention of the idea of using cathode-followers between the sections of the phase-shifting network. It is shown, however, in a patent,[2] the outgrowth of some experimental work at the Naval Ordnance Laboratory in 1944. The basic circuit is shown in Fig. 1. The gain is concentrated in tube V_4 and the amplitude controlled by a thermistor TR.

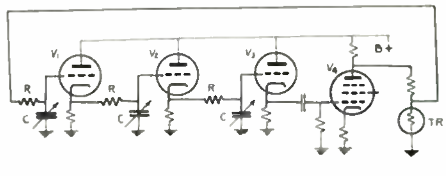

Fig. 1—Basic circuit of experimental phase-shift oscillator with cathode-follower coupling, for video frequency ranges.

The cathode follower expedient did not appear to improve the circuit enough to make it important for application to a general-purpose instrument. Using an ordinary 3-gang variable capacitor, the practical frequency range was from about 100 cycles to 4 or 5 megacycles. Using potentiometers the frequency range can be extended downward, limited by the need for a reasonably short time constant in the thermistor.

A very important limitation in the frequency range of practical phase-shift oscillators lies in the fixed stray phase-shifts, at low frequencies in the supply and bypass circuits, and at high frequencies in the stray capacitance across the resistors R. Hence it is found that the shunt-R configuration works best at low frequencies and the shunt-C configuration best at high frequencies. These effects have been discussed briefly and qualitatively by the writer[3] and by Butler,[4] but do not seem to have been looked into quantitatively. Such an analysis would be valuable.

Since each of the cathode followers requires a tube, the question arises whether these tubes might not as well be used also as the amplifiers. This takes us back to the original disclosure of Barrett;[5] the gain per tube needs only to be 2, and the maximum frequency attainable would appear to be higher than when all the gain is concentrated in one stage. There appears, however, to be no analysis of the respective merits. The "ring" circuit is used to cover the decade 1–10 mc in a commercial oscillator, and a simplified version covering 0.6–6 mc has been described.[6] One disadvantage of the "ring" is that it is capable of generating relaxation oscillations, while the regular phase-shift oscillator is not.[7]

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* Received by the IRE, February, 15, 1955.
[1] H. J. Reich, PROC. I.R.E., vol. 43, p. 229; February, 1955.
[2] L. Fleming, "Variable-Frequency Oscillator," U. S. Patent 2,565,490; August 28, 1951.
[3] L. Fleming, "Thermistor-regulated low-frequency oscillator," *Electronics*, p. 97, October, 1946.
[4] F. Butler, "Variable-frequency r-c oscillator," *Electronic Eng.*, pp. 140–142; April, 1949.
[5] R. M. Barrett, "*N*-phase resistance-capacitance oscillators," PROC. I.R.E., vol. 33, pp. 541–545; August, 1945.
[6] L. Fleming, "A video test oscillator," *Electronics*, vol. 27, p. 196, May, 1954.
[7] W. A. Edson, "Vacuum Tube Oscillators," John Wiley and Sons, New York, p. 190, 1953.

Contributors

For a photograph and biography of M. Balser, see page 62A of the September, 1954 issue of PROCEEDINGS OF THE IRE.

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D. Barbiere was born on August 10, 1924, in Brooklyn, N. Y. He received the A.B. degree in physics in 1945, and the M.S. degree in physics in 1948, both from New York University. During the summers of 1947 and 1948 he was engaged in fundamental research in the field of gas discharges at the Westinghouse Research Laboratories, Pittsburgh, Pa.

D. BARBIERE

From 1948 to 1950, Mr. Barbiere was employed by the Glenn L. Martin Co., Baltimore, Md., where he worked on the design and development of servo systems and antennas. In 1950, he joined the staff of the Radiation Laboratory of The Johns Hopkins University, Baltimore, Md.

Since 1953, Mr. Barbiere has been associated with Bendix Radio, Division of Bendix Aviation Corp., Baltimore, Md., where he is conducting studies on component and system reliability in the missile program.

Mr. Barbiere is a member of Sigma Xi and Phi Beta Kappa.

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D. B. Brick (S'50–A'54) was born on October 1, 1927, in Brooklyn, N. Y. He attended Newark College of Engineering from 1945 to 1946. After serving in the U. S. Army, he entered Harvard University where he received the A.B. degree in engineering sciences and applied physics in 1950, and the S.M. and Ph.D. degrees in applied physics in 1951 and 1954 respectively.

D. B. BRICK

Dr. Brick has been a teaching fellow and research assistant at Harvard and at present he holds a post-doctoral research fellowship there. He is doing research in the fields of electromagnetic scattering, microwave circuits and radiation.

Dr. Brick is a member of Sigma Xi and the Harvard Engineering Society.

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M. H. Brockman (S'43–A'45–M'50–SM'53) was born in Richmond Beach, Wash., on September 8, 1918. He received the B.S. degree in electrical engineering from the University of Washington in 1943. After graduate work at Harvard University, Northwestern University and Polytechnic Institute of Brooklyn, he received the Master of Electrical Engineering degree from the Polytechnic Institute of Brooklyn.

M. H. BROCKMAN

In 1943, Mr. Brockman joined the staff of Radio Research Laboratory of Division 15, OSRD at Harvard University where he participated in receiver development until he became head of the test group in the pre-production section in 1944. In June 1946 Mr. Brockman joined the engineering staff of Airborne Instruments Laboratory in Mineola, N. Y. Mr. Brockman is a member of the applied electronics section of the laboratory.

Mr. Brockman is a member of Tau Beta Pi and Zeta Mu Tau and an associate member of A.I.E.E. and Pi Mu Epsilon.

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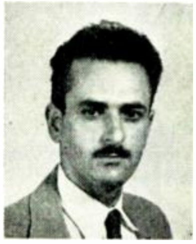
L. A. Freedman (S'49–A'51) was born on March 29, 1927, in Philadelphia, Pa. After serving in the Navy from 1945 to 1946, he received the B.S. degree in electrical engineering in 1948 from Drexel Institute of Technology, and the M.S. degree in electrical engineering from Rutgers University in 1950. While attending Rutgers he was an assistant in the electrical engineering department.



L. A. FREEDMAN

In 1950 he joined the research staff of the David Sarnoff Research Center of RCA Laboratories in Princeton, N. J.

He is a member of Phi Kappa Phi, Tau Beta Pi, Sigma Xi, and Eta Kappa Nu.

❖

D. D. Holmes (A'48) was born in Portland, Maine, on August 12, 1926. He received the B.S.E.E. degree from the University of Maine in 1946. From 1946 to 1947 he was with Hazeltine Electronics Corp., working on television receiver circuitry. From 1947 to 1949 he was instructor in electrical engineering at the University of Nebraska. He received the S.M. degree in electrical engineering from Massachusetts Institute of Technology in 1950.

D. D. HOLMES

Since 1950, Mr. Holmes has been a staff member at the David Sarnoff Research Center, RCA Laboratories, in Princeton, N. J., working in the fields of color television and transistor applications.

❖

E. G. Hopkins was born in Toowoomba, Australia, on December 27, 1918. He joined the staff of Amalgamated Wireless Valve Company as a cadet engineer in 1937 and remained with that company for twelve years, working on receiving tubes, power tubes, gas-filled tubes and industrial tube applications. Dr. Hopkins received the Diploma of Electrical Engineering of the Sydney Technical College in 1944 and the B.E. and Ph.D. degrees from the New South Wales University of Technology in 1952 and 1955 respectively.

E. G. HOPKINS

In 1949 he joined the New South Wales University of Technology to carry out research on self-heating thermionic tubes and, at present, is a lecturer in the School of Electrical Engineering.

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H. R. Johnson (S'45–A'51–SM'55) was born April 26, 1926, in Jersey City, N. J. He attended Cornell University from 1943 to 1947 and the Massachusetts Institute of Technology from 1947 to 1951. He received the B.E.E. from Cornell in 1946 and the Ph.D. in physics from M.I.T. in 1952. At Cornell he was associated with an OSRD project in 1945 and was a teaching assistant in physics from 1946 to 1947. At M.I.T. Dr. Johnson was a research fellow in electronics and a research assistant. Since 1951 he has been with the Electron Tube Laboratory of the Hughes Research and Development Laboratories, where he is currently head of the microwave tube section.

H. R. JOHNSON

He is a member of Tau Beta Pi, Eta Kappa Nu, Phi Kappa Phi, Sigma Xi, Gamma Alpha, and American Physical Society.

❖

H. P. Kalmus (A'39–SM'45) was born in 1906, in Vienna, Austria. He was graduated from the Technical University at Vienna.

From 1930 to 1938 he was with the Orion Radio Corp. in Budapest, Hungary. Mr. Kalmus was development engineer for the Emerson Radio Corp. in New York from 1939 to 1941. From 1941 to 1948 he did research work for the Zenith Radio Corp., where he developed the "Cobra" transducer and a novel light meter, based on magnetic modulation. Since 1948 he has been with the National Bureau of Standards and the Diamond Ordnance Fuze Laboratories in Washington, D. C. In 1954 he received the Exceptional Service Award from the U. S. Department of Commerce.

H. P. Kalmus

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B. Reich was born in New York, N. Y., on January 7, 1926. He received the B.S. degree in physics from the College of the City of New York in 1948. From 1948 to 1954, he did graduate work in electrical engineering and mathematics at Rutgers University, attending on a part-time basis.

B. Reich

Mr. Reich has been employed at the Signal Corps Engineering Laboratories since 1948, serving as a radiation physicist. During the period from May, 1951 to August, 1953, he served as full time Radiological Safety Officer for the Signal Corps Engineering Laboratories and Fort Monmouth. At present, Mr. Reich is working in the Solid State Devices Branch, Evans Signal Laboratory, Belmar, N. J.

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S. D. Robertson (S'37–A'41–SM'45) was born at Aurora, Mo., on August 8, 1915. He received the B.E.E. degree in 1936 from the University of Dayton. He then attended the Ohio State University and was awarded the M.Sc. degree in 1938 and the Ph.D. degree in 1941. He was a Stillman W. Robinson Fellow at Ohio State University from 1937 to 1939.

S. D. Robertson

In 1940, Dr. Robertson taught electrical engineering at the University of Dayton. Since 1940 he has been a member of the technical staff of Bell Telephone Laboratories, where he has been engaged in microwave radio, radar and vacuum tube research.

He is a member of Sigma Xi, Eta Kappa Nu, and Tau Beta Pi.

❖

S. Schneider was born in New York, N. Y., on February 24, 1924. He received the B.A. degree from Brooklyn College in 1946 and the M.S. degree from New York University three years later. From 1943 to 1946 he served with the U. S. Army.

S. Schneider

Mr. Schneider joined the Signal Corps Engineering Laboratories in 1948 and at present he is chief of the Power and Gas Tubes Research Unit.

He is a member of the American Physical Society and the American Institute of Physics.

❖

K. K. Shrivastava was born in Jabalpur, India, on May 25, 1926. He obtained the B.Sc. degree at the University of Sagar in 1947 and the B.E. degree with honors in telecommunication in in 1951 from the same university.

K. K. Shrivastava

In 1951 he joined All India Radio and in 1953 was awarded a Junior Fellowship for two years of post-graduate study at the New South Wales University of Technology in Australia, under the Technical Co-operation scheme of the Colombo Plan. Mr. Shrivastava returned to India in February, 1955.

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For a photograph and biography of R. A. Silverman, see page 82A of the September 1954 issue of Proceedings of the IRE.

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T. O. Stanley was born in Orange, N. J., on May 15, 1927. He served in the Navy from 1945 to 1946 and received his B.E. degree in electrical engineering from Yale University in 1949. He was awarded the Westinghouse Fellowship at Yale, receiving his M.E. degree in 1950.

T. O. Stanley

Since 1950 he has been with RCA, at the David Sarnoff Research Center, Princeton, N. J., where he has worked on color television circuits and transistor circuit applications.

Mr. Stanley is a member of Sigma Xi and Tau Beta Pi.

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M. W. P. Strandberg (A'42–SM'51) was born in Box Elder, Mont., on March 9, 1919. He received his B.S. degree from Harvard College in 1941 and his Ph.D. degree in physics from M.I.T. in 1948.

From 1941 through 1945 he was with the M. I. T. Radiation Laboratory, engaged in advanced developmental work on microwave radar. He spent a year in 1942–43 at T.R.E. in Malvern, Eng. as a visiting scientist working on radar counter measures for the Royal Air Force.

M. Strandberg

He returned to M.I.T. in 1947 where he is now associate professor of physics. His general field of research is microwave physics.

Dr. Strandberg is a member of the American Physical Society and Phi Beta Kappa.

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P. G. Sulzer (A'46) was born in Media, Pa., on August 3, 1922. He attended Drexel Institute of Technology in Philadelphia from 1940 to 1943. During that time he also spent one year with the Westinghouse Radio Division in Baltimore, Md.

P. G. Sulzer

He received the B.S. degree in 1947, and the M.S. degree in 1949, both in electrical engineering, from the Pennsylvania State College, where he had been engaged in designing ionosphere equipment. Since September, 1949, Mr. Sulzer has been employed at the Central Radio Propagation Laboratory of the National Bureau of Standards in Washington, where he is concerned with ionospheric instrumentation.

He is a member of Sigma Xi and Pi Mu Epsilon.

❖

M. Tinkham was born in Green Lake County, Wis., on February 23, 1928. He received his A.B. from Ripon College in 1951 under a joint study program, after two years at Ripon and two years at M.I.T. He received his S.M. degree in physics from M.I.T. in 1951, and his Ph.D. in physics from M.I.T. in 1954.

M. Tinkham

During the 1953–1954 academic year he held a National Science Foundation Predoctoral Fellowship, and the same agency has awarded him a Postdoctoral Fellowship, which he is using for study and research at the Clarendon Laboratory, Oxford, England, during the 1954–1955 academic year.

Dr. Tinkham is a member of the American Physical Society and Sigma Xi.

IRE News and Radio Notes

PG on Electron Devices to Hold Washington Technical Meeting

The Professional Group on Electron Devices will hold its first Annual Technical Meeting October 24 and 25 at the Shoreham Hotel in Washington, D. C. It will be the first of the annual meetings on electron devices to provide an exclusive medium for exchange of engineering information. George D. O'Neill, of Sylvania Electric Products Inc., will be chairman of the symposium. Mr. O'Neill, in describing plans for the meeting, said that specialists will gather from government and industry for an exchange of information on advanced developments and applications of both electron tubes and transistors in radio, television, business machines, and military equipment.

Approximately 1,000 engineers and scientists are expected to attend, with papers by scientists representing organizations in the United States and foreign countries.

Symposium on Electronics and Automatic Production Will Meet Preceding WESCON

Electronics and Automatic Production will be the theme of a national symposium to be held August 22 and 23 in San Francisco. Jointly sponsored by Stanford Research Institute and the National Industrial Conference Board, the meeting will immediately precede the Western Electronic Show and Convention scheduled for San Francisco the same week.

William D. McGuigan will be general chairman for the symposium at which more than two dozen papers will be presented to the group.

One section of the program will deal with the impact of automation on economic welfare, the use of leisure time, and personnel training in the future. Management considerations of capital and plant requirements, production control, and marketing will be covered by another set of papers. How automation will affect the electronics, machine tools, and materials-handling industries also will receive attention on the symposium program.

A session will be devoted to automatic production systems, assessing them in terms of performance, quality control, the effect on personnel requirements, and management and organizational reorientation. Also covered will be factors influencing product design, materials used and the processing of electronic and mechanical products. Papers have been scheduled on electronic applications to automatic control of machines, production lines, and quality. There will be reports on electronic automation development of new materials, components, product designs, and fabrication techniques, and new developments in automatic machinery.

Additional information concerning the symposium is available from Stanford Research Institute at Palo Alto, Calif., and the National Industrial Conference Board, 247 Park Avenue, New York City.

Vehicular Communications Paper Deadline Is Set for August First

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.

Institution of Telecommunication Engineers in India Delivers Report on Progress

The Annual Report of the Institution of Telecommunication Engineers reveals that the organization, inaugurated in New Delhi, India in November, 1953, has, in one year, grown remarkably in membership and activities.

The Institution, a unique professional body comprising all the specialized branches of telecommunication engineering, already has more than a thousand members, and enrolment is increasing rapidly. Its membership is drawn from various government operated communication agencies, the three defence services, research institutes, and industry. Run by a governing council of 24, the Institution, like most professional bodies, prescribes minimum educational qualifications and experience for entry into its several categories, but direct admission into the lower categories is also possible by passing the examinations which will be conducted. A quarterly, called the *Journal of the Institution Telecommunication Engineers*, has been proposed and the first issue is already in press. Talks and discussion meetings are now arranged periodically at New Delhi, headquarters of the Institution, and similar activities are being planned at Bombay, Calcutta, Madras, Poona, Bangalore and Jabalpur. Further particulars about the Institution may be obtained from The Honorary Secretary, The Institution of Telecommunication Engineers, Post Box No. 481, New Delhi, India.

Symposium on Normal Mode Theory to Be Held in San Diego

The Office of Naval Research and the Navy Electronics Laboratory will co-sponsor a symposium on *Normal Mode Theory* to be convened at the Navy Electronics Laboratory in San Diego, July 5–7, 1955. The symposium will be headed by S. A. Schelkunoff of Bell Telephone Laboratories and will consist of a round-table discussion on the present state of theoretical knowledge of tropospheric wave propagation, the known methods of attack, and the outstanding unanswered questions. Further information may be obtained from J. B. Smyth at the Navy Electronics Laboratory.

Calendar of Coming Events

IRE-AIEE Symposium on Electronic Materials and Components, University of Pennsylvania, Philadelphia, Pa., June 2–3

Conference of National Society of Professional Engineers, Bellevue-Stratford Hotel, Philadelphia, Pa., June 2–4

IRE-Rockefeller Institute meeting on Instrumentation in Electroencephalography, Medical Electronics Center, Rockefeller Institute, New York City, June 14

American Society for Engineering Education Annual Meeting, Pennsylvania State University, State College, Pennsylvania, June 20–24

URSI-U. of Michigan International Symposium on Electromagnetic Wave Theory, University of Michigan, Ann Arbor, Mich., June 20–25

Fifth International Aeronautical Conference, IAS Bldg., Los Angeles, California, June 21–24

American Nuclear Society First Annual Meeting, Pennsylvania State University, State College, Pa., June 27–29

Ohio State U. and Wright Air Development Center Radome Symposium, Columbus, Ohio, June 27–29

IRE-West Coast Electronic Manufacturers' Association WESCON, Civic Auditorium, San Francisco, California, August 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

IRE Professional Group on Nuclear Science—Second Annual Meeting, Oak Ridge National Laboratories, Oak Ridge, Tenn., Sept. 14–17

PG on Vehicular Communications Sixth Annual Meeting, Multnomah Hotel, Portland, Ore., Sept. 26–27

IRE-AIEE Conference on Industrial Electronics, Rackham Memorial Building, Detroit, Michigan, September 28–29

National Electronics Conference, Hotel Sherman, Chicago, Ill., October 3–5

IRE East Coast Conference on Aeronautical and Navigational Electronics, Lord Baltimore Hotel, Baltimore, Md., Oct. 31–Nov. 1

IRE-AIEE-ACM Eastern Joint Computer Conference, Hotel Statler, Boston, Nov. 7–9

M.I.T. Offers Special Two Week Summer Program in Numerical Control of Machine Tools

To assist in meeting the demand for technical information on automatic control, a two-week Special Summer Program in *Numerical Control of Machine Tools* will be presented by the Massachusetts Institute of Technology from Aug. 22 through Sept. 2.

Professor J. Francis Reintjes, Director of the Servomechanisms Laboratory of the M.I.T. Department of Electrical Engineering, will direct the program, in co-operation with James O. McDonough and other members of the Numerical Control Group.

Lecture topics will include: principles of information—processing as applied to the use of machine tools; numerical-control systems and their machine-tool applications; equipment design for numerical-control systems; design considerations for system reliability; and management, operation, and maintenance of numerically controlled machine tools.

Afternoon sessions will be devoted to programming techniques and will include such topics as mathematics of programming, practical procedures, and machine aids. The group will prepare a program for machining a work piece and actually execute operation.

The number of registrants is limited, and preference will be given to applicants now engaged in the design or application of automatic machine-tool equipment or anticipating entrance into this field.

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, Mass.

Symposium on Electromagnetic Wave Theory Held this Month

The Symposium on Electromagnetic Wave Theory will be held June 20–25 at the University of Michigan in Ann Arbor. K. M. Siegel, Willow Run Research Center of the Engineering Research Institute, will be Chairman of the symposium. Sponsoring groups include Commission Six of URSI and the University of Michigan.

Among the topics considered by the symposium will be: *Boundry Value Problems of Diffraction and Scattering, Forward and Multiple Scattering, Antenna Theory,* and *Propagation in Doubly Refracting Media in Wave Guides.*

Plans have been made for a number of special events and the university itself offers many recreational opportunities. A reception will be held on June 20, first day of the symposium. On June 22 there will be a trip to the Ford Rouge Plant. A party and a concert are also planned, and on June 23 there will be a banquet.

The papers of the Proceedings of the Conference will be published by the IRE Professional Group on Antennas and Propagation. Copies will be distributed without cost to members of the Group and to nonmembers at a nominal cost.

Further information may be obtained from J. T. Bolljahn, Stanford Research Institute, Menlo Park, California.

Registration fee is five dollars.

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

The following issues of TRANSACTIONS are available from the Institute of Radio Engineers, Inc., 1 East 79 Street, New York 21, N. Y., at the prices listed below.

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
Aeronautical & Navigational Electronics	PGA-5: A dynamic Aircraft Simulator for Study of Human Response Characteristics (6 pages)	$ .30	$ .45	$ .90
	PGAE-6: Ground-to-Air Cochnnel Interference at 2900 NC (10 pages)	.30	.45	.90
	PGAE-8: June 1953 (23 pages)	.65	.95	1.95
	PGAE-9: September 1953 (27 pages)	.70	1.05	2.10
	Vol. ANE-1, No. 1, March 1954 (51 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 2, June 1954 (22 pages)	.95	1.40	2.85
	Vol. ANE-1, No. 3, September 1954 (27 pages)	1.00	1.50	3.00
	Vol. ANE-1, No. 4, December 1954 (27 pages)	1.00	1.50	3.00
Antennas and Propagation	PGAP-4: IRE Western Convention, August 1952 (136 pages)	2.20	3.30	6.60
	Vol. AP-1, No. 1, July 1953 (30 pages)	1.20	1.80	3.60
	Vol. AP-1, No. 2, October 1953 (31 pages)	1.20	1.80	3.60
	Vol. AP-2, No. 1, January 1954 (39 pages)	1.35	2.00	4.05
	Vol. AP-2, No. 2, April 1954 (41 pages)	2.00	3.00	6.00
	Vol. AP-2, No. 3, July 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 4, October 1954 (36 pages)	1.50	2.25	4.50
	Vol. AP-3, No. 1, January 1955 (43 pages)	1.60	2.40	4.80
Audio	PGA-5: Design Interrelations and Records and Reproducers (8 pages)	.30	.45	.90
	PGA-7: Editorials, Technical Papers and News, May 1952 (47 pages)	.90	1.35	2.70
	PGA-10: November-December 1952 (27 pages)	.70	1.05	2.10
	Vol. AU-1, No. 1, January–February 1953 (24 pages)	.60	.90	1.80
	Vol. AU-1, No. 2, March–April 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 3, May–June 1953 (34 pages)	.80	1.20	2.40
	Vol. AU-1, No. 4, July–August 1953 (19 pages)	.70	1.05	2.10
	Vol. AU-1, No. 5, September–October 1953 (11 pages)	.50	.75	1.50
	Vol. AU-1, No. 6, November–December 1953 (27 pages)	.90	1.35	2.70
	Vol. AU-2, No. 1, January–February 1954 (38 pages)	1.20	1.80	3.60
	Vol. AU-2, No. 2, March–April 1954 (31 pages)	.95	1.40	2.85
	Vol. AU-2, No. 3, May–June 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 4, July–August 1954 (27 pages)	.95	1.40	2.85
	Vol. AU-2, No. 5, September–October 1954 (22 pages)	.95	1.40	2.85
	Vol. AU-2, No. 6, November–December 1954 (24 pages)	.80	1.20	2.40
	Vol. AU-3, No. 1, January–February 1955 (20 pages)	.60	.90	1.80
Broadcast Transmission Systems	PGBTS-: March 1955 (102 pages)	2.50	3.75	7.50
Broadcast and Television Receivers	PGBTR-1: Round-Table Discussion on UHF TV Receiver Considerations, 1952 IRE National Convention (12 pages)	.50	.75	1.50
	PGBTR-3: June 1953 (67 pages)	1.40	2.10	4.20
	PGBTR-5: January 1954 (96 pages)	1.80	2.70	5.40
	PGBTR-6: April 1954 (119 pages)	2.35	3.50	7.00
	PGBTR-7: July 1954 (58 pages)	1.15	1.70	3.45
	PGBTR-8: October 1954 (20 pages)	.90	1.35	2.70
Circuit Theory	PGCT-1: IRE Western Convention, August 1952 (100 pages)	1.60	2.40	4.80
	PGCT-2: Papers presented at the Circuit Theory Sessions of the Western Electronic Show & Convention, San Francisco, Calif., August 19–21, 1953 (106 pages)	1.95	2.90	5.85
	Vol. CT-1, No. 1, March 1954 (80 pages)	1.30	1.95	3.90
	Vol. CT-1, No. 2, June 1954 (39 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 3, September 1954 (73 pages)	1.00	1.50	3.00
	Vol. CT-1, No. 4, December 1954 (42 pages)	1.00	1.50	3.00
Communications Systems	Vol. CS-2, No. 1, January 1954 (83 pages)	1.65	2.50	4.95
	Vol. CS-2, No. 3, November 1954: IRE Symposium on Global Communications, June 23–25, 1954	3.00	4.50	9.00

*** Public libraries, colleges, and subscription agencies may purchase at IRE member rate.**

(*Continued on the following page*)

TRANSACTIONS OF IRE PROFESSIONAL GROUPS

(Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
	Washington D. C., & IRE-AIEE Symposium on Military Communications, April 28, 1954, New York, N. Y. (181 pages)			
	Vol. CS-3, No. 1, March 1955: Papers presented at the Symposium on Marine Communications & Navigation, October 13–15, 1954, Boston, Mass. (72 pages)	$1.00	$1.50	$3.00
Component Parts	PGCP-1: March 1954 (46 pages)	1.20	1.80	3.60
	PGCP-2: September 1954 Issue: Papers presented at the Component Parts Sessions at the 1954 Western Electronic Show & Convention, Los Angeles, Calif. (118 pages)	2.25	3.35	6.75
Electronic Computers	Vol. EC-2, No. 2, June 1953 (27 pages)	.90	1.35	2.70
	Vol. EC-2, No. 3, September 1953 (27 pages)	.75	1.10	2.25
	Vol. EC-2, No. 4, December 1953 (47 pages)	1.25	1.85	3.75
	Vol. EC-3, No. 1, March 1954 (39 pages)	1.10	1.65	3.30
	Vol. EC-3, No. 2, June 1954 (65 pages)	1.65	2.45	4.95
	Vol. EC-3, No. 3, September 1954 (54 pages)	1.80	2.70	5.40
	Vol. EC-3, No. 4, December 1954 (46 pages)	1.10	1.65	3.30
	Vol. EC-4, No. 1, March 1955 (48 pages)	1.10	1.65	3.30
Electron Devices	PGED-2: Papers on Electron Devices presented at the IRE Conference on Electron Tube Research, Ottawa, Canada, June 16–17, 1952 & IRE Western Convention, Long Beach (84 pages)	1.60	2.40	4.80
	PGED-4: December 1953 (62 pages)	1.30	1.95	3.90
	Vol. ED-1, No. 1, February 1954 (72 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 2, April 1954 (75 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 3, August 1954 (77 pages)	1.40	2.10	4.20
	Vol. ED-1, No. 4, December 1954: Papers presented at the 1954 Symposium on Fluctuation Phenomenon in Microwave Sources, November 18–19, 1954, New York, N. Y. (280 pages)	3.20	4.80	9.60
Engineering Management	PGEM-1: February 1954 (55 pages)	1.15	1.70	3.45
	PGEM-2: November 1954 (67 pages)	1.30	1.95	3.90
Industrial Electronics	PGIE-1: August 1953 (40 pages)	1.00	1.50	3.00

* Public libraries, colleges, and subscription agencies may purchase at IRE member rate.

(*Continued on the following page*)

Transistor Applications Symposium Will Meet Next Month

The University of Michigan Summer Transistor Applications Symposium will consist of a four-week lecture and conference program offered in co-operation with the Bell Telephone Laboratories, and partially supported by the Army. It will be part of the 1955 summer session of the university and will adhere to the pattern of somewhat similar programs held in Ann Arbor in 1937 and 1950. Within the university, the symposium is operated by the Electrical Engineering Department and directed by Professor W. G. Dow.

The core of the symposium will be two parallel four-week lecture sequences offered at 9:00 and 11:00 a.m. four days a week, beginning Tuesday, July 5, and extending through Friday, July 29. These lectures will treat various aspects of transistor circuit application. About half the lectures will be given by visiting lecturers of Bell Telephone Laboratories, and half by members of the faculties of various universities. In afternoon problem laboratory sessions, the lecture material will be worked into illustrative engineering problems. Opportunities for informal conferences will be provided.

The work of the symposium will be on the graduate level and will carry credit in the Horace H. Rackham School of Graduate Studies for those who desire it and, who qualify in advance for admission to the graduate school.

The fee for the complete course will be $200. For those properly certified as full-time faculty members or graduate students the fee will be $75. This reduced fee is provided through the Bell Telephone Laboratories. All fees are the same for residents of Michigan as for non-residents.

Requests for additional information should be addressed to Professor W. G. Dow, Electrical Engineering Department, Room 2501 East Engineering Building, University of Michigan, Ann Arbor, Mich.

Autumn Conference on Aeronautical and Navigational Electronics to Meet in Baltimore

Plans for *The 1955 East Coast Conference on Aeronautical and Navigational Electronics*, to be held October 31 and November 1, have been announced by the Baltimore Section and the Professional Group on Aeronautical and Navigational Electronics. The conference will be held at the Lord Baltimore Hotel in Baltimore.

The technical portion of the conference will be devoted to Aeronautical and Navigational Electronics. Interested persons are invited to submit papers for consideration: 150 word abstracts should be sent no later than July 1 to Norman Caplan, Bendix Radio Division of Bendix Aviation Corporation, Towson 4, Maryland.

Industrial exhibits are planned as an integral part of the conference. During the two days in November, representatives of many East Coast industries and development organizations will be concentrated in Baltimore to view these exhibits. Organizations interested in sponsoring exhibits should contact C. E. McClellan, Air Arm Division, Westinghouse Electric Corporation, Friendship Airport, Baltimore, Maryland

In addition to the technical programs and exhibits, a banquet will be held on the evening of October 31.

Information concerning advance conference registrations and hotel reservations may be obtained from G. R. White, Bendix Radio Division, Bendix Aviation Corporation, Towson 4, Maryland.

RETMA to Sponsor Automation Symposium in Philadelphia, Pa.

A two-day automation symposium sponsored by RETMA will present the current status of systems, machinery, and components to aid in the automatic fabrication of electronic equipment, and will show the part played by electronics in the automation of other industries. The theme of this year's symposium, to be held at the University of Pennsylvania in Philadelphia, September 19–20, is "Electronics for Automation and Automation for Electronics."

Following is the program for the symposium, the chairmen of the five sessions, and a description of the proposed sessions:

Monday, Sept. 19—*Session I: Mechanization of High Volume Assembly*, J. Harrington, Jr. This session will discuss mechanization of assembly of electronic equipment in lot sizes found in the TV and entertainment radio receiver fields. Working systems will be described, with emphasis on the mechanical design characteristics necessary to insure reliable operation. The extension of these mechanisms to military and industrial electronics applications of similar lot sizes also will be discussed.

Session II: Data Sensing, Processing and Utilization, D. A. Griffin. Electronics opens horizons on methods of rapid data sensing and processing which are having profound effects on the progress of automation in business routines and in industries other than that of electronics itself. This session will give electronics industry engineers a look at the scope of electronics in automation.

Session III: The Future of Automation, W. R. G. Baker. Automation can be expected to have a tremendous impact on our

TRANSACTIONS OF IRE PROFESSIONAL GROUPS
(Continued)

Sponsoring Group	Publications	Group Members	IRE Members	Non-Members*
Information Theory	PGIT-2: A Bibliography of Information Theory (Communication Theory-Cybernetics) (60 pages)	$1.25	$1.85	$ 3.75
	PGIT-3: March 1954 (159 pages)	2.60	3.90	7.80
	PGIT-4: September 1954 (234 pages)	3.35	5.00	10.00
Instrumentation	PGI-2: Data handling Systems Symposium: IRE Western Electronic Show & Convention, Long Beach, Calif., August 27–29, 1952 (111 pages)	1.65	2.45	4.95
	PGI-3: August 1954 (55 pages)	1.05	1.55	3.15
Microwave Theory & Techniques	Vol. MTT-1, No. 2, November 1953 (44 pages)	.90	1.35	2.70
	Vol. MTT-2, No. 1, Papers presented at the Microwave Radio Relay Systems Symposium, New York, N. Y., April 1954 (107 pages)	2.20	3.30	6.60
	Vol. MTT-2, No. 2, July 1954 (67 pages)	1.25	1.85	3.75
	Vol. MTT-2, No. 3, September 1954: Papers presented at the Joint IRE Professional Group—URSI meeting, Washington, D. C., May 5, 1954 (54 pages)			
	Vol. MTT-3, No. 1, January 1955 (47 pages)	1.50	2.25	4.50
Nuclear Science	Vol. NS-1, No. 1, September 1954 (42 pages)	.70	1.00	2.00
Reliability and Quality Control	PGQC-1: Papers presented at the 1951 Radio Hall meeting, and 1952 IRE National Convention (58 pages)	1.20	1.80	3.60
	PGQC-2: March 1953 (51 pages)	1.30	1.95	3.90
	PGQC-3: February 1954 (39 pages)	1.15	1.70	3.45
	PGQC-4: December 1954 (56 pages)	1.20	1.80	3.60
Telemetry and Remote Control	PGRTRC-1: August 1954 (16 pages)	.85	1.25	2.55
Ultrasonics Engineering	PGUE-1: June 1954 (62 pages)	1.55	2.30	4.65
	PGUE-2: November 1954 (43 pages)	1.05	1.55	3.15
Vehicular Communications	PGVC-2: Symposium on What's New in Mobile Radio (32 pages)	1.20	1.80	3.60
	PGVC-3: Theme—Spectrum Conservation, Washington, D. C., December 3–5, 1952 (140 pages)	3.00	4.50	9.00
	PGVC-4: Design, Planning and Operation of Mobile Communications Systems (June 1954) (98 pages)	2.40	3.60	7.20

* Public libraries, colleges, and subscription agencies may purchase at IRE member rate.

way of life. This forum discussion, in contrast to the technical papers of the other sessions, will explore via qualified panel of experts, the expected effects of automation on our society in management, education, sociology, economics and the military.

Tuesday, Sept. 20—*Session IV: Automation for Low Volume Production*, D. Cottle. Contrasted to long run production systems, those designed for short run production stress versatility rather than specialization. The papers in this session will describe several approaches to the problem of building in this versatility while meeting speed and reliability requirements.

Session V: Redesign for Automation of Components and Products, W. Hannahs. As automatic assembly of both high and low volume production becomes more common, the need for maximum efficiency and economy will stimulate new approaches in components and new approaches in design methods. This session presents some of these new components, new products, and new approaches already oriented to automation.

Technical Tape Recordings Now Available for Sections

During the past eighteen months a number of organizations, in cooperation with the IRE, have made tape recordings of technical talks for use by Sections and Student Branches. Full details on the subjects of the talks, length, number of slides, and how to obtain them are given in the IRE Tapescripts Manual which is available on request from the Institute, 1 East 79 Street, New York City. A list of tapescripts and their sponsors is given below.

Bell Telephone Laboratories, Inc.—Experiments With Linear Prediction in Television, C. W. Harrison. Statistics of Television Signals, E. R. Kretzmer. Efficient Coding, B. M. Oliver. The Physics of Music and Hearing, Winston E. Kock. Transistors in Negative Impedance Circuits, J. G. Linvill. A Junction Transistor Tetrode for High Frequency Use, R. L. Wallace, Jr. Some Circuit Properties of Junction Transistors, L. G. Schimpf. Ferroelectric Storage Devices, J. Reid Anderson. The Bell Solar Battery, Gordon Raisbeck.

General Electric—Germanium—The Magic Metal, J. S. Saby and J. P. Jordan.

Naval Ordnance Laboratory—New Magnetic Materials, Carroll W. Lufcy. Magnetic Amplifier Analysis and Design, Carroll W. Lufcy and Herbert H. Woodson. Physical Problems at Hypersonic Speeds, H. H. Kurzweg.

Polytechnic Research and Devel. Co.—Microwave Techniques, H. C. Nelson.

IRE PG on Audio—Method for Time or Frequency Compression—Expansion of Speech, Grant Fairbanks, W. L. Everitt and R. P. Jaeger. Magnetic Recording, Marvin Camras. Phonograph Pickups, B. B. Bauer. Push-Pull Single Ended Audio Amplifier, Arnold Peterson and D. B. Sinclair. An Experimental Co-Channel Television Booster Station Using Crossed Polarization, John H. DeWitt, Jr.

RCA Laboratories—Color Television, D. H. Ewing. The Germanium Story, S. M. Christian. A P-N-P Alloy Junction Transistor for Radio-Frequency Amplification, C. W. Mueller. An Experimental Transistor Personal Broadcast Receiver, L. E. Barton.

Obituary

Rear Admiral Stanford C. Hooper (F'28–A'33–F'46) died recently at his winter home in Miami Beach. He was 70 years old.

S. C. Hooper

Admiral Hooper received many civilian and military honors, primarily for his contributions to the development of radio in Navy. He was retired in 1943 because of a disability incurred in World War II.

He was born in Colton, California in 1884. In 1905 he was graduated from the United States Naval Academy and then instructed there in physics, chemistry, and electricity. Later he served for two years as the first Fleet Radio Officer, resuming that post again from 1923 to 1925. He was in charge of the Radio Division of the Navy Department for eleven years and Director of Naval Communication from 1928–1934.

Admiral Hooper was a leader in developing wireless radio communications in the Navy by carrying out pioneer tests, establishing a chain of land stations for communication between fleet and land, and serving as technical advisor and head of many boards and committees dealing with communications. He suggested the position of Fleet Radio Officer as necessary to the new radio Communications, and served in this post for two years. In the first World War Admiral Hooper was awarded the Navy Cross for distinguished service as commanding officer of the U.S.S. *Fairfax*.

The Franklin Institute of Philadelphia presented him the Eliot Cresson Medal for research in radio electronics in 1945. In 1948 Admiral Hooper received an honorary LL.D. from Drury College. He also received the French Legion of Honor, the Department of Navy Electronics Trophy, and one of few to hold the Marconi Medal of Merit.

Report of the Secretary—1954

To the Board of Directors,
The Institute of Radio Engineers

Gentlemen:

Another annual report is herewith submitted which depicts the status of IRE affairs at the end of the year just closed.

Your attention is directed to the increases in membership, both of the IRE and of the Professional Groups, which are larger percentagewise than normally. The National Convention was the most extensive held both for attendance and the number of exhibitors. In fact, all activities, as the report shows, are at an increased tempo.

The housing situation for the headquarters staff and for committee activities was alleviated during 1954 by the completion of the new building and its facilities at 5 East 79th Street, adjacent to the existing building.

Respectfully submitted,

Haraden Pratt
Secretary

January 24, 1955

Membership

At the end of the year 1954, the membership of the Institute, including all grades, was 41,778, an increase of 4,644, or 12½% over the previous year. The 4,644 member increase in 1954 was more than 4,260 and 3,466, the increases for 1953 and 1952 respectively. The percentage increase was 13% in 1953 and 12% in 1952. The membership trend from 1912 to date is shown graphically in Figure 1.

Actual membership figures for 1952, 1953, and 1954 are shown in Table I. Of the 24,043 non-voting Associates, 6,244 have been in that grade for more than five years.

It is with deep regret that this office records the death of the following members of the IRE during the year 1954.

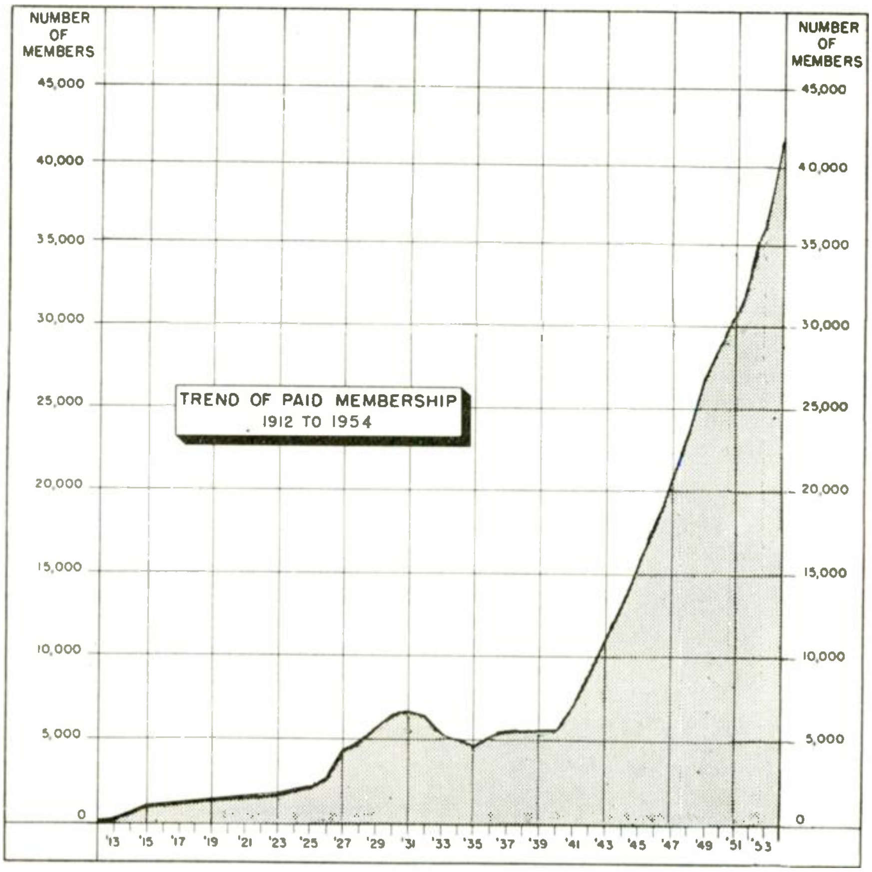

Fig. 1

TABLE I

Total Membership Distribution by Grades, 1952–1954

Grade	As of Dec. 31, 1954		As of Dec. 31, 1953		As of Dec. 31, 1952	
	Number	Per Cent of Total	Number	Per Cent of Total	Number	Per Cent of Total
Fellow	489	1.2	425	1.2	379	1.2
Senior Member	4,780	11.4	4,170	11.2	3,566	10.8
Member	6,107	14.6	5,307	14.3	4,617	14.0
Associate	24,846‡	59.5	22,702†	61.1	20,029*	61.0
Student	5,556	13.3	4,530	12.2	4,283	13.0
Totals	41,778		37,134		32,874	

* Includes 919 Voting Associates. † Includes 856 Voting Associates. ‡ Includes 803 Voting Associates.

Fellows

Armstrong, Edwin H. (A'14, F'27)
Brennecke, Cornelius G. (A'36, M'46, F'51)
Kilgour, C. E. (A'25, M'30, SM'43, F'51)
Mouromtseff, Ilia E. (A'34, SM'45, F'47)
Rush, Walter Albert (M'13, F'24, L'48)

Senior Members

Ainsworth, Archer L. (A'22, M'33, SM'43)
Allison, John L. (SM'47)
Babcock, Stuart M. (A'36, VA'39, SM'52)
Beranek, Jerome A. (S'41, A'42, M'46, SM'48)
Bernreuter, Herbert A. (SM'49)
Burgess, Warren B. (A'20, M'31, SM'43)
Fry, Howard M. (A'46, SM'51)
Garrison, Millard M. (M'35, SM'43)
Gunzbourg, Paul M. (M'43, SM'43)
Hancock, George N. (M'43, SM'43)
Herz, Armin J. (M'49, SM'53)
Holborn, Frederick (M'23, SM'53)
Jones, Walter Roy (A'26, M'32, SM'43)
Lewis, Oliver Irving (M'45, SM'45)
Parker, Thomas J. (SM'46)
Tuller, W. G. (S'37, A'40, M'45, SM'48)

Members

Cook, William F. (M'46)
DeSutter, David H. (M'52)
Goldin, Hyman C. (A'27, M'44)
Groom, Bryan (M'46)
Mack, Pearcy W. (M'47)
Melnicoe, Samuel A. (A'41, M'46)
Patremio, Salvatore R. (M'50)
Robertson, Raymond D. (A'42, M'51)
Slavin, Samuel (A'29, M'48)
Tavaniotis, Constantine (A'47, M'48)
Williams, Roger W. (M'50)

Voting Associates

Porter, Roland Guyer (A'26, VA'39)

Associates

Belle Isle, Armand G. (A'41)
Bonk, George V. (S'52, A'54)
Briscoe, William C. (A'52)
Budelman, Frederick T. (A'41)
Clark, Bayard H. (A'46)
Crockett, John E. (S'49, A'50)
Daum, John A. (A'51)
Fanyo, Wayne P. (A'51)
Ferrell, Dewey (A'46)
Fleer, William A. (A'52)
Garland, O. K. (A'40)
Green, Harry (A'48)

Hammond, Arthur L. (A'44)
Hanson, Louis W. (A'53)
Harris, William C. (A'47)
Hartmann, Charles L. (A'44)
Helmers, Edgar C. (A'52)
Hopf, Edgar W. (A'49)
Hurford, Alvin C., Jr. (A'53)
Jones, Robert E. (S'45, A'47)
Jorgensen, Andrew C. (A'45)
Kraus, Russell B. (A'52)
Martin, William H. (A'49)
Martinez, Clement I. (A'45)
Neuman, Leonard J. (A'44)
Paul, Fred Herman (A'53)
Rogers, Keith Sinclair (A'47)
Teeter, Albert A. (A'24)
Tindall, John Ronald (A'52)
Vassian, Herbert J. (S'49, A'54)
Voegtlin, Elmo (A'37)
White, Robert H. (A'40)
Winkler, William H. (S'49, A'53)
Wolcott, Glenn W. (A'51)
Woodman, Allan J. (A'53)

Students

Carrell, Sanford G. (S'52)
Herzer, Melvin (S'53)
Nye, Henry A., Jr. (S'53)
Stewart, Phil M. (S'52)

Fiscal

A condensed summary of income and expenses for 1954 is shown in Table II, and in Table III a balance sheet for 1954.

TABLE II

SUMMARY OF INCOME AND EXPENSES, 1954

Income		
Advertising	$774,765.52	
Member Dues and Convention	846,892.17	
Subscriptions	107,113.67	
Sales Items—Binders, Emblems, etc.	63,338.89	
Investment Income	20,088.33	
Miscellaneous Income	874.84	
TOTAL INCOME		$1,813,073.42
Expense		
PROCEEDINGS Editorial Pages	$287,831.17	
Advertising Pages	399,042.18	
Directory	153,633.38	
Section and Student Branch Rebates	64,314.29	
Professional Group Expense	68,563.34	
Sales Items	41,549.04	
General Operations	316,002.34	
Convention Cost	265,556.67	
TOTAL EXPENSE		1,596,492.41
Reserve for Future Operations—Gross		$ 216,581.01
Reserve for Depreciation		13,469.05
Reserve for Future Operations—Net		$ 203,111.96

TABLE III

BALANCE SHEET—DECEMBER 31, 1954

Assets		
Cash and Accounts Receivable	$396,052.27	
Inventory	17,451.96	
TOTAL CURRENT ASSETS		$ 413,504.23
Investments at Cost	774,241.75	
Buildings and Land at Cost	910,337.92	
Furniture and Fixtures at Cost	166,649.93	
Other Assets	19,855.23	
TOTAL		1,871,084.83
TOTAL ASSETS		$2,284,589.06
Liabilities and Surplus		
Accounts Payable		$ 21,662.30
*Deferred Liabilities	$540,984.97	
Professional Group Funds on Deposit	109,885.04	650,870.01
TOTAL LIABILITIES		672,532.31
Reserve for Depreciation		59,849.27
Surplus Donated	595,286.61	
Surplus	956,920.87	
TOTAL SURPLUS		1,552,207.48
TOTAL LIABILITIES AND SURPLUS		$2,284,589.06

* 1955 Items, PROCEEDINGS for Members and Subscribers, Advertising and Convention Service.

Editorial Department

IRE publication activities during 1954 were marked by a large increase in volume and important service improvements.

Due principally to the large growth of the transactions of the professional groups, which nearly doubled in volume, the total IRE publication output soared to 79 issues totaling 10,486 pages, a 25% increase over the previous year. It is interesting to note that as recently as 1951, the corresponding totals were only 19 issues and 3,914 pages, indicating that the IRE publication program has expanded 2½ times in the brief span of 3 years.

Two noteworthy improvements in service were instituted during the year: a new publication for student members, the *IRE Student Quarterly*, was successfully launched; the average publication time of PROCEEDINGS papers was cut in half.

PROCEEDINGS OF THE IRE

The Editorial Board began a continuing study of the long-term roles of THE PROCEEDINGS and the transactions. As a result of their initial deliberations, several steps were taken to avoid an overlap of THE PROCEEDINGS and the transactions and, at the same time, to strengthen the coordination between them. These actions involved placing stringent restrictions on reprinting transactions papers in THE PROCEEDINGS, participation of professional group representatives in the review of PROCEEDINGS papers, and the publication in PROCEEDINGS of abstracts of all transactions papers.

Table IV (next page) and Figure 2

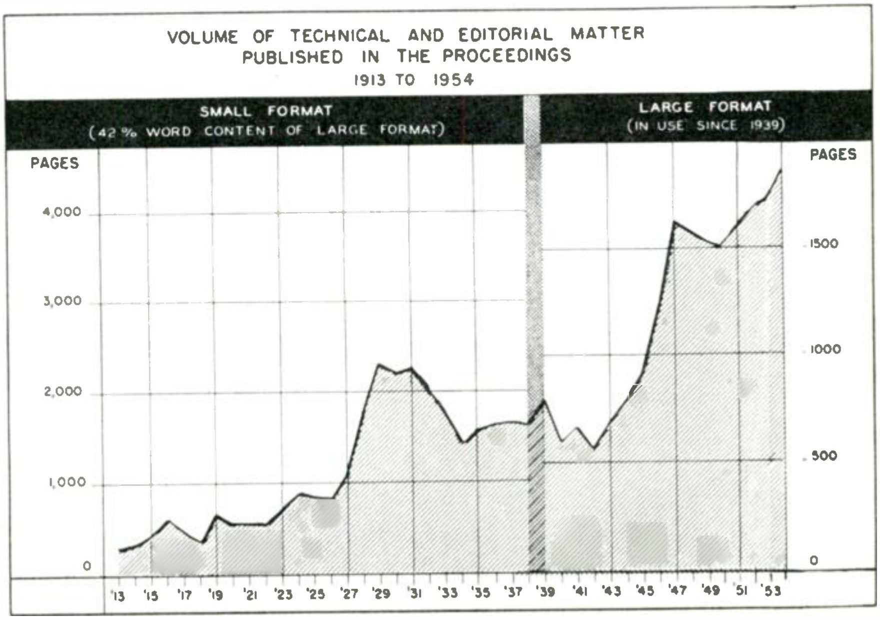

Fig. 2

each show that the number of PROCEEDINGS pages published during 1954 remained about the same as the previous year, with the total of editorial pages showing a slight increase. However, the number of papers, 177, was less than the previous year's total of 212. Nine IRE Standards also appeared. The year's highlight was the appearance of a special jumbo issue on Color Television in January.

TABLE IV
VOLUME OF PROCEEDINGS PAGES

	1954	1953	1952	1951
Editorial	1884	1860	1804	1628
Advertising	2072	2146	1800	1424
Total	3956	4006	3604	3052

Early in the year, the procedure for reviewing papers was considerably streamlined and, as a result, the average time required for publishing PROCEEDINGS papers was reduced from 10 months to 5. In addition, the backlog of accepted material awaiting publication was reduced from the equivalent of 2 issues to a small fraction of an issue, thus helping to speed up publication of papers.

The volume of material reviewed for PROCEEDINGS was less than in the previous year, reflecting the increasingly prominent role being played by the transactions. A total of 305 papers comprising 1,988 pages was considered, as against 376 papers and 2,621 pages in 1953. Of the total, 38% of the papers were accepted, 30% were referred to the appropriate transactions for consideration, and 32% were rejected.

The *Annual Review of Radio Progress*, which formerly appeared in April of each year, was discontinued. In its stead, a series of review papers, prepared by experts in various fields, was started. It is planned to publish additional review papers on each subject at appropriate intervals when sufficient progress has been made since the last paper was published.

Transactions

The year 1954 saw the transactions output of the Professional Groups increase almost 100%, the greatest increase since their inception three years ago. The year also saw an improvement in the transactions themselves, with 6 groups utilizing letterpress composition for improved appearance and 10 groups issuing transactions on a regular schedule.

As a result of this major increase in activity, the Editorial Department published 51 issues of transactions totaling 3,714 pages for 18 groups during 1954, as compared with the 1953 totals of 32 issues totaling 1,798 pages for 15 groups. A breakdown of transactions material published in 1953 and 1954 is given in Table V.

Convention Record of the IRE

The practice of publishing the *Convention Record* containing papers presented at the National Convention, begun in 1953, was repeated. The *1954 Convention Record*, containing 213 papers, 29 abstracts and totaling 1,436 pages, was issued in eleven parts. Every paid member of a professional group received free of charge a copy of that part pertaining to the field of interest of his group.

TABLE V
VOLUME OF TRANSACTIONS PAGES

Group	1954 Issues	1954 Pages	1953 Issues	1953 Pages
Aeronautical and Navigational Electronics	4	144	3	84
Antennas and Propagation	4	188	2	72
Audio	6	208	6	180
Broadcast and Television Receivers	4	312	3	160
Circuit Theory	4	256	1	112
Communications Systems	3	414	1	76
Component Parts	2	172	0	0
Electron Devices	4	524	3	184
Electronic Computers	4	228	4	136
Engineering Management	2	132	0	0
Industrial Electronics	0	0	1	44
Information Theory	2	404	2	290
Instrumentation	1	60	1	116
Medical Electronics	0	0	1	44
Microwave Theory and Techniques	3	244	2	100
Nuclear Science	1	48	0	0
Reliability and Quality Control	2	104	1	56
Telemetry and Remote Control	2	56	0	0
Ultrasonics Engineering	2	116	0	0
Vehicular Communications	1	104	1	144
Totals	51	3714	32	1798

IRE Student Quarterly

A new publication for student members, the *IRE Student Quarterly*, made its first appearance in September, followed by a second issue in December. The publication is sent free to all student members of the IRE as part of a program of increased services to students.

Joint Conference Publications

As a result of offering its services and facilities to conferences jointly sponsored by the IRE and other societies, the Editorial Department published the 132 page *Proceedings of the Eastern Joint Computer Conference*, held in Washington, D. C., December 8–10, 1953, by the IRE, AIEE, and the Association for Computing Machinery.

IRE Directory

The 1954 *IRE Directory* was published in October, containing 1,044 pages including covers, of which 429 were membership listings and information, and 615 were advertisements and listings of manufacturers and products.

IRE Cumulative Index

IRE cumulative indexes were brought up to date with the publication of a cumulative index for 1948–1953, covering the transactions and *Convention Record*, as well as THE PROCEEDINGS.

Technical Activities

Technical Committees

During 1954, the 22 IRE technical committees with their 89 subcommittees held 197 meetings, of which 177 met at IRE Headquarters and 20 elsewhere.

The Annual Review Committee was disbanded by the Executive Committee in April, 1954. Papers are now solicited by the Editorial Board to cover fields formerly covered by this committee.

At the time of this report, the Standards Committee is considering the formation of new Technical Committees on Electromagnetic Interference and on Nuclear Science. The subject of interference received wide attention in committee work in 1954 and the entire industry is concerned with the problem.

The Standards listed herewith were published in THE PROCEEDINGS in 1954 and reprints are now available to the public.

March—

53 IRE 4. S1

Standards on Circuits: Definitions of Terms in the Field of Linear Varying Parameter and Non-Linear Circuits, 1953.

53 IRE 19. S2

Standards on Sound Recording and Reproducing: Methods for Determining Flutter Content, 1953 (ASA Z57.1-1954).

May—

53 IRE 3. S2

Standards on American Recommended Practice for Volume Measurements of Electrical Speech and Program Waves, 1953 (ASA C16.5-1954).

June—

54 IRE 21. S1

Standards on Graphical Symbols for Electrical Diagrams, 1954 (ASA Y32.2-1954).

July—

54 IRE 3. S1

Standards on Audio Techniques: Definitions of Terms, 1954.

54 IRE 23. S1

Standards on Television: Methods of Measurement of Aspect Ratio and Geometric Distortion, 1954 (ASA C16.23-1954).

August—

54 IRE 7. S1

Standards on Electron Devices: Definitions of Terms Related to Phototubes, 1954.

September—

54 IRE 17. S1

Standards on Receivers: Methods of Measurement of Interference Output of Television Receivers in the Range of 300 to 10,000 kc, 1954.

October—

54 IRE 7. S2

Standards on Electron Devices: Definitions of Semi-conductor Terms, 1954.

The IRE Standards on Radio Aids to Navigation: Definitions of Terms, 1954 (54 IRE 12. S1) were given final approval in December, 1954 and will appear in the February, 1955 PROCEEDINGS.

During 1954 a new Measurements Index was set up to supplement the Master Index of IRE Definitions, first issued in 1949. The new document will appear in the PROCEEDINGS during the first part of 1955 with the title *Index to Measurement Methods Described in IRE Printed Standards.* The fourth edition of the *Master Index of IRE Definitions* was distributed in January, 1955.

IRE is represented on 27 American Standards Association Committees and sponsors two of them: the ASA Sectional Committee on Radio and Electronic Equipment, C16, and the ASA Sectional Committee on Sound Recording, Z57. Six IRE Standards received approval of the American Standards Association as American Standards in 1954, and are now available to international standards organizations.

Interest in international standardization was renewed when many IRE committee members participated in the Golden Jubilee Meeting of the International Electrotechnical Commission in Philadelphia during September, 1954. Several members of IRE participated in the meetings of URSI at the Hague in August, 1954.

Professional Group System

General. There are presently 23 Professional Groups actively operating within the framework of the IRE. Two new groups were established in 1954, the Professional Group on Production Techniques and the Professional Group on Automatic Control. More than 50% of all IRE members have taken advantage of the Professional Group system which now has a total membership of 31,797. 593 Student members of the IRE have joined the Groups at the Student member rate of $1.00 for one year covering their particular field of interest.

Twenty-one groups levy publications assessments and 28,158 members have paid these assessments and are receiving the pertinent group transactions regularly. In addition, 361 companies, university and public libraries have subscribed to the transactions of all the Groups.

The IRE has continued to leave the formation and activities of individual Groups to the initiative of interested IRE members. Supplementary financial assistance is provided by Headquarters and provision has been made for reimbursing Sections for Chapter meetings. Included in the many services rendered the professional groups during 1954 were 602 mailings by headquarters to the Groups' membership.

Symposiums. The procurement of papers and the management of Symposiums is now entirely in the hands of Professional Groups. Each of the Professional Groups has sponsored one or more technical meetings of national import each year, and in addition, technical sessions at the IRE National Convention, the WESCON, and the National Electronics Convention, for a total of 70 national meetings in 1954.

Publications. The official publications of the Groups are called IRE Transactions. 20 Groups are currently publishing transactions and to date 113 issues have appeared. In 1954 20 Groups published 53 transactions containing 3,490 pages. Ten of the Groups are now on a regularly stated publication schedule.

Professional Group Chapters. The interest of Group members within Sections has resulted in the organization of 117 Professional Group Chapters in 27 IRE Sections to date. All of the Chapters are meeting regularly and sponsoring Section meetings in the fields of interest of their associated groups.

Appointed Institute Representatives on other Bodies

The IRE appointed representatives on a number of other bodies for a one-year period: May 1, 1954 to April 30, 1955 (as listed on page 771 of this issue) and worked closely with international bodies such as URSI, CCIR, and the International Electrotechnical Commission.

The Joint Technical Advisory Committee

The Joint Technical Advisory Committee and its subcommittees held a total of 12 meetings, in addition to a fiftieth meeting dinner.

Volume XI of the JTAC Proceedings was published in 1954. This includes Section I—Official Correspondence Between the Federal Communications Commission and the Joint Technical Advisory Committee (IRE-RETMA) with Other Items of Correspondence Pertinent to the Action of the JTAC and Section II—Approved Minutes of Meetings of the Joint Technical Advisory Committee for the period 1 July 1953 to 30 June 1954.

The JTAC Subcommittee on Land-Mobile Channeling Arrangements submitted a report to JTAC on December 10, 1954, which was transmitted to the FCC Chairman. This subcommittee was reactivated on February 12, 1954, in reply to the FCC inquiry of January 8, 1954, to Mr. A. V. Loughren, then Chairman of JTAC. Laboratory tests had been conducted and data collected to assist in answering the questions posed in Mr. Hyde's letter to Mr. Loughren.

The Interference Study established by JTAC in December 1952 at the request of Chairman Paul A. Walker of the FCC continued during 1954. JTAC considered the FCC's Notice of Proposed Rule Making in the Matter of the Commission's Docket No. 9288. The Committee reached conclusions which were considered pertinent to the proposed rule making and so presented these views to the FCC Chairman. These included opinions on (I) basic philosophy, (II) basic technical considerations, (the necessity for and practical limits of control) and (III) specific considerations arising under the proposed rules.

Two new subcommittees were formed by JTAC. The Interference Analysis Sub-committee 54.1 was organized on January 28, 1954, to examine the material regarding interference cases and to advise JTAC of other possible sources of information.

On September 23, 1954, a subcommittee 54.2 on RF Interference from Arc Welders was formed as the result of an FCC request to JTAC asking their assistance and advice in connection with certain problems concerning radio frequency stabilized arc welding apparatus. One meeting has been held so far.

Section Activities

We were glad to welcome three new Sections into the Institute during the past year. They are as follows:

Israel	October 1954
Ithaca	March 1954
Northern New Jersey	May 1954

The total number of Sections is now 74. There has been a membership increase in 57 of the 74 Sections.

The Ithaca and Northern New Jersey Subsections became full Sections in the year 1954. The Subsections of Sections now total 17, the following being formed in 1954:

Berkshire County (Connecticut Valley Section) January 1954
Northwest Florida (Atlanta Section) March 1954
Richland (Seattle Section) February 1954
Tucson (Phoenix Section) January 1954

Student Branches

The number of Student Branches formed during 1954 was 3, 1 of which operates as an IRE Branch, and 2 of which operate as Joint IRE-AIEE Branches. The total number of Student Branches is now 121, 91 of which operate as Joint IRE-AIEE Branches.

Following is a list of the Student Branches formed during the year: Colorado Agricultural and Mechanical College, Texas College of Arts and Industries, and Washington University.

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7.4 Camera Tubes, Phototubes, and Storage Tubes in which Photo-Emission is Essential

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7.5 High-Vacuum Microwave Tubes

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7.5.1 Non-Operating Characteristics of Microwave Tubes

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7.5.2 Operating Measurements of Microwave Oscillator Tubes

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27.8 Carrier Current Equipment

27.9 Community Antennas

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NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

The Index to the Abstracts and References published in the PROC. IRE from February, 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.

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

ACOUSTICS AND AUDIO FREQUENCIES

534.2 **1227**
Air-to-Ground Sound Propagation—P. H. Parkin and W. E. Scholes. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1021–1023; November, 1954.) "The attenuation of sound in air, in the vertical direction, has been measured on six occasions using an aircraft flying at various heights over a microphone at ground level. It was found that there was always negligible attenuation at frequencies below 1,000 c/s even though the air was turbulent; at higher frequencies the attenuation was found to be of the same order as the Knudsen-Kneser results for attentuation due to molecular absorption."

534.2-8 **1228**
Absorption of Finite-Amplitude Sound Waves—F. E. Fox and W. A. Wallace. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 994–1006; November, 1954.) Measurements indicate that the absorption coefficient of CCl_4 and water for ultrasonic waves increases with increasing intensity; the effect is attributed to waveform distortion. Theory is presented in which the medium is treated as the acoustic analog of a transmission line with nonlinear elements.

534.23+621.396.67 **1229**
Maximum Directivity Index of a Linear Point Array—R. L. Pritchard. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1034–1039; November, 1954.) The maximum directivity index of an array is computed as a function of number and spacing of elements. For spacings $<\lambda/2$ a significant improvement of both directivity index and major lobe width may be achieved by departing from uniform excitation. Numerical results are presented for three-, five- and seven-element arrays.

534.24-8 **1230**
Reflection of Sound by Thin Plates in Water—L. M. Lyamshev. [*Compt. Rend. Acad. Sci. (URSS)*, vol. 99, pp. 719–721; December 11, 1954. In Russian.] The nonspecular reflection of sound from brass and aluminium plates immersed in water was investigated experimentally using waves of frequency 1 mc in pulses of duration 30 μs and repetition rate 50 per second. Reflection at angles related to the velocity of propagation of longitudinal waves in the plates was observed, in addition to that related to flexural waves [611 of 1949 (Finney)]. In a 0.8-mm-thick Al plate the respective angles were 16 degrees and 43 degrees, in a 0.58-mm brass plate the reflection maxima due to longitudinal waves occurred at 23 degrees, those due to flexural waves at 66 degrees and 74 degrees; in a 0.38-mm brass plate, where the velocity of flexural waves is lower than the velocity of sound in water, only the maximum due to longitudinal waves was observed, at 23 degrees. Experimental results are presented graphically.

534.26+[535.42:538.566 **1231**
Diffraction of Pulses by a Circular Cylinder—F. G. Friedlander. (*Commun. Pure Appl. Math.*, vol. 7, pp. 705–732; November, 1954.) Analysis for a sound-pressure field is presented for the two-dimensional case, i.e., uniformity in the direction parallel to the cylinder axis is assumed. The incident, reflected and diffracted fronts are shown diagrammatically. The interpretation of the problem in em terms is indicated.

534.3/.4 **1232**
The Transient Tones of Wind Instruments—E. G. Richardson. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 960–962; November, 1954.) "Oscillograph records of the initiation of sound in organ pipes and orchestral wind instruments show large variations from one instrument to another. The causes of these variations and their artistic effects are discussed."

534.6:621.395.61 **1233**
Microphone Measurements in a Reverberation Chamber—L. Keidel. (*Arch. Tech. Messen*, no. 226, pp. 249–252; November, 1954.) Experiments are described which indicate that reverberation-chamber measurements can be substituted for anechoic-chamber measurements without introducing large errors. Suitable methods are outlined for determining microphone characteristics by comparison with a known microphone. Since the ratio of the sound pressure at the two microphones is measured, it is not necessary to maintain the sound level constant.

534.614:621.3.018.75 **1234**
Electronic Pulse Method for measuring the Velocity of Sound in Liquids and Solids—N. P. Cedrone and D. R. Curran. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 963–966; November, 1954.)

534.75 **1235**
Cumulative Effects of Repeated Bursts of White Noise on Threshold for 4000-c/s Tone Pips—C. Lightfoot and J. F. Jerger. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1048–1052; November, 1954.)

534.75 **1236**
Masking of Speech by Repeated Bursts of Noise—I. Pollack. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1053–1055; November, 1954.)

534.75 **1237**
Intensity Discrimination Thresholds under Several Psychophysical Procedures—I. Pollack. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1056–1059; November, 1954.) "Discrimination thresholds for the detection of a change in the sound level of a tone were obtained under five experimental procedures."

534.75 **1238**
Method of Reproduction and the Identification of Elementary Auditory Displays—I. Pollack. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1060–1063; November, 1954.) "A psychophysical procedure which combines the salient features of the classical discrimination and identification experiments is described."

534.76:534.86 **1239**
Effect of Arrival Time on Stereophonic Localization—W. B. Snow. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1071–1074; November, 1954.) Summarized report of experimental work previously published in U.S. Patent No. 2137032.

534.861 **1240**
Radio Studios—W. Furrer. (*Tech. Mitt. Schweiz. Telegr.-Teleph Verw.*, vol. 32, pp. 430–436; November 1, 1954. In German.) A brief review of the requirements as regards electro-acoustic equipment and acoustic characteristics, with illustrations of studio details.

621.395.623.5 **1241**
Experimental Investigation of Wedge Horns used with Line Hydrophones—C. M. McKinney and C. D. Anderson. (*Jour. Acoust.*

Soc. Amer., vol. 26, pp. 1040–1047; November, 1954.) Radiation patterns are presented for horns constructed of Al plates covered with cellular rubber set at angles up to 90 degrees. The frequency range investigated is 60–150 kc.

621.395.623.5 1242

Acoustical Studies of the Tractrix Horn: Part 1—R. F. Lambert. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1024–1028; November, 1954.) By assuming the constant-phase surfaces in the horn flare to be spheres, and the flow lines to be tractrixes, a one-parameter wave equation and Ricatti impedance equation are derived. Comparison of measured values of axial response and throat impedance with calculations based on a hemispherical and on a plane piston indicate that the actual configuration is intermediate between these two.

621.395.623.5 1243

Acoustical Studies of the Tractrix Horn: Part 2—A. O. Jensen and R. F. Lambert. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 1029–1033; November, 1954.) Measurements of the free-field radiation characteristics are reported. Small and large baffle mountings were investigated. Results for two-cell units indicate a performance comparable with that of the exponential horn. Multicellular structures give improved uniformity of angular distribution at high frequencies, but exhibit undesirable band-rejection characteristics within the useful frequency range. Part 1: 1242 above.

621.395.625.6 1244

An Optical Method for Reducing the Noise inherent in Sound Films—V. S. Man'kovski. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2025–2035; November, 1954.) A complex electrical device is normally used in sound recording for varying the average density of the film with the modulation level. The same result can be achieved with the aid of a simple optical system in the light modulator, based essentially on the use of an additional wide slot. The theory of the method is discussed in detail and results are given of experiments indicating that the reduction of noise so obtained is equivalent to that given by the electrical method (10 db).

ANTENNAS AND TRANSMISSION LINES

621.315.212:621.397 1245

Progress in the Statistical Treatment of Characteristic-Impedance Irregularities in Television Cables—H. Kaden. (*Arch. elekt. Übertragung*, vol. 8, pp. 523–529; November, 1954.) Formulas obtained previously for the far-end effect (2070 of 1953) were based on the assumption of an exponential autocorrelation function for the cable irregularities. The theory is now generalized to apply to an arbitrary autocorrelation function, and the special case of a Gaussian error function is examined. Permissible values of irregularity are tabulated for wideband and for feeder cable. For the latter type the outer conductor should be of such design that the correlation range is >1m, in which case the requirements as regards uniformity are not too stringent. The reflection coefficient at the antenna should be <1.6 per cent. A rms value of 1 per cent is assumed for the far-end effect in the tables.

621.315.212.1:621.315.682 1246

Coaxial Plug Connectors for Communication Engineering—H. Larsen and F. Haas. (*Frequenz*, vol. 8, pp. 267–276; September, 1954.) The construction of connectors for use at frequencies up to 3kmc is described and discussed. The particular problem of connecting 60-Ω to 75-Ω lines is also discussed and calculations show that reflection can be neglected when using the 6/16-mm (internal/external diameters respectively) cable connectors. The characteristics of the 2.5/6.5-mm and the 11/30-mm standard connectors are also given.

621.372.2 1247

Electromagnetic Wave Propagation on Helical Structures (a Review and Survey of Recent Progress)—S. Sensiper. (Proc. IRE, vol. 43, pp. 149–161; February, 1955.) Includes 58 references.

621.372.2 1248

Electromagnetic Processes in a Multiconductor System—P. I. Kuznetsov and R. L. Stratonovich. [*Bull. Acad. Sci. (URSS), Tech. Sci.*, pp. 3–23; September, 1954. In Russian.] The em field in a multiconductor system is determined by using Maxwell's equations with boundary conditions given by a set of inequalities relating the electric and magnetic constants of the conductor and the surrounding space. The impedance and admittance matrixes of the telegraphy equation are derived and the conditions for the equivalence of Maxwell's and Kirchhoff's equations are deduced. The theory is applied to the case of a two-conductor system parallel to a plane ideally conducting surface, and simplification is obtained by putting $a \ll l$, where a is the radius of the conductors and l is their separation or distance from the surface; the error introduced by this simplification is of the order of $(a/l)^2$. The interaction of signals transmitted along two parallel conductors is also briefly considered.

621.372.2:621.315.212 1249

Coaxial Line with Helical Inner Conductor—W. Sichak. (Proc. IRE, vol. 43, p. 148; February, 1955.) Correction to paper abstracted in 3123 of 1954.

621.372.43 1250

Theory and Design of Wide-Band Multisection Quarter-Wave Transformers—R. E. Collin. (Proc. IRE, vol. 43, pp. 179–185; February, 1955.) "A general theory of the n-section quarter-wave transformer is presented. It is shown that optimum bandwidth with a minimum pass band tolerance is obtained when the power loss ratio is chosen to give Tchebycheff behavior in the pass band. A comparison is made of the Tchebycheff transformer and the maximally flat transformer, and shows that the former gives a large increase in bandwidth—e.g., up to 44 per cent for a 2-section transformer and up to 75 per cent for a 4-section transformer. Design formulas are given for the 2, 3 and 4-section transformers."

621.372.8 1251

Propagation Constant in Rectangular Wave Guide of Finite Conductivity—D. M. Kerns and R. W. Hedberg. (*Jour. Appl. Phys.*, vol. 25, pp. 1550–1551; December, 1954.) A formula is presented which holds for frequencies both above and below cut-off.

621.372.8 1252

H-Plane Bifurcation of Rectangular Waveguides—R. A. Hurd and H. Gruenberg. (*Canad. Jour. Phys.*, vol. 32, pp. 694–701; November, 1954.) "Using a method based on the calculus of residues, a rigorous solution has been obtained for the problem of the bifurcation of a rectangular waveguide. Expressions are given for the amplitudes of all the reflected and transmitted modes in the guide. A comparison is made with results obtained by the transform method of Wiener and Hopf."

621.372.8 1253

Propagation in Curved and Twisted Waveguides of Rectangular Cross-Section—L. Lewin. (*Proc. IEE*, part B, vol. 102, pp. 75–80; January, 1955.) The problem is investigated by putting the wave equation in a form in which the coordinates in a waveguide cross section are also the independent variables in the differential equation. The wavelength is found for the *E*-curved and *H*-curved guides, and for twisted guides. In the latter case there is a degeneracy when the guide sides become equal. Curves are presented for the correction to the wavelength in a straight guide caused by the bending or twisting.

621.372.8 1254

Waveguide Phase Changer—R. E. Collin. (*Wireless Eng.*, vol. 32, pp. 82–88; March, 1955.) Various known types of phase changer are surveyed briefly and a description is given of a new type in which the phase shift is produced by longitudinal displacement of a central dielectric strip in a rectangular waveguide with fixed side dielectric strips. Over the wavelength range 3–3.5 cm, the departure from linearity is <1 per cent and the swr is <1.05.

621.372.8:621.317.7 1255

Waveguides and Waveguide Junctions—J. M. C. Dukes. (*Wireless Eng.*, vol. 32, pp. 65–72; March, 1955.) Problems relating to microwave measurements made through mismatched junctions are considered. The analysis is performed using simple plane geometry, the junction being shown to constitute a bilinear transformation. The required measurement and calculation procedure is detailed. See also various papers by Deschamps (e.g. 14 of 1954).

621.372.8:621.318.134 1256

Analogy of Ferrite in Waveguide—G. F. Nicholson. (*Wireless Eng.*, vol. 32, p. 89; March, 1955.) Asymmetrical attenuation produced by a ferrite rod in a waveguide with a steady transverse magnetic field is explained by using as model a gyroscope suspended at its center of gravity.

621.372.8.012.3 1257

Nomographs for Rectangular Waveguides—T. S. Chen. (*Electronics*, vol. 28, pp. 172–176, 178; January, 1955.)

621.396.67+534.23 1258

Maximum Directivity Index of a Linear Point Array—Pritchard. (*See* 1229.)

621.396.677.71 1259

The Slot Aerial—B. L. Morley. (*Wireless World*, vol. 61, pp. 129–131; March, 1955.) Slot antennas have been found effective for eliminating ghost-reception in television. Simple theory and design details are given. The cavity resonator and skeleton slot are mentioned as special cases.

621.396.677.71 1260

Radiation from a Slot on a Cylindrically Tipped Wedge—J. R. Wait and S. Kahana. (*Canad. Jour. Phys.*, vol. 32, pp. 714–721; November, 1954.) The system considered is an infinite wedge with the edge replaced by a portion of a concave cylinder with an arbitrary slot; it constitutes an idealized case of the winged-slot arrangement discussed by Alford (3034 of 1947). Radiation patterns are calculated for thin transverse $\lambda/2$ slots in the equatorial plane, for wedge angles of 90 degrees, 180 degrees and 270 degrees.

621.396.677.85:621.372.22:537.226 1261

Some Experiments on Artificial Dielectrics at Centimetre Wavelengths—M. M. Z. El-Kharadly. (*Proc. IEE*, Part B, vol. 102, pp. 17–25; January, 1955.) A parallel-plate transmission line is found to be a convenient system for investigating artificial dielectrics. For the wavelength range 8–11 cm, the width of the plates is about 60 cm and the spacing between them about 3 cm to ensure that only the TEM mode propagates. Details are given of the line construction. The method was shown to be accurate to within about 2 per cent by making measurements on ordinary dielectrics. Results of refractive-index and impedance measurements are presented for arrays of spheres, disks and rods. Dispersion is related to (a) resonance of the array elements and (b) spacing between elements.

621.396.677.85:621.372.22:537.226 1262
A General Experimental Method to determine the Properties of Artificial Media at Centimetre Wavelengths, applied to an Array of Parallel Metallic Plates—R. I. Primich. (*Proc. IEE*, Part B, vol. 102, pp. 26–36; January, 1955.) Measurements of refractive index and magnitude and phase of reflection coefficient were made on an artificial dielectric comprising a semi-infinite array of thin metal plates, using a parallel-plate transmission-line system as described by El-Kharadly (1261 above). Use of the parallel-plate system permits precise location of the interface between the artificial dielectric and free space. Satisfactory accuracy is obtained at wavelengths between 9 and 10 cm.

621.396.677.85 1263
Modified Luneberg Lens—A. S. Gutman. (*Jour. Appl. Phys.*, vol. 25, p. 1553; December, 1954.) Correction to paper abstracted in 3472 of 1954.

621.396.67.012.12 1264
Antennas Diagrams. [Book Review]—International Radio Consultative Committee (C.C.I.R.). Publishers: International Telecommunication Union, Geneva, 52 pp., S.Fr. 18.45. (*Wireless Eng.*, vol. 32, p. 90; March, 1955.) Contains formulas and power-distribution diagrams for vertical, horizontal dipole, curtain and rhombic antennas and antennas for tropical broadcasting.

AUTOMATIC COMPUTERS

681.142 1265
An Autocorrelogram Computer—G. Revesz. (*Jour. Sci. Instr.*, vol. 31, pp. 406–410; November, 1954.) A comparatively simple and inexpensive instrument is described, with detailed circuit diagram. The signal, recorded on a magnetic tape, is played back through two separate heads, thus producing the functions $f(t)$ and $f(t-\tau)$. These are amplified, multiplied and then integrated by a modified dc watt-hour meter. The computer was originally developed as a yarn cross-section analyzer.

681.142 1266
Frequency Analysis of Computer Systems—R. Boxer. (Proc. IRE, vol. 43, pp. 228–229; February, 1955.)

681.142 1267
Australian Guided Weapons Analogue Computer—(*Aust. Jour. Instr. Tech.*, vol. 10, pp. 145–146; November, 1954.) Brief description of AGWAC, whose design is based on that of the Royal Aircraft Establishment simulator TRIDAC. The equipment is built up of 280 plug-in units ("bricks").

681.142 1268
Differential Analyser: the N.P.L.'s New Analogue Computer for solving Differential Equations—[*Engineering* (*London*), vol. 178, pp. 659–660; November 19, 1954.] Brief description of a large analyzer, recently brought into use. The machine is actuated by a motor-driven shaft representing the independent variable. The three principal types of mechanism used are integrators, gear trains and differential gears. Function tables permit an arbitrary function of any variable to be inserted from the graph and also allow solutions to be obtained in graphical form. Setting-up time is low, since all units are coupled by means of servomechanisms which may be connected in any combination through a telephone-type switchboard.

681.142 1269
Algebraic Method of Synthesis of Multicontact Relay Systems—V. I. Shestakov. [*Compt. Rend. Acad. Sci.*, (*URSS*), vol. 99, pp. 987–990; December 21, 1954. In Russian.] A vector-algebraic method of synthesis of switching circuits suitable for computers is developed theoretically.

681.142:621.37 1270
Note on the Resistance-Network Analogue Solution of Field Problems of Spherical Symmetry—G. Liebmann. (*Brit. Jour. Appl. Phys.*, vol. 5, p. 412; November, 1954.) Results of a previous paper (945 of May) are applied to the derivation of design formulas for a resistance-chain analog for solving the equation div K grad $U=g$ in cases of spherical symmetry.

CIRCUITS AND CIRCUIT ELEMENTS

534.213–8:621.396.6 1271
Bibliography on Ultrasonic Delay Lines—M. D. Fagen. (*Trans. IRE*, no. PGUE-2, pp. 3–8; November, 1954.) 78 references are listed.

621.3.066.6:621.396.822 1272
Current Fluctuation Phenomena in Current-Carrying Sliding Contacts—G. W. Epprecht. (*Jour. Appl. Phys.*, vol. 25, pp. 1473–1480; December, 1954.) An account is given of an experimental investigation of sliding contacts as sources of interference voltages. The dependence of the noise characteristics on sliding speed and current intensity is examined. Most of the observed phenomena have a thermal origin. Several ranges of operation involving different mechanisms are distinguished.

621.314.263 1273
Odd-Integer Magnetic Frequency Multipliers—L. J. Johnson and S. E. Rauch. (Proc. IRE, vol. 43, pp. 168–173; February, 1955.) For various applications in the frequency range 400 cps–20 kc, frequency multiplication can be performed more efficiently by magnetic than by electronic equipment. The operation of the magnetic multiplier is analyzed and experimental units are described. For higher multiples, multistage arrangements are preferred. Harmonic distortion can be kept low by suitable choice of core materials.

621.318.5 1274
Component Design Trends—New Relay Materials improve Performance—F. Rockett. (*Electronics*, vol. 28, pp. 144–148; January, 1955.) Tungsten-carbide contacts, nickel-iron armatures, stainless-steel springs, silicone damping fluids and self-shielded cores are among the materials discussed.

621.318.57 1275
Flip-Flop Counter has Expanded Range—H. Beckwith. (*Electronics*, vol. 28, pp. 149–151; January, 1955.) Design procedure is indicated for counters having up to six stable states, without using feedback or matrixing. Long-term reliability is a primary objective. Crystal diodes are used in cross-coupling networks.

621.318.57:621.314.7 1276
Junction Transistor Switching Circuits—T. A. Prugh. (*Electronics*, vol. 28, pp. 168–171; January, 1955.) A 75-transistor experimental switching system includes blocking oscillator, multivibrator, flip-flop, gating and adding circuits.

621.318.57:621.314.7 1277
A Method of designing Transistor Trigger Circuits—F. C. Williams and G. B. B. Chaplin. (*Proc. IEE*, Part B, vol. 102, p. 74; January, 1955.) Discussion on 2914 of 1953.

621.319.4 1278
The Behaviour of Polystyrene Capacitors under Alternating Voltage—A. P. Butra and V. T. Renne. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1974–1982; November, 1954.) Experiments at 50 cps are reported. The main conclusions reached are: (a) heat treatment has no adverse effect on life; (b) for a given value of applied voltage the life is shortened if the voltage gradient is increased; (c) for a given ratio between applied voltage and initial ionization voltage the life is considerably shortened if ambient temperature is raised. Certain considerations regarding the impregnation of capacitors with mineral oil are discussed.

621.319.4:621.372.54 1279
Discoidal vs Tubular Feed-Through Capacitors—H. M. Schlicke. (Proc. IRE, vol. 43, pp. 174–178; February, 1955.) Tubular and disk-type ceramic feed-through capacitors for filtering purposes are compared in respect of their coupling impedance and its frequency dependence. The disk type is superior as regards hf performance and mechanical ruggedness. The necessary improved shielding methods are described.

621.37:621.314.7 1280
Interesting Applications of Transistor Technique—W. W. Diefenbach. [*Funk-Technik* (*Berlin*), vol. 9, pp. 621–622; November, 1954.) Circuit diagrams and photographs are presented of various applications, including an af amplifier molded inside a record-player pick-up arm.

621.37:621.314.7 1281
Practical Circuits using Transistors—H. Salow. (*Fernmeldetech. Z.*, vol. 7, pp. 581–588; November, 1954.) Diagrams and descriptions of operation are given for amplifiers, oscillators and switching circuits.

621.37.029.64.049 1282
Integrated Microwave Circuits—E. Jamieson. (*Electronic Eng.*, vol. 27, pp. 60–63; February, 1955.) Use of a milled-block construction for producing microwave assemblies is shown to offer advantages over more usual waveguide assemblies.

621.372 1283
A Definition of Passive Linear Networks in Terms of Time and Energy—G. Raisbeck. (*Jour. Appl. Phys.*, vol. 25, pp. 1510–1514; December, 1954.)

621.372.413 1284
Thermodynamic Consideration of Electromagnetic Cavity Resonators—C. H. Papas. (*Jour. Appl. Phys.*, vol. 25, pp. 1552–1553; December, 1954.) A simple method is presented for calculating the shift of resonance frequency produced by introducing a foreign body into the cavity.

621.372.413:621.385.029.6 1285
Tunable Cavity for X-Band Oscillators—N. A. Spencer. (*Electronics*, vol. 28, pp. 135–137; January, 1955.) A rectangular-waveguide tuner for external-cavity reflex klystrons is described.

621.372.5 1286
Equivalent Circuit for a Passive Nonreciprocal Network—H. A. Haus. (*Jour. Appl. Phys.*, vol. 25, pp. 1500–1502; December, 1954.) A nonreciprocal quadripole is considered, comprising an ideal amplifier and phase shifter. Compared with the gyrator representation, this method facilitates the linking of the network parameters with measurements.

621.372.5:621.318.134 1287
Behavior and Applications of Ferrites in the Microwave Region—A. G. Fox, S. E. Miller and M. T. Weiss. (*Bell Sys. Tech. Jour.*, vol. 34, pp. 5–103; January, 1955.) Both reciprocal and nonreciprocal devices comprising ferrite-loaded waveguides are studied, the nonreciprocal effects in the latter being based on birefringence, phase shift, field displacement, and coupling through apertures, as well as on rotation of polarization. A method called "point-field" analysis facilitates qualitative investigation. Definitions are given of the gyrator, the isolator and the circulator. Performance of some experimental devices is reported.

621.372.5.029.3.018.78 1288

An Investigation of the Nonlinear Distortion of a Quadripole using a Continuous Spectrum Signal—V. M. Vol'f. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2054–2063; November, 1954.) For measuring the nonlinear distortion introduced by an af transmission system, it is necessary to use a continuous spectrum of frequencies with a given shape of the envelope. Relations are derived from which it is possible to calculate the spectra at the output of certain nonlinear systems for input signals covering a continuous band of frequencies and having envelopes represented by an exponential function or a difference between two exponential functions. As compared with tests using one or two spot frequencies, the results so obtained are more in accordance with results of subjective listening tests.

621.372.56.029.64:621.387 1289

The Theory and Design of Gas-Discharge Micro-Wave Attenuators—E. M. Bradley and D. H. Pringle. (*Jour. Brit. IRE*, vol. 15, pp. 11–24; January, 1955.) "The theory of the interaction between a d.c. gas discharge and an r.f. field is considered and expressions developed for the complex conductivity of the discharge. The simple Lorentz theory is shown to give values of complex conductivity not very different from those derived with more detailed analyses. Problems connected with the gas discharge including that of noise production are considered. Practical microwave attenuators developed by the authors are described in detail. Two main classes of device are considered: (*a*) devices with resonant structures, and (*b*) distributed devices."

621.373:621.396.822:530.145 1290

Vacuum Fluctuation Noise and Dissipation—J. Weber. (*Phys. Rev.*, vol. 96, pp. 556–559; November 1, 1954.) Continuation of previous investigation (2603 of 1954) of the interaction of an electron beam with the field in an enclosed region. An expression is derived for the observable mean squared emf, and is used to calculate the fluctuations induced in the beam when interacting with a damped oscillator. As the damping becomes very large the fluctuations approach zero.

621.373.4:621.375.232.3 1291

Cathode-Follower-Coupled Phase-Shift Oscillator—H. J. Reich. (Proc. IRE, vol. 43, p. 229; February, 1955.) A note on the advantages of using cathode-follower amplifiers between the sections of the phase-shifting network.

621.373.422 1292

Circuits for producing High Negative Conductance—H. J. Reich. (Proc. IRE, vol. 43, p. 228; February, 1955.) Several arrangements using two triodes are illustrated; negative conductances of the order of 10 mhos can be produced. Stable oscillators can be made to operate with a supply voltage of 10V or less.

621.373.43:621.375.23 1293

The Regenerative Pulse Generator—C. C. Cutler. (Proc. IRE, vol. 43, pp. 140–148; February, 1955.) Pulses are circulated repeatedly in a feedback loop including an expander and a filter, being alternately shortened in the expander and lengthened in the filter, and giving an output at each traversal. A loop gain of unity is maintained by agc. For microwave pulses the nonlinear expander characteristic may be obtained by means of a crystal. Microwave pulses of duration down to 0.002μs and dc pulses down to 0.005μs have been produced. Possible applications include communications, radar and measurements, where coherent pulse trains are required.

621.373.44 1294

A Versatile Pulse Shaper—G. E. Kaufer. (*Electronic Eng.*, vol. 27, pp. 78–81; February, 1955.) A triggered miniature-thyratron circuit is described for deriving sharp pulses with uniform amplitude and duration from an input of nonuniform pulses. A table shows the output waveform for various circuit arrangements and component values.

621.374.43 1295

Frequency Multiplication by Regenerative Modulation—D. Makow. (*Canad. Jour. Technol.*, vol. 32, pp. 206–219; November, 1954.) A multiple-frequency generator is discussed in which operation depends on mixing a feedback and an input voltage. Theory is presented for a circuit designed to produce four multiples of the input frequency; experimental confirmation of the results is reported. Multiples up to the 40th have been obtained with good amplitudes; maximum amplitudes can be arranged to fall in desired frequency ranges. Even and odd multiples can be made available at separate output terminals. Operation is unaffected by variations of anode, filament or input voltage or by phase shifts due to aging of components in the feedback loop.

621.375.2:621.3.018.75 1296

Effect of Steepness of Rise and Fall of the Input Pulse on the Response of Pulse Amplifiers: Part 2—B. K. Bhattacharyya. (*Indian Jour. Phys.*, vol. 28, pp. 371–395; August, 1954.) The response of a shunt-compensated amplifier to ramp-function and triangular pulses is analyzed. Part 1: 2894 of 1954.

621.375.2.024 1297

A Stable and Sensitive D.C. Amplifier with High Input Resistance—S.O. Nielsen and T. Rosenberg. (*Jour. Sci. Instr.*, vol. 31, pp. 401–404; November, 1954.) "A line-operated d.c. amplifier for automatic regulation using a.c. amplification and a breaker modulator has been designed with the following characteristics: direct reading instrument; zero drift after 'warming-up' less than 30 μV peak-to-peak in 12 hours' test; sensitivity (short-time stability) usually 10 μV; input resistance $3\times10^{9}\Omega$; and damping about 1 sec with $3\times10^{8}\Omega$ source resistance."

621.375.2.029.3 1298

New Amplifier has Bridge-Circuit Output—D. J. Tomcik and A. M. Wiggins. (*Audio*, vol. 38, pp. 17–19; November, 1954.) The basic circuit is described and complete circuit diagrams are given of 20-w and 30-w af amplifiers. The advantages of the circuit include (a) absence of switching transients, (b) absence of dc current through primary of output transformer, and (c) low quiescent current.

621.375.232 1299

Differential-Amplifier Design—A. M. Andrew. (*Wireless Eng.*, vol. 32, pp. 73–79; March, 1955.) The possibility is discussed of providing good rejection of in-phase signals without using a balancing adjustment or specially selected components. To achieve the desired result by use of in-phase negative feedback, care must be taken that the feedback and input voltages are correctly combined. Two circuits are described illustrating the principles involved; both have the disadvantage of requiring floating batteries and isolated heater supplies.

621.375.234 1300

The Possibility of adjusting the Internal Impedance and varying the Gain in Amplifiers with Combined Current and Voltage Feedback—W. Benz. (*Fernmeldetech. Z.*, vol. 7, pp. 362–370; July, 1954.) The internal impedance can be reduced to and maintained at any desired level R_i, independent of the gain or the amount of feedback, by using a fixed amount of the total feedback to achieve the impedance reduction and apportioning the remainder so that $r/k=R_i$, where r and k are the current and voltage feedback factors of this remainder. Three commonly used circuits incorporating this feature are discussed.

621.375.3:621.317.3 1301

Magnetic Amplifiers and Weak Currents—R. Klein. (*Bull. Soc. Franç. Élect.*, vol. 4, pp. 649–674; November, 1954.) Particular attention is devoted to the use of magnetic amplifiers for measurement purposes.

621.375.4.029.3:621.314.7:621.396.822 1302

Investigations of Noise in Audio-Frequency Amplifiers using Junction Transistors—P. M. Bargellini and M. B. Herscher. (Proc. IRE, vol. 43, pp. 217–226; February, 1955.) Noise-factor measurements were made using various types of transistor in different circuit arrangements; the influence of input termination and operating point are discussed. Results are presented graphically. Three distinct sources are identified; shot noise and thermal noise are independent of frequency over the af band, semiconductor noise varies approximately as $1/f$. Noise properties of transistors and thermionic tubes are compared.

621.396.6 1303

Economic Aspects of Printed Foil Circuits—P. Eisler. [*Research (London)*, vol. 7, pp. 441–447; November, 1954.]

GENERAL PHYSICS

530.19 1304

Search for a General Definition of Energy and of the Parameters of Extension and Intensity—P. Renaud, M. Joly and D. G. Dervichian. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 1603–1605; December 8, 1954.] A scheme of relations between energy and extension and intensity is presented which leads to definitions generally valid for all forms of energy including those as yet unknown.

534.21+538.566]:537.228.1 1305

Conductivity and Viscosity Effects on Wave Propagation in Piezoelectric Crystals—J. J. Kyame. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 990–993; November, 1954.)

534.26+[535.42:538.566 1306

Diffraction of Pulses by a Circular Cylinder—Friedlander. (*See* 1231.)

535.37 1307

Electroluminescence with Nonsinusoidal Fields—S. Nudelman and F. Matossi. (*Jour. Electrochem. Soc.*, vol. 101, pp. 546–553; November, 1954.) Experiments using pulses of rectangular, sawtooth and exponential form were made to supplement the information obtained by applying sinusoidal voltages [e.g. 3439 of 1952 (Piper and Williams)]. The phosphor investigated was ZnS:Cu, Pb, giving green and blue luminescent bands which were examined separately. Measurements were made of field and frequency dependence and decay law with square pulses; only qualitative observations were made with the other waveforms. A theoretical interpretation of the results is advanced.

535.376 1308

Voltage Dependence of Electroluminescent Brightness—B. T. Howard, H. F. Ivey and W. Lehmann. (*Phys. Rev.*, vol. 96, pp. 799–800; November 1, 1954.) An equation of the form $L=aV\exp[-b/(V+V_0)]$ relating the brightness L to the voltage V is found to fit experimental data well, a, b and V_0 being constants.

537.21 1309

The Two-Dimensional Electric Field of a Curved-Sided Wall or Groove or an Infinite Plane—N. H. Langton and N. Davy. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 405–410; November, 1954.) Results of a theoretical investigation are tabulated and presented graphically. The calculation involves conformal transformations and elliptic integrals for complex moduli; values of the latter are tabulated.

537.226 1310

Some Properties of Dielectrics with High Permittivity—M. Gourceaux. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1777–1778; December 20, 1954.] Analysis shows that for a conductor surrounded by a dielectric of sufficiently high permittivity, the depth of penetration of hf current is dependent only on the constants of the dielectric. Formulas are derived, for solid and for hollow dielectric cylinders and spheres subjected to an alternating electric field, expressing the conditions under which the amplitude of oscillation of the bound electrons becomes infinite; the results can be stated in terms of characteristic radii for a given material and frequency.

537.311.1 1311

Theory of Electrical Conduction—K. Ljoljc. (*Z. Phys.*, vol. 139, pp. 388–401; November 26, 1954.) General theory of conduction in pure metals is developed without assuming spherical symmetry for the energy levels. This is applied to calculate the conductivity of the alkali metals.

537.311.1 1312

The Theory of Electrical Resistance of Good Metallic Conductors—C. A. Busse and F. Sauter. (*Z. Phys.*, vol. 139, pp. 440–447; November 26, 1954.) Reasonable values are obtained for the resistivity of the alkali metals (except Li) by calculation based on the propagation of compression waves in an ionic continuum.

537.311.62:621.365.5 1313

Electromagnetic Heating of Thin Layers of Conductors or Semiconductors; Generalizations relating to the Problem of the Determination of the Ideal Thickness—M. Gourceaux. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1607–1609; December 8, 1954.] Formulas are quoted for the thickness of a cylindrical and of a spherical metal shell to make optimum use of the heating effect of a hf magnetic field. If the metal is replaced by a material which is both a conductor and a dielectric, the corresponding expressions are quite different though the ratio between them is the same. Consideration is also given to the case of a cylindrical shell rotated rapidly in a uniform constant transverse magnetic field.

537.52 1314

Artificial Increase of Electrical Breakdown Strength of Air at Low Pressure in the Region of 300 Mc/s—G. W. E. Stark and G. L. Fougere. [*Nature* (*London*), vol. 174, p. 1066; December 4, 1954.] Measurements covering the pressure range 5–120 mm Hg, using gap widths of 1–4 mm and frequencies of 227 and 379 mc are reported. Under these conditions, the superposition of a moderate direct voltage on the alternating voltage can increase the uhf breakdown stress quite considerably. Typical results obtained with direct voltages up to 125 v are shown graphically; the effect is large enough to find practical application as an inhibitor of breakdown in electronic equipment for operation at high altitudes.

537.529:537.58 1315

Theory of Electric Breakdown in the High-Temperature Region and its Relation to Thermionic Emission—K. Lehovec. (*Phys. Rev.*, vol. 96, pp. 921–928; November 15, 1954.) Discussion is based on consideration of the distribution of electrons over the energy states, rather than the behavior of individual electrons.

537.533.8 1316

Energy Distribution of Secondary Electrons from Magnesium and Beryllium Alloys—V. N. Lepeshinskaya and V. M. Tumorin. (*Zh. Tekh. Fiz.*, vol. 24, pp. 1933–1941; November, 1954.) Experiments were carried out at a pressure of $2-4\times10^{-6}$ mm Hg with alloys CuMg, AgMg, AlMg and CuBe activated by heating either in a vacuum or in an atmosphere of oxygen. Curves showing the dependence of the secondary emission coefficient on the energy of primary electrons and of the secondary current on the voltage at the collector are plotted. The mechanism of secondary emission from complex surfaces is not considered to be similar to that of the Malter effect; the role of the accelerating field is probably reduced here to the compensation of local retarding fields above the inhomogeneous surface of the activated layer.

537.533.8 1317

The Secondary Electron Emission of Sodium and Zinc—J. Woods. (*Proc. Phys. Soc.*, vol. 67, pp. 843–844; November 1, 1954.) Experiments are described in which the metal to be examined was deposited on a copper target block in a tube with a guiding grid accurately aligned with a collector grid. The collector potential was high enough for saturation of secondary current. To introduce the sodium an electrolytic method was used. For sodium, the maximum secondary emission factor reached a peak value of 1.2 during the coating process and then fell to a constant value of 0.82, occurring at a target potential of 300 v. For zinc, the maximum value of 1.15 occurred at a target potential of 200 v.

537.56:538.63 1318

Non-Maxwellian Theory of Homogeneous Anisotropic Plasmas—R. Jancel and T. Kahan. (*Nuovo Cim.*, vol. 12, pp. 573–612; November 1, 1954. In French.) Using theory based on Boltzmann's integro-differential equation, a calculation is made of the distribution of electron velocities in an ionized gas subjected to a constant magnetic field, for non-Maxwellian conditions. Explicit expressions are derived for the magneto ionic conductivity, dielectric tensor, Hall effect, deflection of an electron beam, and a generalized form of Langevin's mobility formula. Results are compared with those obtained by other methods based on mean free path. Wave propagation is studied, formulas being derived for refractive index, birefringence, phase and group velocities, attenuation, polarization, and critical frequencies. Classical results of ionosphere theory are confirmed. See also 3196 of 1954 and back references.

537.56:538.65 1319

Magnetic Fields in Plasmas with Turbulent Motion—P. O. Schilling and W. Lochte-Holtgreven. (*Z. Naturf.*, vol. 9a, pp. 520–526; June, 1954.) When a plasma comprising electrons and ions is in turbulent motion, diffusion currents give rise to magnetic fields. These can be demonstrated by rotating the gases produced by burning a mixture of oxygen and propane. An account is given of suitable experimental technique.

538.221 1320

Spin Precession in a Vibrating Bloch Wall at Small Amplitudes—I. Lucas. (*Z. Naturf.*, vol. 9a, pp. 373–376; May, 1954.) The natural frequency of vibration of the wall is derived on the basis of an effective internal magnetic field. The formula obtained for a 180 degree wall is identical, except for a factor $\sqrt{2}$, with that given by Döring (*ibid.*, 1948, vol. 3a, p. 373) for a 90 degree wall, on the basis of the wall mass.

538.3 1321

Electromagnetism, Old and New—F. W. Warburton, P. Moon and D. E. Spencer. (*Jour. Frank. Inst.*, vol. 258, pp. 395–400; November, 1954.) Comment on 2923 and 2924 of 1954 and authors' reply.

538.3 1322

Energy of the Harmonic Oscillator in Vacuum—P. Braffort and C. Tzara. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1779–1780; December 20, 1954.] Continuation of work abstracted previously [3200 of 1954 (Brafford et al.)]. The "energy of the oscillator" is used to indicate the additional energy acquired by a free charge on passing into the oscillating state. The energy of the harmonic oscillator in the fluctuating field of the absorber is equal to the product of its natural frequency and a universal constant.

539.15 1323

Electronic Band Structure of Solids—C. A. Coulson. [*Nature* (*London*), vol. 174, pp. 949–950; November 20, 1954.] Brief report of conference held at the Radar Research Establishment at Malvern in September, 1954, at which criticisms of the energy-band theory were advanced.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.16 1324

The Isotropic Component of Cosmic Radio-Frequency Radiation—J. R. Shakeshaft. (*Phil. Mag.*, vol. 45, pp. 1136–1144; November, 1954.) Possible sources of the isotropic component deduced by Westerhout and Oort (419 of March) are discussed. The integrated radiation from normal extragalactic nebulas is shown to be inadequate to account for the observations, which may be explained on the basis of collisions between galaxies. The intensity to be expected depends strongly on the cosmological theory assumed, hence accurate measurements of the extragalactic component may yield information of value for distinguishing between different theories.

523.746 1325

Prediction of the Present Sunspot Cycle—A. G. McNish and J. V. Lincoln. (*Trans. Amer. Geophys. Union*, vol. 35, pp. 709–710; October, 1954.) The reliability of the prediction method presented previously (*ibid.*, 1949, vol. 30, pp. 673–685) is examined with reference to observed sunspot data for the period 1944–1954.

523.78 1326

Observation of the Total Eclipse of the Sun on 30th June 1954 at Hagaby (Sweden), and of the Partial Eclipse using the Radio Telescope at Meudon Observatory—M. Laffineur, B. Vauquois, P. Coupiac and W. N. Christiansen. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1589–1590; December 8, 1954.] The curve obtained from the observations in Sweden at 545 mc was less irregular than that observed at Khartoum in 1952; this is consistent with the lower level of solar activity. At totality, the residual radiation was 11 per cent, compared with 19.5 per cent in 1952.

550.3:551.5 1327

Symposium on Scientific Aspects of the International Geophysical Year 1957–1958—[*Proc. Nat. Acad. Sci.* (*Washington*), vol. 40, pp. 922–982; October, 1954.] The text is given of a number of papers, including the following:

"Morphology of Ionospheric Storms,"—H. G. Booker (pp. 931–943).

"The Morphology of the Aurora,"—A. B. Meinel (pp. 943–950).

"Diurnal and Seasonal Variation of the Airglow,"—F. E. Roach (pp. 950–956).

"Transport Problems in the Atmosphere,"—H. Wexler (pp. 956–966).

"Evidence for Winds in the Outer Atmosphere,"—F. L. Whipple (pp. 966–972).

"Solar Activity and Terrestrial Disturbances,"—D. H. Menzel (pp. 973–978).

"Geographic Basis for Antarctic Scientific Observations,"—P. A. Siple (pp. 978–982).

550.38 1328

Homogeneous Dynamos and Terrestrial Magnetism—E. Bullard and H. Gellman. (*Phil. Trans. A*, vol. 247, pp. 213–278; November 30, 1954.) Solutions of Maxwell's equations are obtained for a sphere of electrically conducting fluid in which there are specified velocities. Solutions exist when harmonics up to degree four are included, but for a greater number of harmonics the convergence of solutions has not been proved. Orders of magnitude are derived for a number of quantities connected with the dynamo theory of terrestrial magnetism.

550.385.21:551.510.535 1329

Possible Cause of Disturbance of Solar Diurnal Variation of Geomagnetic Field—B. E. Bryunelli. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 741–743; December 11, 1954. In Russian.] A mechanism is considered whereby currents in the ionosphere acquire a pattern such as to produce the observed S_D variations of the geomagnetic field.

550.388.8:551.510.53 1330

Infrared Radiation of Night Sky—V. I. Krasovski. (*Uspekhi Fiz. Nauk*, vol. 54, pp. 469–494; November, 1954.) A survey. Production of infrared radiation in the upper atmosphere is discussed and the estimates given by various authors for the heights at which radiation is produced are quoted. The heights coincide roughly with the *D* and *E* layers. 70 references, including about 30 to Russian literature.

551.510.5:523.72 1331

Seasonal Variation in the Absorption of Solar Radiation by Atmospheric Ozone at 9.6 Microns—A. Adel. (*Bull. Amer. Met. Soc.*, vol. 35, pp. 250–252; June, 1954.)

551.510.52:551.594.13 1332

Aircraft Investigation of the Large Ion Content and Conductivity of the Atmosphere and their Relation to Meteorological Factors—R. C. Sagalyn and G. A. Faucher. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 253–272; November, 1954.) Continuation of work described previously [1289 of 1952 (Callahan et al.)]. Simultaneous measurements of conductivity, large-ion concentration, temperature and humidity over the altitude range 700–15,000 feet are reported. The results indicate that over continental areas in fair weather there exists a layer adjacent to the ground, varying in height from 1,000 to 10,000 feet, in which the vertical distributions of large ions and conductivity are controlled primarily by atmospheric turbulence. At the upper boundary of this layer there is a transition region about 800 feet high in which the conductivity increases by a factor of 1.5–6 while the large-ion content decreases by a factor of 1.5–100. Above this region the large-ion content is reduced to a very low value and the conductivity is determined primarily by the intensity of cosmic radiation.

551.510.53 1333

Reviews of Modern Meteorology—11. The Physical State of the Upper Atmosphere—K. Weekes. (*Quart. Jour. Roy. Met. Soc.*, vol. 80, pp. 2–15; January, 1954.) The composition of the atmosphere at various levels is discussed, mainly as regards the constituents N_2 and O_2; this provides an indication of the value to be assigned to the mean molecular weight in the formula for converting scale height to temperature. Following the discussion of the properties of the static atmosphere, the evidence for movement in the high atmosphere is reviewed, and the complications to be expected in the upper levels of the ionosphere as a result of the interaction of the ionized gas with the earth's magnetic field are considered.

551.510.535 1334

Determination of Electron Densities in the Ionosphere from Experimental (h',f) Curves—H. A. Whale. (*Jour. Atmos. Terr. Phys.*, vol. 5, p. 351; November, 1954.) An error in the argument used in the original paper (2177 of 1951) was pointed out by Kelso (2088 of 1954); it is suggested that it may nevertheless be possible to provide a rigorous argument justifying the use of the curves given.

551.510.535 1335

Concentration of Hydroxyl in the Upper Atmosphere—V. I. Krasovski. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 979–981; December 21, 1954. In Russian.] A brief critical discussion of possible reactions and their equilibria at heights of about 100 km.

551.510.535 1336

The Physics of the Ionosphere—J. A. Ratcliffe. (*Proc. IEE*, Part I, vol. 101, pp. 339–346; November, 1954.) A brief survey of selected aspects of recent research on the ionosphere.

551.510.535 1337

The Anomalous Equatorial Belt in the F_2-Layer—E. V. Appleton. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 348–351; November, 1954.) An examination has been made of the relation between N_m (the maximum electron density) and geomagnetic latitude for hours of the equinox day other than noon. The abnormally low values of N_m found over a 4,000-km-wide equatorial belt at noon tend to disappear during the afternoon and to be replaced by abnormally high values during the evening; the anomaly disappears between midnight and 0300 h. These variations appear to be due to a contraction of the vertical extent of the F_2 layer during the first half of the night, and an expansion prior to noon.

551.510.535 1338

F_2-Layer Regularities at Ibadan—N. J. Skinner and R. W. Wright. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 290–297; November, 1954.) Analysis of records for the period December, 1951–January, 1953 confirms that the daily variation of n (the total electron content in a column of unit cross section below the height of maximum ionization in the F_2 layer) is relatively free from anomalies exhibited by the corresponding variation of N_m (the maximum electron density of the F_2 layer). In calculating n, the nonparabolic electron distribution at Ibadan must be taken into account. Inclusion of the F_1-layer electron contribution has no appreciable effect on the regularity of the daily variation. A new method is described for deducing the recombination coefficient and the rate of ion production from the variation of n. The phenomenon of sunset minimum of $h'F_2$ is discussed; reasons are given for considering it a spurious effect arising from errors in the reduction of records.

551.510.535:523.78 1339

Preliminary Results of Ionospheric Observations of the Solar Eclipse of 30th June 1954—K. Bibl and F. Delobeau. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1658–1660; December 8, 1954.] Ionosphere soundings were made in the zone of totality, from a ship off the coast of southern Norway. A panoramic instrument was used covering the frequency range 1–20 mc in 12 seconds; a second instrument was used for absorption measurements at 1.43, 2.09, 2.86 and 4.18 mc. The variation of the maximum electron concentrations for the different layers is shown graphically. The recombination coefficients are estimated (a) from the delay between the optically and the ionospherically observed maximum phase, and (b) by trying different values in the equilibrium equation expressing the dependence of electron concentration on unobscured fraction of the solar disk. α_2, the recombination coefficient for a normal law, is greater than or equal to 5×10^{-8} cm³ for the *E* and F_1 layers and over 5×10^{-9} cm³ for the F_2 layer.

551.510.535:621.396.11 1340

The Physics of the Ionosphere—[*Nature* (*London*), vol. 174, pp. 866–868; November 6, 1954.] Report of conference held at Cambridge, England, in September, 1954, at which papers were presented and discussed summarizing existing knowledge of the subject.

551.510.535:621.396.11 1341

Simultaneous Ionospheric Absorption Measurements at Widely Separated Stations—Beynon and Davies. (*See* 1469.)

551.510.535:621.396.822 1342

Ionospheric Thermal Radiation at Radio Frequencies: Part 2—Further Observations—F. F. Gardner. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 298–315; November, 1954.) Continuation of work reported previously [406 of 1952 (Pawsey et al.)]. Observations were made at frequencies near 2 mc, over a period of about a year, at a site chosen to minimize man-made noise. The location of the absorbing regions was deduced from consideration of pulse-echo observations of the *D* region made on 2.28 mc [132 of 1954 (Gardner and Pawsey)]. The ionosphere temperatures deduced ranged from about 200 degrees to 250 degrees K, varying appreciably from day to day and being lower and less variable in winter. An increase of up to 40 degrees during sudden ionospheric disturbances is probably due to a change of height of the 2-mc absorbing region. Polarization measurements in September, 1951 indicated that the absorbing regions were then located at a level where the temperature decreased with height.

551.510.535:621.396.822:523.16 1343

Effects of Solar Flares on the Absorption of 18.3-Mc/s Cosmic Noise—C. A. Shain and A. P. Mitra. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 316–328; November, 1954.) Abnormal variations of ionospheric absorption were investigated during the course of the observations reported previously (1426 of 1954). Most cases of unusually high absorption appear to be associated with solar flares. Observations of the increases in absorption during the resulting sudden ionospheric disturbances are compared with observations of sudden phase anomalies in waves reflected from the ionosphere; a 1-db absorption difference appears to correspond to a phase anomaly of about 200 degrees at the frequency considered. As a method of detecting sudden ionospheric disturbances, observation of the absorption variation is as sensitive as and more convenient than observation of phase anomalies. Evidence was found of a delayed increase of absorption about 30 hours after some solar flares; the magnetic records showed no abnormal features at these times, and it is uncertain whether the excess absorption occurs within the ionosphere or outside it.

551.593 1344

Theoretical Considerations regarding the Dayglow—D. R. Bates and A. Dalgarno. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 329–344; November, 1954.) Results of rocket observations made by Miley et al. (*Trans. Amer. Geophys. Union*, vol. 34, p. 680; 1953) are examined; the reported great altitude and luminosity of the emissive source cannot be reconciled with the assumption that dayglow is due to direct photo-action by incident solar radiation. A determination is made of the emission to be expected from processes known to be operative, including resonant and fluorescent scattering, photo-dissociation, photo-ionization and recombination.

551.594.21 1345

Chain-Process of Accumulating Charges in Thunderclouds—V. M. Muchnik. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 537–538; December 1, 1954. In Russian.] Theoretical considerations lead to the conclusion that the buildup of charge in a thundercloud follows an exponential law.

551.594.21 1346

Field Intensities and Charge Densities in Thunderclouds—H. Norinder and W. Pucher. (*Ark. Geofys.*, vol. 2, pp. 97–107; November 22, 1954.) Calculations based on the assumption that the thundercloud is ellipsoidal give the maximum field strength in the cloud as 1–5 kv/cm and the charge density as $10^{-5}-2.10^{-5}$ esu/cm³. The conditions under which lightning discharges may be initiated are discussed.

551.594.22:621.396.963.3 1347

Radar Echoes from Lightning—Jones. (*See* 1356.)

551.594.6:621.396.11 1348

The Wave-Forms of the Electric Field in Atmospherics recorded simultaneously by Two Distant Stations—H. Norinder. (*Ark. Geofys.*, vol. 2, part 2, pp. 161–195; November 22, 1954.) Electric-field variations were recorded at two stations, one near Uppsala, the other at Fotevik at the southwestern tip of Sweden, during September, 1949, and the variations were correlated with thunderstorms which occurred at distances between 240 and 2,400 km from Uppsala. In pre-discharge atmospherics a quasi-periodic variation was observed which corresponded, most frequently, to a wavelength of 11 km. The waveforms of the main discharges varied greatly with distance, often in a quasi-periodic manner. A unipolar form was also observed. The mean waveform, obtained from several hundreds of oscillograms, shows five pulses reflected from the ionosphere in addition to the ground pulse. Waveform changes with distance traveled can only be calculated by making large simplifications. Numerous oscillograms are reproduced and discussed.

LOCATION AND AIDS TO NAVIGATION

621.396.93 1349

H.F. Direction Finding—C. G. McCue. (*Wireless Eng.*, vol. 32, pp. 79–81; March, 1955.) "Some night-time d.f. observations were made at the Radio Research Organization's stations at Slough and Winkfield on h.f. radio signals from Sterling, Virginia, U.S.A., during February and March 1953. The purpose of these measurements was to test whether the observed bearings depended upon the direction of the transmitting aerial beam. No significant correlation of this nature was disclosed."

621.396.93(083.71) 1350

I.R.E. Standards on Radio Aids to Navigation: Definitions of Terms, 1954—(Proc. IRE, vol. 43, pp. 189–209; February, 1955.) Standard 54 IRE 12S1.

621.396.933 1351

Radio Navigation Methods and their Evaluation—E. Kramar. (*Fernmeldetech. Z.*, vol. 7, pp. 571–576; November, 1954.) A review of known systems of aircraft navigation aids, from the point of view of the operator.

621.396.933.4 1352

Ground-Controlled Approach Systems—A. Hunkeler. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 45, pp. 1018–1023; November 27, 1954.) An illustrated description of civilian and military uses of G.C.A.

621.396.96 1353

Application of Radar in National Defence—J. H. Leutwyler. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 45, pp. 1009–1018; November 27, 1954.) An illustrated account is presented of some ground and airborne radar equipment tested by the Swiss Army. Requirements for various units and for the coordination of units are outlined.

621.396.96:519.21 1354

On the Effect of Integration in a Pulsed Radar, Randomly Modulated Carrier—W. M. Stone. (*Jour. Appl. Phys.*, vol. 25, pp. 1543–1548; December, 1954.) Results obtained previously (3602 of 1953) on statistical theory of detection are extended to include the case of integration of N pulses before a decision is reached by the detector associated with a pulsed radar system. Sets of curves show the probability of detection as a function of average signal/noise power ratio.

621.396.96:621.396.621 1355

Design of a Logarithmic Receiver—S. Rozenstein. (*Proc. IEE*, Part B, vol. 102, pp. 69–74; January, 1955.) A receiver of the successive-detection type for pulsed radar is discussed. Details are given of an IF stage designed to maintain the screen-grid voltage constant. A video amplifier is described including a differentiating circuit for signals of large dynamic range with negligible overshoot. Conclusions are drawn for the practical application of this receiver with various indicators.

621.396.963.3:551.594.22 1356

Radar Echoes from Lightning—R. F. Jones. (*Quart. Jour. Roy. Met. Soc.*, vol. 80, pp. 579–582; October, 1954.) A-scope records are shown which were obtained in August, 1953 using a wavelength of 10 cm and a stationary antenna. A possible distribution of the charge in a thundercloud is deduced.

621.396.963.325 1357

Suppression of Fixed-Target Echoes in Radar: Application of the Doppler Effect—M. Wildi. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 45, pp. 1023–1047; November 27, 1954.) An outline of moving-target-indicator systems is given. Suppression is based on the electrical subtraction of two successive echoes; the effectiveness of this depends on the level of instabilities and these are estimated. Some measured values are also given.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 1358

A Thermionic Ionization Gauge of High Sensitivity employing a Magnetic Field—G. K. T. Conn and H. N. Daglish. (*Jour. Sci. Instr.*, vol. 31, pp. 412–415; November, 1954.) The gauge comprises a cylindrical anode, an axial filament and a pair of ion-collecting end-plates. In operation an axial magnetic field of the order of 150 oersted is used; the gas pressure is then determined by measurement of the ion current. The sensitivity is 800 μa per micron pressure at an emission corresponding to 1 ma anode electron current, with no field, and 7,000 μa/micron at 2 ma; this is of the order of 100 times the sensitivity of a conventional thermionic gauge.

531.788.7 1359

The Influence of the Ballast Resistance on the Performance of Penning Vacuum Gauges—G. K. T. Conn and H. N. Daglish. (*Jour. Sci. Instr.*, vol. 31, pp. 433–434; November, 1954.) A brief note.

531.788.7:621.376.3 1360

Recording Vacuum Gauge—D. T. Hurd and M. L. Corrin. (*Rev. Sci. Instr.*, vol. 25, pp. 1126–1128; November, 1954.) The gauge described is a modification of the Langmuir type; the frequency of an oscillator is arranged to depend on the varying capacitance between a stationary electrode and an oscillating fiber whose damping is a function of the pressure of the ambient gas.

533.5 1361

An Instrument for Leak Detection and Pressure Measurement in High Vacuum Systems—E. G. Leger. (*Canad. Jour. Tech.*, vol. 32, pp. 199–205; November, 1954.) The instrument described is suitable for measurements of gas pressure in the range $10^{-1}-10^{-7}$ mm Hg. Leaks are detected by spraying oxygen on the outside of the vacuum system and noting the decrease in electron emission from a tungsten filament in an ionization gauge when it is operated as a saturated diode. Indication is given either by a meter or by the variation of frequency of an af oscillator.

535.37 1362

Modifications of Electroluminescence under the Influence of Temperature—J. Mattler. [*Compt. Rend. Acad. Sci. (Paris)*, vol. 239, pp. 1616–1618; December 8, 1954.] Measurements were made on ZnS phosphors, variously activated, over the temperature range −166 degrees to +155 degrees C. An experimental arrangement was used which did not introduce temperature effects due to the supporting dielectric. The applied voltage ranged from 125 v to 900 v. In one group of phosphors the electroluminescence decreased monotonically over the temperature range, in another group the electroluminescence passed through a maximum value at room temperature. Variations of spectral distribution as well as intensity were observed. A theoretical interpretation of the results is outlined.

535.376 1363

Cathodo-Luminescence of MgO—B. D. Saksena and L. M. Pant. (*Proc. Phys. Soc.*, vol. 67, pp. 811–816; November 1, 1954.) The effects on the luminescence spectra of exposure to cathode rays and of heating are examined experimentally and the results are interpreted in terms of impurity-center processes.

535.376:546.47-31 1364

Luminescence Mechanism of Zinc Oxide Phosphors—H. Gobrecht, D. Hahn and K. Scheffler. (*Z. Phys.*, vol. 139, pp. 365–371; November 26, 1954.) Investigation shows that the decay of luminescence in the ultraviolet and green bands after electron-beam excitation and, down to −60 degrees C., in the green band after long-wave ultraviolet excitation is exponential. A thermoluminescence maximum occurs at 118 degrees K. The temperature characteristic of luminescence is independent of the excitation intensity. This is not compatible with a recombination mechanism suggested earlier [3442 of 1952 (Gobrecht et al.)] Similarity with the luminescence characteristics of barium platinocyanide is noted.

537.226:621.315.612.4 1365

Structural and Dielectric Studies in the System (Ba, Ca) (Ti, Zr)O_3—M. McQuarrie and F. W. Behnke. (*Jour. Amer. Ceram. Soc.*, vol. 37, pp. 539–543; November, 1954.) "The crystalline phases in the system (Ba, Ca) (Ti, Zr)O_3 were identified and their lattice dimensions were determined from X-ray powder patterns of fired ceramic disks. Compositions near the $BaTiO_3$-$BaZrO_3$ and $CaTiO_3$-$CaZrO_3$ sides of the composition square gave evidence of complete solid solution, whereas all intermediate compositions contained two separate phases. Variations in the temperature of the Curie point and in aging rate were found to correlate reasonably well with variations in the character of the crystal structure."

537.226:621.372.22 1366

The Relative Permittivity of Tetragonal Arrays of Perfectly Conducting Thin Discs—J. Brown and W. Jackson. (*Proc. IEE*, Part B, vol. 102, pp. 37–42; January, 1955.) "The methods available for the calculation of the interaction field of an array of dipoles are discussed with reference to an artificial dielectric formed from a tetragonal array of conducting discs. A modified method is described and used to derive a formula for the relative permittivity, valid when the discs are not too closely spaced. For closely packed discs, the fields due to higher-order multipoles become important and a second formula for the relative permittivity is obtained for this case. Comparison with experimental values already obtained shows that the two formulae together enable the calculation of the relative permittivity over the whole range of spacings likely to be used in practice to be made."

537.226:621.372.22 1367

The Properties of Artificial Dielectrics at Centimetre Wavelengths—J. Brown and W. Jackson. (*Proc. IEE*, Part B, vol. 102, pp. 11–16; January, 1955.) The properties of the artificial dielectric are compared with those of a solid dielectric, and possible definitions of refractive index and wave impedance are examined. The latter is most appropriately defined in terms of the reflection coefficient at an interface between the dielectric and free space. The reactive fields set up near such an interface are examined qualitatively, and ways of representing them are discussed. Possible applications of artificial dielectrics other than as antennas are mentioned.

537.226:621.396.677.85:621.372.22 1368

Some Experiments on Artificial Dielectrics at Centimetre Wavelengths—El-Kharadly. (*See* 1261.)

537.226:621.396.677.85:621.372.22 1369

A General Experimental Method to determine the Properties of Artificial Media at Centimetre Wavelengths, applied to an Array of Parallel Metallic Plates—Primich. (*See* 1262.)

537.227+621.315.612 1370

Dielectric Properties and Phase Transitions of $NaNbO_3$ and (Na, K)NbO_3—G. Shirane, R. Newnham and R. Pepinsky. (*Phys. Rev.*, vol. 96, pp. 581–588; November 1, 1954.) Optical, X-ray and dielectric measurements on single-crystal and ceramic specimens of $NaNbO_3$ and solid solutions of (Na, K)NbO_3 are reported. No evidence was found of ferroelectricity in $NaNbO_3$; a small addition of $KNbO_3$ produced a ferroelectric phase. The phase transitions in $NaNbO_3$ are quite different from those in $KNbO_3$.

537.227 1371

Ferroelectric Properties of Single Crystals—C. F. Oxbrow. [*Nature* (*London*), vol. 174, pp. 1091–1093; December 11, 1954.] Report of a colloquium held at Christchurch, England, in September, 1954.

537.227:[546.431.824-31+547.476.3 1372

Ferroelectric Ceramics with Very Pronounced Nonlinear Properties—T. N. Verbitskaya. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 100, pp. 29–32; January 1, 1955. In Russian.] A comparison is made of the variation of the dielectric constant with applied electric field for Rochelle salt, $BaTiO_3$, T-7500, and Varikond VK-1. The last-mentioned is similar to Rochelle salt in respect of the large change of the dielectric constant in fields up to 1 kv/cm. Experimental results are presented graphically.

537.227:546.431.824-31 1373

Infrared Spectrum of Barium Titanate—R. T. Mara, G.B.B.M. Sutherland and H. V. Tyrell. (*Phys. Rev.*, vol. 96, pp. 801–802; November 1, 1954.) Observations are reported and are discussed in the light of theories about the ferroelectric character of $BaTiO_3$.

537.227:546.431.824-31:621.396.822 1374

Noise Generation in Crystals and in Ceramic Forms of Barium Titanate when subjected to Electric Stress—A. C. Kibblewhite. (*Proc. IEE*, Part B, vol. 102, pp. 59–68; January, 1955.) Variation of the noise is investigated experimentally as a function of the electric stress on the material and of its temperature relative to the Curie point. The existence of a Barkhausen effect associated with the ferroelectric domain structure is established. From the results obtained on crystals an estimate is made of the effective volume of the individual domains. In ceramics additional noise is observed, the origin of which is not yet certain.

537.228.1 1375

Piezoelectrically Excited Vibrations of Plates and Rods, and Dynamic Determination of Elastic and Piezoelectric Constants—R. Bechmann. (*Arch. elekt. Übertragung*, vol. 8, pp. 481–490; November, 1954.) Two methods of determining the piezoelectric constants are discussed, (a) from equivalent-circuit dynamic-capacitance considerations, and (b) from measurements of resonance and antiresonance frequencies. Method (a) is independent of the dielectric properties of the material; method (b) depends on the parallel capacitance and hence on the dielectric properties, and permits evaluation of an equivalent dielectric constant whose value depends on the mode of vibration. The theoretical results are supported by measurements on ADP and $BaTiO_3$ ceramic plates.

537.228.1 1376

Piezoelectric Effect in Minerals—M. P. Volarovich and E. I. Parkhomenko. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 239–242; November 11, 1954. In Russian.] A mainly qualitative experimental investigation of piezoelectricity in natural minerals is reported. Oscillograms are shown. The largest effect, observed in granite, is of the order of several hundredths of the effect in pure quartz. Smaller effects were observed in sandstone and quartzite; no effect in jasper, sienite, marble, dolomite, and limestone.

537.311.31:539.23 1377

Conductivity of Films in Longitudinal Magnetic Field—M. Ya. Azbel'. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 519–522; December 1, 1954. In Russian.] A theoretical treatment of electronic conduction in metallic films, for the case of parallel electric and magnetic fields.

537.311.33 1378

Electrical Conduction in Non-metallic Solids—C. W. Oatley. (*Proc. IEE*, Part B, vol. 102, pp. 7–10; January, 1955.) Conduction by electrons in nearly perfect crystals is considered, a gravitational analog being used to explain the motion of the electrons in the electric field. Production of free carriers by heating and by radiation is discussed.

537.311.33 1379

A Unidimensional Model of Semiconductor Compounds of Type $A^{III}B^{V}$—B. Seraphin. (*Z. Naturf.*, vol. 9a, pp. 450–456; May, 1954.) The main differences between semiconductor compounds of this type and semiconductors of Group-IV type are increased energy gap and electron mobility, lower hole mobility, and increased bond strength. The one-dimensional representation of the potential variation through the lattice shows how these differences result from the presence of alternate shallow and deep troughs in place of equal-depth troughs.

537.311.33 1380

Measurements on some Semiconducting Compounds with the Zinc-Blende Structure—F. A. Cunnell, J. T. Edmond and J. L. Richards. (*Proc. Phys. Soc.*, vol. 67, pp. 848–849; November 1, 1954.) The infrared transmission was measured for polished specimens of AlSb, GaSb, InSb, InAs, InP and GaAs; the curves are shown and deduced values of the optical activation energy are tabulated. Hall constant and conductivity were measured for GaSb, InSb and GaAs, and the electron concentrations and mobilities were determined; these figures are included in the table.

537.311.33 1381

Stability of Vertical Fusion Zones—W. Heywang and G. Ziegler. (*Z. Naturf.*, vol. 9a, pp. 561–562; June, 1954.) Recrystallization of semiconductors from the molten lower end of a vertically held rod is considered. A brief analysis indicates a maximum height of the molten zone for stability at a given radius of the solid rod.

537.311.33:535.215.3:538.639 1382

Theory of the Photomagnetic Effect in Semiconductors in High-Intensity Magnetic Fields—A. I. Ansel'm. (*Zh. Tekh. Fiz.*, vol. 24, pp. 2064–2069; November, 1954.) A brief review of theoretical papers on the subject is given, and a differential theory of the effect is developed, i.e. the value of the electric field along one of the axes is determined for a given gradient of the concentration of current carriers along this axis and a given value of the magnetic field along the other axis. Differences in the results obtained for the ionic and atomic types of semiconductor are indicated and interpreted.

537.311.33:537.213 1383

The Potentials of Infinite Systems of Sources and Numerical Solutions of Problems in Semiconductor Engineering—A. Uhlir, Jr. (*Bell. Sys. Tech. Jour.*, vol. 34, pp. 105–128; January, 1955.) "Tables and charts are given of mathematical functions related to the potential of a line of point charges. The use of these functions is illustrated by applications to semiconductor resistivity measurements and to calculations of the base resistance of point-contact transistors."

537.311.33:537.322.1:621.56 1384

The Use of Semiconductors in Thermoelectric Refrigeration—Goldsmid and Douglas. (*See* 1465.)

537.311.33:538.63:535.215.9 1385

Origin of E.M.F. on Illumination of Semiconductors in Inhomogeneous Magnetic Field—I. K. Kikoin, I. Kh. Ganev and A. I. Karchevski. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, p. 51; November 1, 1954. In Russian.] Further investigation of the new photomagnetic effect reported by Kikoin (3580 of 1954) shows that the observed emf is caused by a temperature gradient in the specimen. The direction of this emf depends on the direction of the magnetic field and on the magnetic nature (paramagnetic or diamagnetic) of the surrounding gas.

537.311.33:539.23 1386

Titanium Nitride and Titanium Carbide as Semiconductors—A. Münster, K. Sagel and G. Schlamp. [*Nature* (*London*), vol. 174, pp. 1154–1155; December 18, 1954.] Experiments are described with thin films of TiN deposited on Al_2O_3 and on SiO_2. The former exhibited metallic and the latter semiconductor properties. Examination of the structure of the films by X-ray diffraction indicated no difference between the two. Essentially the same behavior was observed with TiC and with mixed crystals of TiN and TiC.

537.311.33:539.23 1387

Thin Semiconductor Films on Glass—A. Fischer. (*Z. Naturf.*, vol. 9a, pp. 508–511; June, 1954.) Oxide films having useful conducting properties combined with good transparency are obtained by depositing suitable metallic compounds on heated glass surfaces; Zn and In give particularly good results.

537.311.33:[546.28+546.289 1388

Aging of Crushed Silicon and Cermanium Crystals—S. B. Brody. [*Acta Cryst.* (*Cambridge*), vol. 7, part 11, pp. 772–774; November

10, 1954.] Si and Ge have been investigated for the occurrence of time-dependent changes in the X-ray diagrams of freshly crushed crystals; Ge does not show these changes, Si does.

537.311.33:[546.28+546.289 1389
Elastoresistance in *p*-Type Ge and Si—E. N. Adams. (*Phys. Rev.*, vol. 96, pp. 803–804; November 1, 1954.) Formulas are derived for the energy states in the strained lattice and values are hence found for the three elastoresistance coefficients, referred to as piezoresistance coefficients by Smith (2418 of 1954).

537.311.33:546.28 1390
Hyperfine Splitting of Donor States in Silicon—J. M. Luttinger and W. Kohn. (*Phys. Rev.*, vol. 96, pp. 802–803; November 1, 1954.) A theoretical estimate of the hyperfine splitting of phosphorus donor states is derived giving reasonable agreement with experimentally observed values [453 of March (Fletcher et al.)].

537.311.33:546.289 1391
Properties of Germanium Doped with Iron: Part 1—Electrical Conductivity—W. W. Tyler and H. H. Woodbury. (*Phys. Rev.*, vol. 96, pp. 874–882; November 15, 1954.) "Measurements of the temperature dependence of electrical resistivity in *p*- and *n*-type iron-doped germanium crystals indicate that Fe introduces impurity levels in Ge at 0.34 ± 0.02 eV from the valence band and $0.27 + 0.02$ eV from the conduction band. Such samples show very high resistivity at 77 degrees K. At this temperature, *n*-type samples also show high photosensitivity and slow photoresponse, presumably because of hole traps."

537.311.33:546.289 1392
Properties of Germanium Doped with Iron: Part 2—Photoconductivity—R. Newman and W. W. Tyler. (*Phys. Rev.*, vol. 96, pp. 882–886; November 15, 1954.) Measurements were made on *n*-type and *p*-type specimens at a temperature of 77 degrees K. The spectral distribution of the photoconductive response due to the im purity is the same in both types, indicating ionization energies of about 0.3 ev. For a given incident power, the intrinsic photoconductive response was 10–10^4 times greater for *n*-type than for *p*-type specimens; part of this difference is probably due to a difference of time constants, while the remainder may be due to differences in the numbers of active centers and in their transition probabilities. Quenching phenomena are discussed in relation to trapping effects.

537.311.33:546.289 1393
Germanium Technology: Part 1—Preparation of the Germanium—J. M. Mercier. (*Bull. Soc. Franç. Élect.*, vol. 4, pp. 629–644; November, 1954.) Physical and chemical purification methods and the production of single crystals are described. See also 168 of February.

537.311.33:546.289 1394
Germanium Technology: Part 2—Measurement of the Semiconductor Properties of Germanium—B. Pistoulet. (*Bull. Soc. Franç. Élect.*, vol. 4, pp. 644–648; November, 1954.) A survey of methods for measuring resistivity and Hall coefficient. Part 1: 1393 above.

537.311.33:546.289 1395
Statistics of the Occupation of Dislocation Acceptor Centres—W. T. Read, Jr. (*Phil. Mag.*, vol. 45, pp. 1119–1128; November, 1954.) Continuation of work on dislocations in Ge reported previously (457 of March). Further approximations are developed taking account of nonuniform spacing of accepted electrons along the dislocation. The case where the dislocation acceptors have a range of energy levels rather than a single level is discussed. The various approximations are compared by plotting the fractions of filled acceptors against temperature for a particular dislocation in 1.7-Ω. cm *n*-type Ge.

537.311.33:546.289 1396
Influence of Free Charge Carriers in Germanium on the Absorption in the Near Infrared Region—J. Jaumann and R. Kessler. (*Z. Naturf.*, vol. 9a, p. 476; May, 1954.) Measurements were made on *p*- and *n*-type specimens over the temperature range corresponding to intrinsic conduction, up to 300 degrees C.; large reversible variations of the carrier concentration and optical absorption are obtained on varying the temperature. The absorption spectrum is structure free and indicates a value of 0.752 ev for the activation energy, in agreement with the value found from measurements of conductivity and Hall constant. Anomalously high absorption in the asymptotic part of the curve must be due to impurities, and affords a method of measuring impurities much more sensitive than electrical methods.

537.311.33:546.289 1397
The Effect of Special Lattice Imperfections on the Electrical Properties of Germanium—K. Blank, D. Geist and K. Seiler. (*Z. Naturf.*, vol. 9a, pp. 515–520; June, 1954.) An experimental investigation was made of the dependence of conductivity and carrier lifetime on imperfections such as individual misplaced atoms, dislocations and impurity centers. After 15 minutes heating at 800 degrees C. and cooling at the rate of 400 degrees/s, a sample which was initially an intrinsic semiconductor at room temperature contained fewer than 10^{14} ionized acceptors or donors per cm^3 when impurities were excluded. A similar result was obtained with cooling at 40 degrees/s combined with plastic deformation. Carrier lifetime was unaffected by simple heating, but was considerably reduced when plastic deformation was superimposed. With doped Ge similarly heated and cooled at 40 degrees/s, fewer than 10^{15} acceptors and some 10^{14} donors per cm^3 were produced which were not ionized at room temperature.

537.311.33:546.289 1398
On the Relation between the Sum of Donor and Acceptor Concentration and Lifetime in Single Crystal Germanium—P. Ransom and F. W. G. Rose. (*Proc. Phys. Soc.*, vol. 67, pp. 646–650; August 1, 1954.) An approximate relation between the total impurity concentration, the activation energy of the traps, and the lifetime of minority carriers in *n*-type single-crystal Ge is derived. Lifetime measurements were made on samples of Sb-doped *n*-type Ge in the temperature range 242 degrees K–302 degrees K, and on undoped samples in the range 242 degrees K–319 degrees K. Values of total impurity concentration are computed from resistivity measurements at various temperatures, and computations of other quantities are made from the experimental data. These show that the approximations made are valid over a considerable range of temperatures. It therefore seems possible from resistivity measurements at various temperatures to determine directly the trap concentration and, in conjunction with lifetime measurements at room temperature, to differentiate between samples having different traps.

537.311.33:546.289:541.135 1399
Experiments on the Interface between Germanium and an Electrolyte—W. H. Brattain and C. G. B. Garrett. (*Bell. Sys. Tech. Jour.* vol. 34, pp. 129–176; January, 1955.) A report is presented of studies of Ge electrodes in contact with aqueous solutions of KOH, KCl and HCl, with anodic and cathodic currents. The anodic current, apart from a small leakage term, is determined by the flow of holes to the surface, so that saturation occurs for *n*-type but not for *p*-type material; the current gain is 1.4–1.8. The cathodic current is determined by the supply of electrons; the current gain is of the order of unity and the leakage term is larger than for anodic current. The effects of illumination are discussed. The results can be understood from simple thermodynamic considerations. See also 1698 of 1953 (Brattain and Bardeen) and 458 of March (Garrett and Brattain).

537.311.33:546.289:548.7 1400
Etching of Single Crystal Germanium Spheres—R. C. Ellis, Jr. (*Jour. Appl. Phys.*, vol. 25, pp. 1497–1499; December, 1954.) As a result of etching, the spheres were transformed into solid figures bounded by characteristic polyhedra. The only twins found were of (111)-plane type.

537.311.33:546.289:621.314.632 1401
Investigation of Change of Rectifying Properties of Metal-Semiconductor Point Contact under the Influence of Bombardment of Semiconductor by Alkali Metal Ions—M. M. Bredov, R. F. Komarova and A. R. Regel'. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 69–72; November 1, 1954. In Russian.] An experimental investigation is reported of the effect on an *n*-type-Ge/metal contact of bombardment by a monochromatic beam of K ions with energy in the range 1–4 kev and of the subsequent treatment of the surface. The observed decrease of the rectification property depends on the energy of the ion beam, but no effect was noticed when this was less than 1 kev. The treatment included polishing and immersion in a 30 per cent solution of H_2O_2. Results are presented graphically and discussed.

537.311.33:546.472.21 1402
On the Properties of Single Cubic Zinc Sulfide Crystals—S. J. Czyzak, D. C. Reynolds, R. C. Allen and C. C. Reynolds. (*Jour. Opt. Soc. Amer.*, vol. 44, pp. 864–867; November, 1954.) Optical and electrical properties were measured for five crystals with impurity content <0.005 per cent. Conductivity is largely intrinsic, and there is a rapid rise and decay of the small photocurrents obtained.

537.311.33:546.482.21:548.5 1403
Microscope Observations and Conductivity Measurements on CdS Crystals—L. Herforth and J. Krumbiegel. (*Z. Naturf.*, vol. 9a, pp. 432–434; May, 1954.)

537.311.33:546.482.21:548.5 1404
The Position of the Twinning Plane in Hexagonal CdS Crystals—K. H. Jost. (*Z. Naturf.*, vol. 9a, pp. 435–436a; May, 1954.)

537.311.33:546.621.86 1405
Electrical and Optical Properties of Intermetallic Compounds: Part 3—Aluminum Antimonide—R. F. Blunt, H. P. R. Frederikse, J. H. Becker and W. R. Hosler. (*Phys. Rev.*, vol. 96, pp. 578–580; November 1, 1954.) "Measurements of resistivity, Hall coefficient, and optical absorption of AlSb are reported. The width of the forbidden energy band, derived from electrical and optical data, is 1.6 ev at absolute zero. Absorption bands are observed at 0.75 ev in *p*-type samples and at 0.31 ev in tellurium-doped *n*-type specimens. An energy level diagram with several acceptor and donor levels in the forbidden energy band is suggested to explain these observations."

537.311.33:546.681.86 1406
Electrical and Optical Properties of Intermetallic Compounds: Part 2—Gallium Antimonide—R. F. Blunt, W. R. Hosler and H. P. R. Frederikse. (*Phys. Rev.*, vol. 96, pp. 576–577; November 1, 1954.) "The Hall effect and resistivity of GaSb have been investigated over the temperature range 78 degrees to 750 degrees K. Hole mobilities as high as 400 cm^2/volt-sec and a forbidden energy gap of 0.775 ev at absolute zero were found. Optical absorption studies at temperatures between 10 degrees and 300 degrees K confirm this value of the band separation."

537.311.33:546.682.86 1407
Electrical and Optical Properties of Intermetallic Compounds: Part 1—Indium Antimonide—R. G. Breckenbridge, R. F. Blunt, W. R. Hosler, H. P. R. Frederikse, J. H. Becker and W. Oshinsky. (*Phys. Rev.*, vol. 96, pp. 571–575; November 1, 1954.) Measurements of conductivity and Hall coefficient over the temperature range 78 degrees–750 degrees K are reported. Values deduced for charge-carrier mobility and effective mass are in good agreement with results of other workers. The value 0.23 ev is found for the width of the energy gap at 0 degrees K. A study of the optical absorption at temperatures between 13 degrees and 300 degrees K shows that the position of the absorption edge of degenerate *n*-type samples depends on the impurity content and is in good agreement with Burstein's predictions (2134 of 1954).

537.311.33:546.682.86 1408
Electrical Properties of Indium Antimonide: Part 2—O. Madelung and H. Weiss. (*Z. Naturf.*, vol. 9a, pp. 527–534; June, 1954.) Careful measurements of Hall constant and conductivity are reported. The value of the energy gap at absolute temperature T is found to be $0.27–3.10^{-4}T$ ev. The electron mobility is about 65,000 $(T/300)^{-1.66}$ cm per v/cm. The hole mobility is considerably lower and is strongly dependent on temperature. For previous work see 2707 of 1954 (Weiss).

537.311.33:546.72 1409
Some Electrical Properties of Natural Crystals of Iron Pyrite—J. C. Marinace. (*Phys. Rev.*, vol. 96, p. 593; November 1, 1954.) Measurements of Hall coefficient and resistivity were made on *p*-type and *n*-type specimens cut from natural single crystals. A value about 1.2 ev is estimated for the width of the energy gap.

537.311.33:546.817.221 1410
Electrical Properties and the Solid-Vapor Equilibrium of Lead Sulfide—R. F. Brebrick and W. W. Scanlon. (*Phys. Rev.*, vol. 96, pp. 598–602; November 1, 1954.) "Natural PbS crystals were exposed to various pressures of sulfur-vapor near 500 degrees C and then quenched. Calculations based on the penetration of *p-n* junctions gave an interdiffusion constant of 2×10^{-6} cm²/sec at 550 degrees C. The temperature dependence of the Hall coefficient and resistivity for several treated crystals was determined. A forbidden energy gap of 0.37 ± 0.01 ev and an electron to hole mobility ratio of 1.4 was obtained."

538.221 1411
Reversible Susceptibility of Ferromagnetics—D. M. Grimes and D. W. Martin. (*Phys. Rev.*, vol. 96, pp. 889–896; November 15, 1954.) Formulas are developed for both transverse and parallel reversible susceptibilities. Measurements on ferrite cores are reported. The theoretical and experimental results show satisfactory agreement.

538.221 1412
Ferromagnetic Properties of Oxidized Mn_2Sb—G. D. Adam and K. J. Standley. (*Proc. Phys. Soc.*, vol. 67, pp. 1022–1023; November 1, 1954.) The formula Mn_2SbO is tentatively assigned to a material which has been prepared by heating finely powdered Mn_2Sb in air to temperatures between 200 degrees and 350 degrees C. Measurements of magnetization, Curie point, and microwave resonance absorption, as well as X-ray powder photographs, indicate that the material is quite distinct from Mn_2Sb.

538.221 1413
Magnetic Anisotropy Constants of Ferromagnetic Spinels—R. S. Weisz. (*Phys. Rev.*, vol. 96, pp. 800–801; November 1, 1954.)

538.221 1414
Ferromagnetic Resonance in Iron-Nickel Alloys—R. Hoskins and G. Wiener. (*Phys. Rev.*, vol. 96, pp. 1153–1154; November 15, 1954.) Experiments were made with disk-shaped bulk specimens to verify the results obtained by Bagguley (443 of 1954) with suspensions of ferromagnetic particles. No evidence was found of a variation of g value with Ni concentration.

538.221 1415
Measurement of the Complex Tensor Permeability of Ferrites—J. H. Rowen and W. von Aulock. (*Phys. Rev.*, vol. 96, pp. 1151–1153; November 15, 1954.) Disadvantages of the cavity-resonator method described by Artman and Tannenwald (769 of 1954) are removed by using a disk in place of a spherical sample.

538.221:538.652 1416
Rapid Method of Evaluating Magnetostrictive Materials for Electromechanical Transducers—M. T. Pigott and P. M. Kendig. (*Jour. Acoust. Soc. Amer.*, vol. 26, pp. 974–976; November, 1954.) Methods described by van der Burgt (3624 of 1953) are adapted for investigating metal laminations as well as ferrites.

538.221:546.73 1417
Magnetic Powder-Patterns on Cobalt and the Structure of the Elementary Domains—W. Andrä. [*Ann. Phys. (Leipzig)*, vol. 15, pp. 135–140; November, 1954.] Powder patterns, photographs of which are shown, are interpreted to give an indication of the structure in depth of the magnetic domains.

538.221:621.318.134 1418
Investigation of Magnetic Properties of Co-Zn and Mn-Zn Ferrites—G. I. Dmitrakova. [*Compt. Rend. Acad. Sci. (URSS)*, vol. 99, pp. 45–46; November 1, 1954. In Russian.] Graphs are presented showing the variation of magnetic moment and Curie point of the ferrites with the ZnO content, the temperature dependence of magnetization and the reciprocal susceptibility. The results are briefly discussed on the basis of Néel's theory (1171 of 1950).

621.3 1419
Materials used in Radio and Electronic Engineering—(*Jour. Brit. IRE*, vol. 15, pp. 47–64; January, 1955.) This first part of a survey of the materials most commonly used in electronic equipment deals with (a) aluminium and its alloys, and (b) piezoelectric crystals.

621.315.612.4 1420
Low-Temperature Sintering of Titanium Dioxide for Ceramic Capacitors—T. V. Ramamurti, C. V. Ganapathy and S. Saran. [*Nature (London)*, vol. 174, p. 1187; December 25, 1954.] Experiments made at the National Physical Laboratory of India have resulted in the production of ceramic bodies with a typical permittivity value of 110 when fired at temperatures as low as 1,100 degrees C. The raw material was of rutile form, in a finely divided active state.

621.315.616 1421
Transition Temperature of Polymethyl Methacrylate at Ultrasonic Frequencies—P. Hatfield. [*Nature (London)*, vol. 174, pp. 1186–1187; December 25, 1954.] Results of measurements by various workers of the temperature variation of the velocity of ultrasonic waves in perspex are shown graphically and discussed. Ultrasonic frequencies up to 11 mc were used.

MATHEMATICS

517 1422
Symmetric Linear Transformations and Complex Quadratic Forms—C. L. Dolph, J. E. McLaughlin and I. Marx. (*Commun. Pure Appl. Math.*, vol. 7, pp. 621–632; November, 1954.) Discussion of application of the Rayleigh-Ritz method as an approximation procedure for investigating characteristic values and characteristic functions occurring in the theory of wave propagation.

517.562 1423
Locus Diagrams of some Elementary Transcendental Functions—H. Wahl. (*Frequenz*, vol. 8, pp. 346–351; November, 1954.) The functions considered are e^{pA} and sin pA, where p is a real parameter and A is a constant which may be complex. The form of the curve, distribution of the parameters, special points on the curve, tangent and radius vectors, and the construction of the curves are briefly discussed.

MEASUREMENTS AND TEST GEAR

621.3.018.41(083.74)+621.396.91 1424
C.C.I.R. Study Commission No. VII: Organization of World-Wide Standard-Frequency and Time-Signal Service as Task of C.C.I.R.—U. Mohr. (*Fernmeldetech. Z.*, vol. 7, pp. 596–598; November, 1954.) Important points from the report presented at the 1953 plenary session of the CCIR are summarized. It is suggested that reports in preparation regarding the derivations and corrections of standard-frequency transmissions should be directly exchanged between interested stations. Possible improvements in the nature of time signals are indicated. Plans for a service within Western Germany are outlined; this would operate from a Post Office transmitter at Mainflingen (call sign DCF77) on a frequency of 77.5 kc constant to within 1 part in 10^8.

621.3.018.41(083.74) 1425
World-Wide Frequency and Time Comparisons by means of Radio Transmissions—J. A. Pierce, H. T. Mitchell and L. Essen. [*Nature (London)*, vol. 174, p. 922; November 13, 1954.] Results obtained at the National Physical Laboratory show that the standard frequency transmission MSF on 60 kc can be used in the United Kingdom with an accuracy of ±1 part in 10^9 in an observation time of a few minutes; it has also been received and measured at Harvard with the same accuracy in a time of about 10 minutes. A typical photographic record taken at Harvard shows variations due to changes in the height of the reflecting ionosphere layers.

621.3.018.41(083.74):621.373.421 1426
High-Stability One-Mc/s Frequency Standard—(*Tech. News Bull. Nat. Bur. Stand.*, vol. 38, pp. 162–163; November, 1954.) Portable equipment developed at the NBS to serve as a secondary standard for general laboratory and field use comprises a crystal-controlled Meacham-bridge oscillator.

621.3.087.9:621-52:621.394.625.2 1427
Automatic Recording and Analysis of Data using Teleprinter Technique—Nettell. (See 1456.)

621.314.7.001.4 1428
A Versatile Transistor Tester—R. Bailey. (*Electronic Eng.*, vol. 27, pp. 64–69; February, 1955.) Equipment is described by means of which the characteristics of *p-n-p* or *n-p-n* transistors can be displayed as families of curves on the screen of a cro accurate to within about 10 per cent. Facilities for point-by-point plotting with greater accuracy are incorporated.

621.317.3:621.317.723 1429
The Measurement of Voltage, Current, Power and Impedance at High Frequencies—D. Karo. (*Beama Jour.*, vol. 62, pp. 33–35; November, 1954.) An electrometer is used with the four plates not connected in pairs, so that it forms a double electrometer with moving vane. Input resistance is $\sim 35M\Omega$ at 1 mc and input capacitance $\sim$ 3 pf. Asymmetry and contact potential effects are eliminated either by calibrating one pair of plates against the other, or by using an electrometer with three pairs of plates. The method of use for each type of measurement and some of the forms the instrument may take are explained.

621.317.3:621.375.3 1430
Magnetic Amplifiers and Weak Currents—R. Klein. (*Bull. Soc. Franç. Élect.*, vol. 4, pp. 649–674; November, 1954.) Particular attention is devoted to the use of magnetic amplifiers for measurement purposes.

621.317.3:621.396.11.029.65 1431
Experimental Equipment and Techniques for a Study of Millimetre-Wave Propagation—Willshaw, Lamont, and Hickin. (*See* 1476.)

621.317.318:621.385.3 1432
The Application of the Electrometer Valve to Charge Measurements—M. J. Morant. (*Jour. Sci. Instr.*, vol. 31, pp. 391–395; November, 1954.)

621.317.33 1433
Point Source and Sink in a Parallelepiped—A. C. Sim and L. Lewin. [*Nature* (*London*), vol. 174, p. 1155; December 18, 1954.] Analysis is presented relevant to the evaluation of the conductivity of semiconductors from measurements by the four-probe method.

621.317.337 1434
The Correction of *Q* Meter Readings—M. V. Callendar and J. P. Newsome. (*Electronic Eng.*, vol. 27, p. 92; February, 1955.) Comment on 3622 of 1954 and author's reply.

621.317.361:519.2 1435
Application of Statistical Theory to the Measurement of Frequency—H. Tanaka, G. Taguchi and M. Ota. [*Rep. Elect. Commun. Lab.* (*Japan*), vol. 2, pp. 9–14; August, 1954.] Frequency deviations of radio sw transmissions are considered. A method is developed for obtaining an unbiased estimate without assuming a normal distribution of the variation of oscillator frequency or of the error in the measurement equipment. This is applied in analyzing the results of measurements made on eight frequencies at three places simultaneously over a period of four days, to obtain probable values for the frequency stability of the transmissions and the accuracy of the measurement equipment.

621.317.39:537.228.1 1436
New Sensitive Piezoelectric Detector—R. Pepinsky, J. Baecklund, H. Diamant and W. G. Perdok. (*Rev. Sci. Instr.*, vol. 25, pp. 1076–1078; November, 1954.) An instrument using the frequency-sweep technique of Perdok and van Suchtelen (1966 of 1950) is described.

621.317.7:621.372.8 1437
Waveguides and Waveguide Junctions—Dukes. (*See* 1255.)

621.317.7.001.4 1438
Test Methods and Construction of Shock-Proof [electrical] Measurement Instruments—A. von Weiss. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 45, pp. 972–977; November 13, 1954.) Usual test methods, such as subjection to acceleration, vibration, or shock, are compared and their practical value is discussed. New shock-proof constructions using pivot and strip suspension systems are described.

621.317.715 1439
Rapid-Oscillation Galvanometers: Part 1—Fundamental Theory—K. H. Winterling. (*Arch. Tech. Messen.*, no. 226, pp. 259–262; November, 1954.) Galvanometers with oscillation period <1 second are considered; these are used mainly as recording instruments. Frequency characteristics and delay times, optimum damping and power requirements are discussed.

621.317.715 1440
The Response of String Galvanometers with Heavy Electromagnetic Damping—D. A. Senior. (*Instr. Practice*, vol. 8, pp. 968–974; November, 1954.)

621.317.729:621.385.032.2 1441
Use of a Rubber Sheet Model for Investigation of Electron Trajectories—K. R. Allen and K. Phillips. (*Electronic Eng.*, vol. 27, pp. 82–85; February, 1955.) Experiments are reported which indicate that the rubber-sheet model can be used satisfactorily in the design of electron guns. A novel method of obtaining equipotentials is described.

621.317.73 1442
Direct-Reading Capacitance Tester—O. E. Dzierzynski. (*Wireless World*, vol. 61, pp. 141–143; March, 1955.) The instrument described covers the range 10 pF–1mF. The basic circuit comprises a 50-cps source driving current through the capacitance in series with a resistance of magnitude such that the current is proportional to the capacitance. The voltage across the resistance is amplified and rectified to give direct readings on a milliammeter.

621.317.73.029.55/.62 1443
Theory of the Reflectometer—H. Wolf. (*Arch. elekt. Übertragung*, vol. 8, pp. 505–512; November, 1954.) Reflectometers comprising an auxiliary line arranged parallel to a main transmission line [477 of 1953 (Grosskopf)] are discussed. Differential equations are established from which voltage and current distributions are calculated. Approximate formulas are presented for the case of loose coupling. A numerical example is given.

621.317.755 1444
A Sensitive Continuously-Evacuated Cathode-Ray Oscillograph—L. J. Griffiths, R. M. Davies and D. A. Richards. (*Electronic Eng.*, vol. 27, pp. 48–52; February, 1955.) An instrument for recording nonrecurrent transients with peak values not exceeding 100 mv and durations up to 100 μs is described. The record is produced by direct action of the electron beam on a photographic emulsion, giving greater photographic sensitivity and lower distortion than when the trace is photographed from a luminescent screen. The apparatus has been used for measurements of stress waves in solids.

621.317.755 1445
A Portable High-Speed Cathode-Ray Oscillograph—L. S. Allard. (*Jour. Sci. Instr.*, vol. 31, p. 433; November, 1954.) Note on precautions required to safeguard the cr tube in the oscillograph described by Waring and Murphy (1862 of 1954).

621.317.772 1446
Direct-Indicating Phase Meter for 50 c/s to 100 kc/s—H. Haller. (*Frequenz*, vol. 8, pp. 325–333; November, 1954.) Rectangular waves derived from the signals under examination are applied to separate grids of a tube which conducts only when the two are of the same sign. The average anode current is then a function of their phase difference. A complete circuit diagram is given and briefly described. 26 references to other phase indicators and comparators are also given and the various principles of operation are classified.

621.317.78 1447
A Very-Wide-Band Dummy Load for measuring Power at Very-high and Ultra-high Frequencies—W. Hersch. (*Proc. IEE*, Part B, vol. 102, pp. 96–98; January, 1955.) A dummy load is described consisting of a short coaxial line filled with carbon tetrachloride, the inner conductor being built up from short sections of thin gold-film resistors in which nearly all the power loss occurs. Its input impedance, of 75Ω, is purely resistive, and it permits measurement of power from 10w to 1kw accurate to within $\pm(2\frac{1}{2}$ per cent of measured power$+3$w) at frequencies over 100 mc.

621.318.4(083.74) 1448
A Study of Absolute Standards of Mutual Inductance and in particular the Three-Section National Bureau of Standards Type—F. W. Grover. (*Jour. Res. Nat. Bur. Stand.*, vol. 53, pp. 297–320; November, 1954.) "The results of a study of the number and location of the circles of zero field surrounding a multisection coil are presented. The configuration of the field surrounding the equatorial region of several three-section coils has been partially mapped out. One arrangement yields a design in which the mutual-inductance contribution of any secondary turn differs but little from that of the median turn. By locating the secondary symmetrically about the circles of zero field, the correction for winding distribution can be made small, and the effect of uncertainties in location of individual secondary turns minimized."

621.373.421 1449
The Bridge-Stabilized Oscillator as Standard Source of A.C. Voltage—G. Hoffmann. (*Arch. Tech. Messen*, no. 226, p. 263–266; November, 1954.) Methods of stabilizing the oscillation amplitude by means of barretter bridges are indicated, and applications of the circuit as a calibration source and as a standard-level source are described.

621.397.5:535.623].001.4 1450
Phase Meter Analyzes Color TV Systems—Houghton. (*See* 1495.)

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.1–8:621.9 1451
Ultrasonic Machining—(*Trans. IRE*, no. PGUE-2, pp. 9–33; November, 1954.) The following three papers are given:
"An Ultrasonic Machine Tool,"—N. Clark, Jr. (pp. 10–18).
"Ultrasonic Machining of Tungsten Carbide," —D. Goetze (pp. 19–22).
"Mechanical Impedance Transformers in relation to Ultrasonic Machining,"—L. Balamuth (pp. 23–33).

534.2–8:620.179.1 1452
How to Check Accuracy of Ultrasonic Flaw Detection—N. Grossman. (*Mater. and Meth.*, vol. 40, pp. 100–101; November, 1954.) Reference blocks machined from regular production materials are used to provide a simple method of setting the equipment to detect flaws above a specified minimum size.

535.4:537.533 1453
An Electron Interferometer—L. Marton, J. A. Simpson and J. A. Suddeth. (*Rev. Sci. Instr.*, vol. 25, pp. 1099–1104; November, 1954.) The instrument described makes use of crystalline diffraction for beam splitting and recombining.

535.4:537.533 1454
The Theory of the Three-Crystal Electron Interferometer—J. A. Simpson. (*Rev. Sci. Instr.*, vol. 25, pp. 1105–1109; November, 1954.)

621.3:621.791 1455
Electronic Control of Resistance Welding—C. R. Bates. (*Jour. Brit. IRE*, vol. 15, pp. 31–46; January, 1955. Discussion, p. 46.) Basic circuits of timing and sequence controls are described and the principles of operation of complete control units are indicated.

621.3.087.9:621–52:621.394.625.2 1456
Automatic Recording and Analysis of Data using Teleprinter Technique—D. F. Nettell. (*Instr. Practice*, vol. 8, pp. 975–980 and 1078–1082; November and December, 1954.) The principles and the advantages of the punched-type system in continuous recording and processing of data are discussed and examples of applications in production-control are given.

621.37/.38:531.383 1457
A Precision Dynamic Balancer (for Small Gyroscopes)—D. Williamson. (*Electronic Eng.*, vol. 27, pp. 53–59; February, 1955.)

621.384.613 1458
Experimental Studies of Betatron Orbit Stability—G. C. Baldwin, F. R. Elder and W. F. Westendorp. (*Jour. Appl. Phys.*, vol. 25, pp. 1553–1554; December, 1954.)

621.384.622 1459
Strong Focusing in Linear Accelerators for Ions—M. Y. Bernard. [*Ann. Phys.* (*Paris*), vol. 9, pp. 633–682; November/December, 1954.] The theory of the motion of heavy particles in a linear accelerator, and the strong-focusing principle, are examined in detail. Theoretical and experimental work is reported on the design of lenses for strong-focusing devices, both electrostatic and magnetic types being practicable.

621.385.833 1460
Some Properties of Symmetrical Slit ("Cylindrical") Electron Lenses—G. D. Archard. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 395–399; November, 1954.) "The focal lengths of a series of symmetrical slit ("cylindrical") electrostatic electron lenses are determined as a function of aperture size and applied potential. It is shown that all practical lenses may be represented on a single curve by means of a parameter simply related to potential and geometry. From this it is deduced that lens strength depends more on the size of the central aperture than on that of the outer apertures."

621.387.424 1461
Time-Delay Distribution in Geiger-Müller Counters—E. Picard and A. Rogozinski. (*Jour. Phys. Radium*, vol. 15, pp. 767–775; November, 1954.) Statistical analysis of results of delay measurements shows that to minimize the response-time delay and make it largely independent of the sensitivity of the associated amplifier the counter should have as small a diameter and as high an applied voltage as possible.

621.387.424 1462
The Ionic Multiplication Coefficient in CO_2-Filled Counters—R. Fourage and L. Miramond. (*Jour. Phys. Radium*, vol. 15, pp. 780–781; November, 1954.)

621.387.424 1463
Influence of the Limited Proportionality Region on the Value of the Geiger Threshold of Counters with Externally Graphited Glass Cylinders—D. Blanc. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 239, pp. 1621–1622; December 8, 1954.]

621.387.464 1464
The Scintillation Counter in Industry with Reference to Stability Problems—J. S. Eppstein. (*Jour. Brit. IRE*, vol. 15, pp. 25–29; January, 1955.) A circuit is described for minimizing the effects of changes of photomultiplier gain.

621.56:537.311.33:537.322.1 1465
The Use of Semiconductors in Thermoelectric Refrigeration—H. J. Goldsmid and R. W. Douglas. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 386–390; November, 1954.) Factors governing the selection of suitable semiconductors are discussed. The material should have a high atomic weight and the thermoelectric power should lie between 200 and 300 μv per degree C. Calculations show that the maximum cooling obtainable with an *n*-type-Ge/metal junction is about 2 degrees C., but that cooling of as much as 28 degrees C. should be obtainable with a *p*-type-Bi_2Te_3/Bi junction; cooling of 26 degrees C. was actually obtained using elements of length 1 cm and respective cross-sections of 0.25 cm^2 and 0.03 cm^2, with a current of 6.5a. The calculated figure of merit was 0.475. The possible application of the inverse effect in solar-energy thermoelectric generators is mentioned.

PROPAGATION OF WAVES

538.566 1466
Reflection of Waves by an Inhomogeneous Medium—V. A. Bailey. (*Phys. Rev.*, vol. 96, pp. 865–868; November 15, 1954.) New approximations are proposed which give good agreement with the known fundamental pair of solutions for the general second-order differential equation of propagation, but which remain finite. A corresponding approximate expression is derived for the coefficient of reflection of plane waves by a specified inhomogeneous medium; by means of an iterative process, a sequence of approximations is developed which converge rapidly on the true value.

538.566:535.312 1467
Statistics of Electromagnetic Radiation Scattered by a Turbulent Medium—R. A. Silverman and M. Balser. (*Phys. Rev.*, vol. 96, pp. 560–563; November 1, 1954.) "The theory of Villars and Weisskopf [2747 of 1954] is used to calculate the univariate and bivariate amplitude distributions of electromagnetic radiation scattered from turbulent fluctuations. The univariate distribution is Rayleigh, and is in excellent agreement with measurements made on a 49.6-Mc/s ionospheric scatter link. Using the bivariate distribution, we relate the amplitude correlation function to a velocity correlation function which appears in the von Weizsäcker-Heisenberg statistical theory of turbulence. In this way the theoretical velocity correlation can be compared with experiment."

538.566:535.42 1468
Diffraction of Electromagnetic Waves at Non-ideally Conducting Wedge and Problem of Coastal Refraction—V. A. Il'in. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 99, pp. 47–50; November 1, 1954. In Russian.] Results are given of the theoretical consideration of four cases: (a) diffraction of em waves from point or line source, (b) an alternative method for (a) for the particular case when the wedge angle is an integral submultiple of π, (c) diffraction of a plane wave, and (d) coastal diffraction. The solutions are obtained by modification of the known solutions for an ideally conducting wedge as in 17 of February. In case (d) the conductivity is assumed to vary continuously at the sea-land boundary; for waves with the electric field perpendicular to the coastline the signal strength should increase towards the coastline, for the magnetic field perpendicular to the coastline it decreases.

621.396.11:551.510.535 1469
Simultaneous Ionospheric Absorption Measurements at Widely Separated Stations—W. J. G. Beynon and K. Davies. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 273–289; November, 1954.) Noon vertical-incidence measurements for the period September, 1950–August, 1951, made at Slough and Swansea (230 km apart), are compared. The frequency used was 2 mc. The average correlation coefficient is 0.85 for winter months and 0.81 for summer months. Oblique-incidence intensity measurements were also made on cw signals of equivalent frequencies from Skelton, about 366 km north of Swansea, and from Allouis, about 666 km southeast of Swansea. The observations give some indication of the progressive decrease in the degree of correlation with increasing distance between stations. The use of simple field-strength recording as a measure of the absorption is briefly considered.

621.396.11:551.510.535 1470
The Physics of the Ionosphere—(*See* 1340.)

621.396.11:551.594.5 1471
Polarization of Radio Echoes from Aurorae—A. G. McNamara and B. W. Currie. [*Nature* (*London*), vol. 174, pp. 1153–1154; December, 18, 1954.] Observations made at Saskatoon are described and the results analyzed. Frequencies of 56 and 106.5 mc were used. Horizontally polarized signals were transmitted towards magnetic north, and a multiple antenna system was used to determine the angle of arrival and the different types of polarization of the echoes. A histogram shows numbers of echoes for different values of polarization ratio (ratio of horizontal to vertical component of field strength). A wide variation of this ratio was observed on 56 mc; on 106.5 mc all the echoes were polarized in the same plane as the transmitted signals. Multiple scatter within the reflecting region is considered more likely than magneto-ionic effect to cause depolarization.

621.396.11:551.594.6 1472
The Wave-Forms of the Electric Field in Atmospherics recorded simultaneously by Two Distant Stations—Norinder. (*See* 1348.)

621.396.11.029.6 1473
Review of Long-Distance Radio-Wave Propagation above 30 Mc/s—W. J. Bray, H. G. Hopkins, F. A. Kitchen and J. A. Saxton. (*Proc. IEE*, Part B, vol. 102, pp. 87–95; January, 1955.) Factors affecting normal propagation are reviewed. Surface roughness modifies the structure of the refractive-index profile. Diffraction effects produced by single large obstacles and by groups of obstacles are indicated. Attenuation due to absorption by oxygen and water vapor, and to absorption and scattering by clouds, fog and precipitation, become significant at frequencies over 3 kmc. Ionospheric propagation via the F_2 layer, via the E_s layer, by scattering from the *E* layer and via the auroras, and tropospheric propagation by way of elevated inversions, under conditions of superrefraction near the earth's surface, and by way of scattering from atmospheric turbulence, are discussed briefly. 61 references.

621.396.11.029.62 1474
Some Features of V.H.F. Tropospheric Propagation—M. W. Gough. (*Proc. IEE*, Part B, vol. 102, pp. 43–58; January, 1955.) An analysis is made of continuous vhf signal-strength records over numerous and diverse paths in tropical and Mediterranean regions; the observed variations can be satisfactorily explained in terms of well-known tropospheric states. In some cases it is possible to reconstruct from the records the salient features of tropospheric *M* profiles operative during the tests. Prolonged measurements over a Mediterranean path have yielded statistics of variations of the effective earth-radius factor which should be useful for predicting the statistical behavior of other paths in that area. A limited connection is found between median path attenuation and fading range for paths experiencing comparable tropospheric influences; on this basis the statistical behavior of paths can be forecast approximately without a preliminary radio survey.

621.396.11.029.62/.63:551.509 1475
Propagation on 144 and 420 Mc/s—C. E. Newton, G. M. C. Stone, A. J. Worrall and H. W. Parker. (*R.S.G.B. Bull.*, vol. 30, pp. 210–215; November, 1954.) A simple introduction is given to the interpretation of meteorological conditions for forecasting anomalous propagation, with particular reference to the London region.

621.396.11.029.65:621.317.3 1476
Experimental Equipment and Techniques for a Study of Millimetre-Wave Propagation—W. E. Willshaw, H. R. L. Lamont, and E. M. Hickin. (*Proc. IEE*, Part B, vol. 102, pp. 99–111; January, 1955.) Report of a program of research, carried out between 1945 and 1949, on propagation at wavelengths of 4–6 mm. Crystal harmonic generators and pulse and cw magnetrons were used as generators; the design of crystal detectors and mixers is discussed. Measurements were made of attenuation due to oxygen absorption over paths of a few kilometers; values of the reflection coefficient of the sea surface were derived.

RECEPTION

621.376.232.2.018.75 1477

Pulse Response of Signal Rectifiers—W. B. Lewis. (*Wireless Eng.*, vol. 32, p. 89; March, 1955.) Comment on 849 of April (Callendar), referring to previously published partial solutions of some of the problems discussed (*ibid.*, vol. 9, pp. 487–499; September, 1932.)

621.396.62:621.314.63 1478

Influence of Image Frequency in Microwave Receivers with [crystal] Detector Mixer—E. Willwacher. (*Fernmeldetech. Z.*, vol. 7, pp. 608–615; November, 1954.) Conversion loss and IF output admittance are considered; analysis is presented indicating how the terminating admittance for the image frequency should be taken into account in the design of the receiver. A numerical example is given. Experiments with a receiver for the frequency range 2.1–2.3 kmc emphasize the importance of providing a line of suitable length between the input filter and the mixing circuit in order to keep the noise factor constant.

621.396.62:621.314.7 1479

Transistor Radio Receiver—D. Nappin. (*Wireless World*, vol. 61, pp. 123–124; March, 1955.) Details are given of a four-stage receiver with rf amplifier and push-pull output, using five point-contact transistors of which only that in the rf stage needs to be specially selected for a particular characteristic. The output is about 50 mw.

621.346.62.029.62:621.314.7 1480

A Transistor Superregenerative Receiver for 10 and 6 Meters—W. A. Wadsworth. (*QST*, vol. 38, pp. 17, 134; November, 1954.) A two-transistor receiver is briefly described. The total power requirement of 0.06w is provided by two 7.5-v batteries.

621.396.621:621.396.822 1481

Recovering Hidden Signals—J. Franklin. (*Wireless World*, vol. 61, pp. 137–140; March, 1955.) An outline is given of the use of correlation techniques, especially in relation to the detection of periodic signal components masked by random variations.

621.396.621.54 1482

Two-Valve Superhet—R. C. Lever. (*Wireless World*, vol. 61, pp. 145–146; March, 1955.) The circuit uses a Type-12AH8 triode-heptode as frequency changer and a Type-ECL80 triode pentode as detector and af-amplifier. The conventional IF amplifier is omitted, and the required gain is provided by regeneration.

621.396.621.54 1483

The TLR 190: a Novel 4-Valve Superheterodyne Receiver—J. Marsac. (*Toute la Radio*, no. 190, pp. 387–391; November, 1954.) Good sensitivity, selectivity and fidelity together with simplicity are claimed for a receiver using a high-slope IF amplifier with a cathode-loaded detector which is capable of producing an avc voltage for application to the IF suppressor grid.

621.396.621.54:621.314.7 1484

The Transistor as a Mixer—P. D. Strum and J. Zawels. (PROC. IRE, vol. 43, p. 230; February, 1955.) Comment on 1898 of 1954 and author's reply.

621.396.812.3 1485

Performance of Space and Frequency Diversity Receiving Systems—R. E. Lacy, M. Acker and J. L. Glaser. [*Proc. IRE (Australia)*, vol. 15, pp. 313–317; December, 1954. *Convention Record IRE*, Part 2, 148–152; 1953]. Report of an investigation covering a period of four years, carried out over three different paths. Six different frequencies were used, in the range 2.63–22 mc. With the space-diversity systems, little improvement was obtained on increasing antenna separation beyond about 600 feet. Differences are indicated between the results for separation perpendicular to and parallel to the direction of propagation. The improvement in reception varied inversely as the frequency. With a dual frequency-diversity system operating at 11.66 mc nominal, the improvement was only slightly less than with a dual space-diversity system with 600-feet separation.

STATIONS AND COMMUNICATION SYSTEMS

534.861 1486

Radio Studios—W. Furrer. (*Tech. Mitt. Schweiz. Telegr.-Teleph Verw.*, vol. 32, pp. 430–436; November 1, 1954. In German.) A brief review of the requirements as regards electro-acoustic equipment and acoustic characteristics, with illustration of studio details.

621.39.001.11 1487

Information Theory and Knowledge—A. Gamba. (*Jour. Appl. Phys.*, vol. 25, p. 1549; December, 1954.) Comment on 3032 of 1954 (MacDonald).

621.39.001.11 1488

Quantity of Information in a Gaussian Stationary Process contained in a Second Process Stationarily Connected with it—M. S. Pinsker. [*Compt. Rend. Acad. Sci. (URSS)*, vol. 99, pp. 213–216; November 11, 1954. In Russian.] Formulas are derived for determining the entropy and quantity of information in a Gaussian stationary process, when the spectral functions are known. See also 3105 of 1951 (Elias).

621.396.44:621.395.82 1489

The Influence of Radio Transmissions on Carrier-Frequency Telephony on High-Voltage Lines—J. Grosskopf. (*Fernmeldetech. Z.*, vol. 7, pp. 623–636; November, 1954.) An extensive experimental investigation was made of the interference effect of communication and broadcast transmissions in the frequency range 100 kc–1.5 mc, both by direct measurement of the interference voltage and by estimating it from the strength of the local rf field. The interference voltages measured at the carrier-frequency input at the telephony stations were analyzed statistically and related to the distance from the interfering transmitters and the position of the hv lines.

621.396.712(494) 1490

Extension of the Swiss National Broadcasting Station at Sottens—R. Pièce. (*Tech. Mitt. Schewiz. Telegr.-Teleph Verw.*, vol. 32, pp. 409–430; November 1, 1954. In French.) A short historical account is given of the development of the Sottens station, followed by a detailed description of equipment installed in 1951. The existing 100-kw transmitter has been supplemented by two 100-kw units which can be operated in parallel or independently, thus avoiding a break in transmission if one of the units fails.

SUBSIDIARY APPARATUS

621.316.722.1:621.387 1491

The Processing and some Characteristics of Coaxial Cylinder, Corona Stabilizer Tubes—A. J. L. Collinson and D. W. Hill. (*Jour. Sci. Instr.*, vol. 32, pp. 13–17; January, 1955.) The running voltage of tubes filled with hydrogen has been found to be stable to within 0.02 per cent over 24 h and to within 1 per cent over 1,000 h. The dependence of slope resistance on pressure and current has been investigated. The temperature coefficient of running voltage is about +0.005 per cent per degree C. It may be possible to obtain higher current by glow discharge treatment.

621.319.3 1492

A New Type of High-Voltage Machine—R. E. D. Clark and F. T. Farmer. [*Nature (London)*, vol. 174, pp. 1065–1066; December 4, 1954.] A machine is described which operates on a principle related to that of the van de Graaff generator, with the difference that the moving belts carry not surface charges but capacitors which become charged on reaching a certain point and are then connected in series. Performance figures are given for an experimental model in which the capacitors are stationary and the electrical contacts are rotated.

TELEVISION AND PHOTOTELEGRAPHY

621.397.242:621.376.2 1493

Modulators and Demodulators for the Transmission of Television Pictures via Cables—W. Dillenburger. (*Fernmeldetech. Z.*, vol. 7, pp. 589–596; November, 1954.) German practice, where the distance between studio and transmitter is several kilometers, is to transmit the picture signal on a 21-mc carrier. Systems with the modulation applied to the control grid are preferred to those with the modulation on the suppressor grid. Automatic control of the synchronizing-pulse level at the demodulator output is necessary; the simplest method is to subject the synchronizing pulses to enhanced amplification followed by limiting. A magnetic voltage stabilizer is included in the tube-heater circuit. A simple black-level-setting arrangement including a keyed diode is used.

621.397.5 1494

Special Television Numbers—(*Onde élect.*, vol. 34, pp. 824–945 and 958–1040; November and December, 1954.) Papers are included on numerous aspects of the subject, including the physiological optics of viewing, a spiral scanning system, transmitting-station equipment, the "cigar type" and related types of antenna for long-distance reception on meter wavelengths, asymmetrical-sideband transmission systems, European program-exchange arrangements, outside-broadcast equipment, cm-λ link equipment, large-screen projection systems, telecine and telerecording equipment, and industrial television equipment.

621.397.5:535.623].001.4 1495

Phase Meter analyzes Color TV Systems—R. W. Houghton. (*Electronics*, vol. 28, pp. 156–160; January, 1955.) The phase specifications for the NTSC system are discussed. To ensure that there is no perceptible color error, receiver phase errors must be below 5 degrees, corresponding to a time delay of 0.004 μs at the color subcarrier frequency of 3.58 mc. Test equipment suitable for use in receiver factory or broadcast studio is described.

621.397.5:778.5 1496

The Recording, Reproduction and Processing of Sound Films for Television—H. Friess. (*Fernmeldetech. Z.*, vol. 7, pp. 346–348; July, 1954.)

621.397.5(083.74)(94) 1497

Australian Television. Revision of Technical Standards—[*Proc. IRE, (Australia)*, vol. 15, pp. 318–320; December, 1954.] The revised standards issued on October 1, 1954 are presented.

621.397.5(083.74)(94) 1498

Technical Requirements of the Australian Television System—A. J. McKenzie. [*Proc. IRE (Australia)*, vol. 15, pp. 301–312; December, 1954.] A table is presented of the television standards of various countries, including those for the projected Australian system. The latter are discussed in detail; they differ only slightly from the CCIR 625-line standards, the most important difference being that a channel width of 7.5 mc is specified as compared with the CCIR 7 mc; this is largely with a view to the later introduction of color television. Receiver IFs of 31.25 mc and 37.25 mc for sound and picture respectively are recommended. Frequency details are given of vhf and uhf channel reservations. It is estimated that an allocation of three stations to each capital city and one to each town with a population over 5,000 is reasonable, subject to economic considerations.

621.397.6 1499
Industrial Television Equipment using Spiral Scanning—A. V. J. Martin. (*Toute la Radio*, no. 190, pp. 357–363; November, 1954.) Description of equipment developed by Derveaux, in which the scanning waveform is a 15-kc sine wave modulated with a 50-cps sawtooth.

621.397.62 1500
The 625-Line C.C.I.R. System: a Line Timebase Investigation—R. R. Harman. (*Jour. Telev. Soc.*, vol. 7, pp. 334–337; October/December, 1954.) From field trials in Holland of a prototype receiver which incorporated both normal and flywheel synchronization circuits, the necessity of providing a flywheel circuit in order to obtain acceptable pictures for input signals <1 mv and its desirability even at higher input levels were established. The circuit will also deal with pulse interference and receiver noise, when these are not severe.

621.397.62(083.74) 1501
An Intermediate Frequency for C.C.I.R. Television Standards—H. B. S. Brabham. (*Jour. Telev. Soc.*, vol. 7, pp. 338–342, 346; October/December, 1954.) Conditions are discussed for ensuring that a superheterodyne receiver shall be free from spurious response generated in the mixer and detector, shall not itself cause interference and shall be reasonably immune from outside interference. A chart is presented which shows that in order to avoid spurious responses, the IF must have a mid-band frequency between 36 and 39 mc. 37.25 mc is recommended. This choice will also avoid local-oscillator radiation within band I. Field tests in Holland gave satisfactory results. External transmitter interference can be dealt with by providing a simple frequency-selective trap.

621.397.621.2 1502
Spurious Line Scan Resonances—K. G. Beauchamp. (*Wireless World*, vol. 61, pp. 109–114; March, 1955.) The generation of various types of unwanted oscillation by the line-scan-circuit output tube is discussed and methods for reducing their effects on the picture are outlined.

621.397.7+621.397.61 1503
Rowridge Television Transmitting Station—[*Engineer* (*London*), vol. 198, p. 711; November 19, 1954.] Brief details are given of this BBC station, where regular service was inaugurated in November, 1954. Operation is in channel 3, with vision on 56.75 mc and sound on 53.25 mc. Modulation of the main vision transmitter is at low level; peak-white output is 5 kw; no separate filter is used for shaping the vestigial sideband. The sound transmitter is a conventional class-B modulated equipment rated at 2 kw output. The vision program reaches Rowridge from London via a repeater station and a microwave radio link; the sound program is conveyed by telephone circuits. A single tower carrier the microwave receiving antenna and the two-stack-dipole temporary transmitting array.

621.397.7 1504
Operating Experiences with the B.B.C. Television Studios at Lime Grove—D. C. Birkinshaw. (*Jour. Telev. Soc.*, vol. 7, pp. 315–332; October/December, 1954.) Developments at Lime Grove since April, 1952 are critically discussed. Extra channels including a flying-spot/Mechau system have been added to the telecine equipment. For telerecording the inverted Mechau system now has a better performance than the suppressed-frame system. Since all large studios have to be used for different types of production, a compromise acoustic treatment imparting a moderate degree of "aliveness" has been given by applying a certain amount of plasterboard to the walls. Lighting technique is under review. Complex vision mixers are to be replaced by simpler units. Other equipment and problems are also discussed.

621.397.7 1505
Special Effects for Television Studio Productions—A. M. Spooner and T. Worswick. (*Proc. IEE*, Part B, vol. 102, p. 68; January, 1955.) Discussion on 561 of 1954.

621.397.7:535.623:628.972 1506
Color Television Light Sources—H. M. Gurin. (*Jour. Soc. Mot. Pic. Telev. Engrs.*, vol. 63, pp. 51–54; August, 1954.) Factors leading to the choice of tungsten illumination for use with the NTSC system are discussed. Operating experience is described.

621.397.743 1507
I.T.A. [Independent Television Authority] Transmitters—(*Wireless World*, vol. 61, pp. 120–122; March, 1955.) Problems of siting the ITA band-III stations are discussed. Co-siting with BBC band-I stations leads to simplification of receiving antennas but involves difficulties in the design of the transmitting antennas. Because of the different population distributions in the service areas of the BBC stations, the desirability of co-siting must be considered separately for each case.

TRANSMISSION

621.396.61:621.373.42 1508
Navy Transmitter uses Frequency Synthesizer—H. Romander and R. Watson. (*Electronics*, vol. 28, pp. 138–143; January, 1955.) A transmitter for telegraphy or facsimile is set to any frequency from 30 to 600 kc, accurate to within 3 cps, by setting four dials. Details are given of the signal generator, in which the desired frequency is obtained by synthesis of selected multiples of 1 kc, using a continuously variable Wien-bridge oscillator for interpolation.

TUBES AND THERMIONICS

621.314.63+621.314.7 1509
Semiconductor Devices made with Single Crystal Germanium-Silicon Alloys—R. A. Logan, A. J. Goss and M. Schwartz. (*Jour. Appl. Phys.*, vol. 25, pp. 1551–1552; December, 1954.) The formation of *p-n* junctions by solid-phase diffusion of impurities into the crystal [2226 of 1954 (Pearson and Fuller)] is well suited for use with Ge-Si alloys. Characteristics of some diodes and transistors are reported briefly.

621.314.63:546.289 1510
A Note on the Effect of Temperature Gradients at Point Contacts on Germanium—H. K. Henisch and P. M. Tipple. (*Proc. Phys. Soc.*, vol. 67, pp. 651–652; August 1, 1954.) The effect of a temperature gradient on the resistance of the point contact is shown graphically, data being obtained from pulse measurements and by inference from dc measurement; for a given contact temperature the dc reverse resistance increases with increasing temperature gradient. If there is any appreciable hole current at large reverse voltages, the current should be sensitive to changes in the hole replacement rate. This should depend on temperature gradient as well as on actual contact temperature. An alternative explanation is considered, but it fails to account for the magnitude of the resistance changes.

621.314.63:546.623.86 1511
Use of AlSb for Crystal Valves—G. Zielasek. (*Arch. elekt. Übertragung*, vol. 8, pp. 529–533; November, 1954.) Properties of AlSb relevant to its use in diodes are compared with those of Ge and Si; operational and economic advantages are indicated. For small signals, the forward slope of the AlSb diode is higher than that of the Ge diode. The diffusion voltage of AlSb is less than that of Ge and Si.

621.314.632:546.28 1512
A New High-Temperature Silicon Diode—C. G. Thornton and L. D. Hanley. (PROC. IRE, vol. 43, pp. 186–188; February, 1955.) An account is given of the preparation of a point-contact diode by a process involving bombardment of the Si surface with oxygen ions, using technique described by Ohl (1337 of 1952). The diode is characterized by (a) very low saturation currents, of the order of 1μa, (b) high inverse-voltage operation (70–200 v), (c) low barrier capacitance (<0.3 pF), and (d) satisfactory operation at temperatures up to 200 degrees C.

621.314.632:621.372.632 1513
Crystal Convertor—J. Shimizu. [*Rep. Elect. Commun. Lab.* (*Japan*), vol. 2, pp. 1–8; August, 1954.] The processing and design of a Si crystal for frequency conversion in microwave receivers and transmitters are discussed. Experimental studies are reported on the effect of contact pressure and heat treatment on the dc characteristic and in particular the influence of spreading resistance and barrier capacitance on the conversion loss. Results of measurements made at 4 kmc on a series of rectifier type IN23B and SH5A are given.

621.314.7 1514
Reliability of Quantity-Produced Transistors in Low-Power Audio Applications—F. M. Dukat. (*Trans. IRE*, no. PGQC-4, pp. 32–39; December, 1954.) A report based on the performance of about one million diffused-junction *p-n-p* transistors, with plastic encapsulation and leads sealed in glass, used in hearing aids.

621.314.7 1515
Theory of Diffusion-Type and Drift-Type Transistors: Part 3—Design—H. Krömer. (*Arch. elekt. Übertragung*, vol. 8, pp. 499–504; November, 1954.) In drift transistors with the drift field produced by nonuniform doping, undesirably high emitter capacitance can be reduced by using several different semiconductors. The field can alternatively be produced by altering the width of the forbidden band, by preparing the base from a nonstoichiometric mixed crystal (e.g. Ge-Si) of graded composition. Variants of this method are indicated. If the current density is above a certain value the drift transistor behaves as an ordinary diffusion type; high distortion is produced in the transition range. Requirements for avoiding this condition are indicated. The characteristic curves are examined; they are straighter and more nearly parallel for the drift transistor than for the diffusion type. The results indicate that for the same base thickness the drift type is superior as regards frequency cut-off, gain, and undistorted power. Part 2: 579 of February.

621.314.7 1516
Introduction to Transistor Electronics—H. K. Milward. (*Wireless World*, vol. 61, pp. 60–65 and 133–136; February and March, 1955.) The operation of junction transistors is explained simply in terms of the potential gradients and current-carrier paths in the semiconductor.

621.314.7:546.289 1517
On the Current Gain of Germanium Filamentary Transistors—R. Lawrance, A. F. Gibson and J. W. Granville. (*Proc. Phys. Soc.*, vol. 67, pp. 625–635; August 1, 1954.) Current-gain/temperature measurements were made on filamentary transistors of *n*-type Ge single crystals having resistivities between 2 and 8Ω. cm. Results indicate that hole trapping is responsible for the high current gain observed at low temperatures and low emitter currents. Evidence from ion-bombardment and illumination experiments suggests that the holes are trapped in a surface channel produced by the minimum potential energy for holes near the

surface. The mobility of a hole in the channel is believed to be very small and the lifetime of a hole in the channel very long.

621.314.7:546.289 1518

Current Multiplication Processes in *n*-Type Germanium Point-Contact Transistors—C. A. Hogarth. (*Proc. Phys. Soc.*, vol. 67, pp. 636–643; August 1, 1954.) Experiments are reported which indicate that, for type-A transistors, the rise in current gain associated with low emitter current is a feature of the collector contact and is not associated with the type of traps discovered by Lawrance (2332 of 1953 and 2130 of 1954). The increase in current gain could be accounted for by a decrease in barrier height due to holes arriving in the vicinity of the collector. Experiments gave results supporting this theory.

621.314.7:621.375.4.029.3:621.396.822 1519

Investigations of Noise in Audio Frequency Amplifiers using Junction Transistors—Bargellini and Herscher. (*See* 1302.)

621.314.7.001.4 1520

A Versatile Transistor Tester—Bailey. (*See* 1428.)

621.383.5 1521

Physical Properties of Thin Photosensitive Films of Cadmium Sulphide—G. Wlérick and F. Prégermain. (*Jour. Phys. Radium*, vol. 15, pp. 757–764; November, 1954.) The suitability of CdS cells for measurements of low luminous intensities was investigated. For each cell tested a critical temperature seems to exist above which there is no equilibrium value of current. At lower temperatures photoelectric inertia is troublesome, and it is essential to use only cells that have been kept in the dark for at least 24 hours. A self-quenching effect at high intensities and anomalies in the initial slope of the decay curve after exposure to light before the commencement of the test were also observed.

621.385.029.6 1522

Complex-Frequency Analysis for Transit-Time Valves—F. W. Gundlach. (*Fernmeldetech. Z.* vol. 7, pp. 598–607; November, 1954.) Conditions for growth of oscillations in transit-time tubes are analyzed in terms of complex frequencies and complex transit angles by extending the linear theory presented previously (1825 of 1950). The solutions depend on the frequency variation of the admittance of the tube-circuit combination. The admittance variations with complex frequency are determined for a number of passive circuit elements and are shown graphically. Types of tube considered are (a) space-charge diode, (b) reflex klystron, (c) dynatron, and (d) traveling-wave.

621.385.029.6 1523

Noise in Cut-Off Magnetrons—R. C. Glass, G. D. Sims and A. G. Stainsby. (*Proc. IEE*, Part B, vol. 102, pp. 81–86; January, 1955.) Measurements have been made of the noise fluctuations in the anode current at anode voltages ranging from zero to the value at which oscillations start. Noise peaks are observed at frequencies which vary in a regular manner with anode voltage and with magnetic field strength. The results indicate that the space charge is in a state of oscillation even at very low voltages, and these oscillations explain the very high electron temperatures that have been observed. The results also explain the high rates of rise of rf amplitude obtained with pulsed magnetrons. It seems probable that the pre-oscillation modes largely determine the oscillation mode. An investigation of the influence of parameters such as cathode diameter on the space-charge oscillations should be useful.

621.385.029.6 1524

A Coupled-Mode Description of Beam-Type Amplifiers—H. Heffner. (PROC. IRE, vol. 43, pp. 210–217; February, 1955.) Analysis for klystron and traveling-wave tubes is presented in which the electron beam and the circuit are treated as coupled transmission lines. Expressions are derived for the ac power flow.

621.385.029.6 1525

On the Minimum Noise Figure of Traveling-Wave Tubes—S. W. Harrison. (PROC. IRE, vol. 43, p. 227; February, 1955. *Sylvania Technologist*, vol. 7, p. 123; October, 1954.) A general proof is given that minimum noise is attained at the same beam voltage as maximum gain, assuming absence of correlation between initial-velocity and initial-current fluctuations.

621.385.032.216 1526

International Congress to mark the Fiftieth Anniversary of the Oxide Cathode—(*Le Vide*, vol. 9, pp. 224–316; November, 1954.) The text is given of the following further papers:

"Properties of Hot Cathodes for Microwave Valves,"—D. Charles and H. Huber (pp. 224–233).

"Examples of Hot Cathodes for Microwave Valves,"—H. Huber and D. Charles (pp. 234–243).

"Cathodes for Magnetrons,"—A. Dubois (pp. 244–248).

"Conditions Imposed on Hot Cathodes for C.R. Tubes,"—C. Dufour and G. Wendt (pp. 249–252).

"Oxide Cathodes in Gas-Discharge Lamps,"—P. L. Voreaux (pp. 253–256).

"A Self-Heating Emissive Electrode with Fused Active Layer, for Luminescent and Fluorescent Tubes,"—P. Delrieu and R. Penon (pp. 257–260).

"Cathodes for Hydrogen-Filled Thyratrons,"—P. Leduc (pp. 261–265).

"Active Cold Cathode for Gas-Filled Tubes,"—C. Biguenet (pp. 266–268).

"Applications of the L Cathode,"—A. Venema (pp. 269–272).

"Ionic Impact and Gas Contamination of the L Cathode,"—S. Fugawa and H. Adachi (pp. 273–278). English version included.

"Evaporation of the Barium from Capillary Metal Cathodes,"—H. Katz (pp. 279–283). German version included.

"The 'Impregnated' Cathode and its Structural Characteristics compared with the 'L' Cathode,"—R. Levi (pp. 284–289). English version included.

"Thermionic Properties of Moulded Cathodes Made from Metal and Alkaline-Earth Oxides,"—R. Uzan (pp. 290–296).

"Directly Emissive Sintered Thermionic Cathodes,"—Nguyen Thien-Chi and P. Dussaussoy (pp. 297–301).

"A New Type of Diffusion Cathode,"—A. H. Beck, A. D. Brisbane, A. B. Cutting and G. King (pp. 302–309). English version included.

"Preliminary Work on Cathodes of the 'Barium-Salt-Based Moulded' Type,"—H. Huber and J. Freytag (pp. 310–316).

For previous lists see 1216 of May.

621.385.15:621.318.57 1527

Storage Tube employs Secondary Emission—W. E. M. Uhlmann. (*Electronics*, vol. 28, pp. 161–163; January, 1955.) A switching tube ("memotron") designed for telephone systems is described. The anode comprises a cylinder surrounding a coaxial squirrel cage, with the 20 secondary emitting storage elements located in the annular space between them.

621.385.3+621.385.5 1528

Electron-Optical Treatment of Current Division in Radio Valves—S. Deb. (*Indian Jour. Phys.*, vol. 28, pp. 348–364; August, 1954.) The method of investigation described previously in relation to the planar triode (1533 of 1953) is extended to the case of nonplanar triodes and pentodes. Formulas are derived for the focal lengths of the aperture constituted by adjacent grid wires and the ratio of grid current to total current in a cylindrical triode, and for the ratio of screen-grid current to anode current in a pentode. Experiments with rubber-membrane models are reported.

621.385.832 1529

Use of Electron Guns with Spherically Curved Electrodes in Cathode-Ray Tubes—J. Laborderie. (*Ann. Radioélect.*, vol. 9, pp. 366–373; October, 1954.) The design of a Pierce-type gun for low-intensity beams is considered, in particular the ratio of the radius of curvature of the cathode to that of the anode, which ratio must be large. In an experimental tube with cathode-target distance 20 cm and voltage 1 kv, spot radius was 0.15 mm with current density 50 ma/cm².

621.385.832:681.142 1530

The Physics of Cathode-Ray Storage Tubes—C. N. W. Litting. (*Jour. Sci. Instr.*, vol. 31, pp. 351–356; October, 1954.) A nonmathematical account is given of the principal theories of the cr-storage-tube mechanism and of experiments made to test these theories. Theory based on the triode tube explains the experimental results obtained at Manchester and also explains the continued usefulness of the original theory of Williams and Kilburn (2258 of 1949).

621.387:621.318.57 1531

A Novel Gas-Gap Speech Switching Valve—A. H. Beck, T. M. Jackson and J. Lytollis. (*Electronic Eng.*, vol. 27, pp. 7–12; January, 1955.) Cold-cathode gas-filled tubes are described, designed to give low noise (at least 60 db below signal level) and low striking resistance (about 100 Ω) independent of signal frequency up to at least 50 kc. The electrode arrangement comprises cathode and twin anodes; several practical forms are illustrated. Multi-gap and single-gap variants are also discussed. The devices are intended for telephone switching applications.

621.387.032.213 1532

The Emission from Hot Cathodes in Gas Discharges—A. E. Pengelly and D. A. Wright. (*Brit. Jour. Appl. Phys.*, vol. 5, pp. 391–395; November, 1954.) The emission was investigated experimentally for both directly and indirectly heated cathodes, over a wide range of current. The condition of zero-field thermionic emission was identified in the presence of a discharge in inert gas; emission from cathodes in Ar, at zero field, is generally equal to the emission in vacuum. At very high currents the total electron emission is greater than the thermionic emission, either because of field enhancement or because of the Townsend γ multiplication process at the cathode. Results are tabulated and presented graphically.

MISCELLANEOUS

061.6:[621.396.9+681.142 1533

Electronics Research Laboratory in Scotland—[*Engineer* (*London*), vol. 198, pp. 565–566; October 22, 1954.] A brief account of the layout of the new Ferranti laboratory at Edinburgh and mention of some of the items under investigation, including control of machines by digital computers, and radar development. A similar account is given in *Engineering* (*London*), vol. 178, pp. 537–538; October 22, 1954.

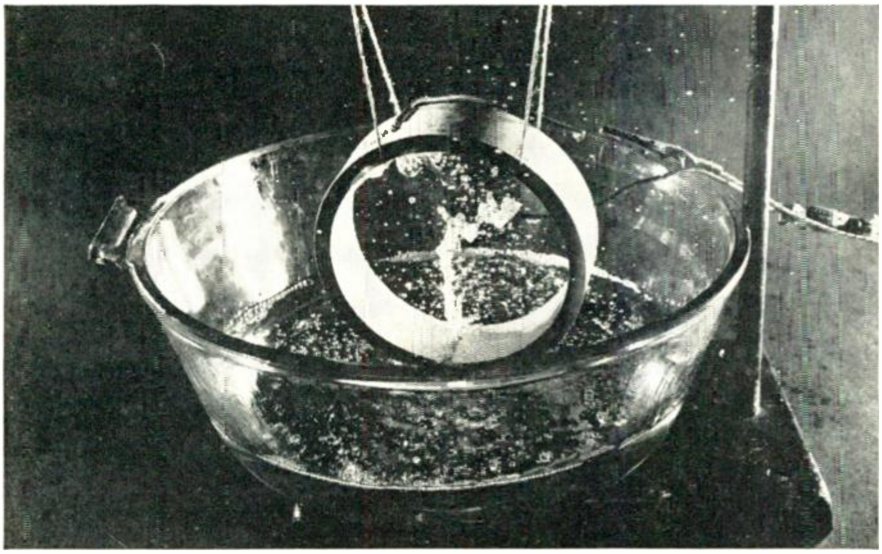
621.396:061.24 1534

The Seventh Plenary Assembly of the International Radio Consultative Committee—E. W. Allen. (PROC. IRE, vol. 43, pp. 132–139; February, 1955.) An outline is given of the organization and functions of the CCIR and of the discussions and recommendations of the Assembly held in London in 1953.

Gulton abstracts

6

MATERIALS RESEARCH • ELECTRONIC COMPONENTS • PRECISION INSTRUMENTS • SYSTEMS ENGINEERING



Wide field for industrial developments opened up by new ceramic transducer

Where would extraordinary simplification of the transfer from electrical to mechanical energy benefit an industrial or manufacturing project? Research and design ingenuity can undoubtedly come up with a multitude of possibilities. However, a potentially fruitful and exciting answer to this basic question lies with the appearance of a new series of piezoelectric ceramic transducers developed by Gulton Mfg. Corp. under the trademark FLEXDUCER.*

These transducers distort in bending, flexural, and torsional modes with four distinct advantages, namely (1) low, sonic, resonant frequencies from 20 cps to 10 kilocycles, (2) high displacement amplitude per applied voltage, (3) small size, and (4) only two electrodes required. The three basic configurations used are the split cylinder, helix, and spiral, but other forms can be fabricated as well.

Owing to the tremendous potential application of the FLEXDUCER, we invite experimental cooperation with industry and the military faced with design problems where these transducers can serve to distinct advantage. For instance, by using such transducers in resonance, low frequency filters and oscillators could be built.

Then, has a sonic pump been considered using the split cylinder transducer which avoids all moving and rotating parts in a completely enclosed circulating system, and eliminating all leaks and contamination? The photo illustrates the basic principle by showing a fountain produced by a FLEXDUCER ring operating at 60 cps.

Would the advantage of a low-impedance transducer appeal to the manufacturer of loudspeakers who might wish to eliminate the need of matching transformers between the amplifier output and the loudspeaker diaphragm?

It is also conceivable that the electric motor might be eliminated in an electric shaver by using the right transducer attached directly to the cutting blades.

Helix and spiral transducers which produce enormous displacement for a given DC voltage might be useful in the design of relays.

Another useful application is foreseen in DC-AC choppers where DC is converted to AC for voltage change in transformers.

Gulton Mfg. Corp. welcomes the opportunity to work with interested organizations or individuals to provide them with experimental FLEXDUCER units and production FLEXDUCER elements intended as vital components in completely new products for the consumer, other industries, or the military.

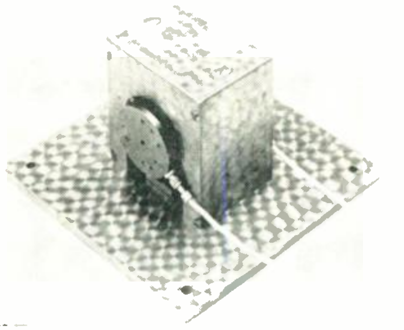
*Patent applied for

New high frequency vibrator designed for tube testing and accelerometer calibration

Tube manufacturers must determine vibrational resonances in electron tubes. Shock and vibration experts now require accelerometer calibrations up to 10 kc. Both will be interested in the new Glennite piezoelectric vibrator (Model AT-10A) which has been specifically developed for the generation of high frequency mechanical vibration in small structures and devices (pictured below).

Utilizing a rigid stack of Glennite piezoelectric transducers, this device has a fundamental resonant frequency of 27 kc. It is the balanced inertia type, since the center of the vibrating structure does not move. Compliant support of the vibrating stack at this node completely isolates vibrations from the housing.

Resonant-free up to 27 kc, the AT-10A has been found extremely useful in calibrating accelerometers up to 10 kc, and can also be used to determine vibrational resonances in small devices, such as electron tubes. The stiff electromechanical design of this instrument minimizes the effect of loading masses, and the vibrator will handle masses up to 3 ounces. This unit is provided with titanium end plates for mounting various types of devices.



Characteristics: resonant frequency 27 kc; sensitivity 2.3 micro-inches per 55 volts input; acceleration range up to ½ g at 1 kc and 50 g at 10 kc; frequency range 1 kc to 10 kc; recommended load up to 3 ounces; size 4" x 6" x 4"; calibration—interferometer calibration furnished up to 9 kc, internal standard accelerometers furnished.

For more information about Vibrator Model AT-10A, write on your letterhead to

GULTON MFG. CORP.
201 DURHAM AVENUE,
METUCHEN, NEW JERSEY

Published by **GULTON INDUSTRIES, INC.**

Comprising: Gulton Mfg. Corp. • Glenco Corporation • Vibro-Ceramics Corporation • Greibach Instruments Corporation • Thermistor Corporation of America

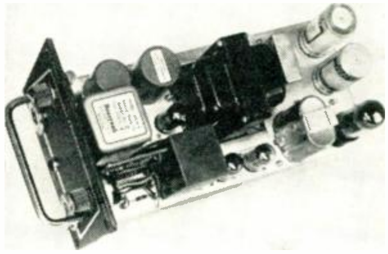
News-New Products

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

(Continued from page 94A)

Proportional Amplifier

A new selective range dc proportional amplifier, designed for use as a preamplifier in recording measurement systems or as an amplifier in highly sensitive control systems, was recently announced by **Doelcam, a division of Minneapolis-Honeywell Regulator Company,** Soldiers Field Rd., Boston 35, Mass.



A feature of the new instrument (known as Model 2HLA-4) is its use of the Doelcam second-harmonic magnetic converter as the input modulator. This converter avoids the inherent limitations of the usual vibrator-type converter and makes possible linear output, high sensitivity and low noise level.

Other features of the instrument include extended frequency response, interchangeable plug-in range units and optional isolated input.

The optional isolated input is useful in temperature measurements using thermocouples or thermistors. The plug-in range unit provides the desired combination of gain and bandwidth which is optimum for any given application. This range plug-in unit can be conveniently interchanged for a different combination of gain and bandwidth by merely replacing one plug-in unit by another.

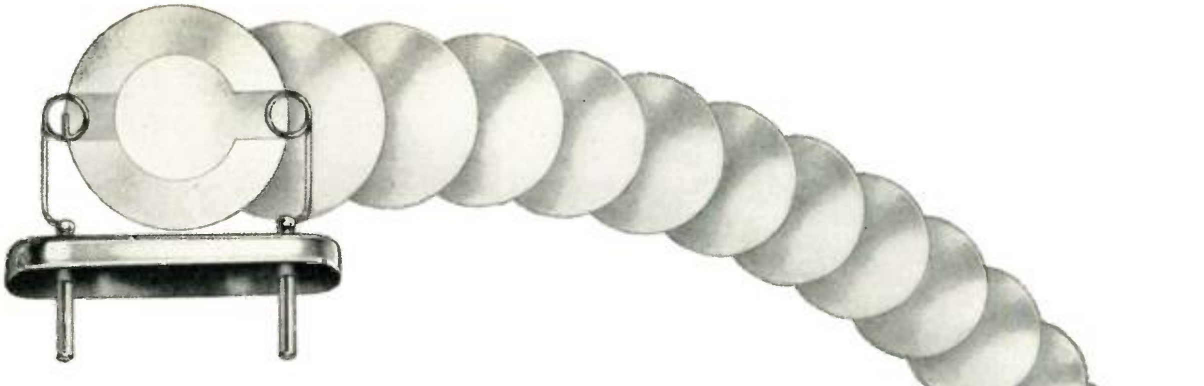
Full-scale voltage ranges as low as 100 microvolts may be selected, and a gain as high as 10^5 may be provided. The noise level is less than 5 microvolts and long term drift is less than 10 microvolts. Linear response within ± 1 per cent may be extended from dc to as high as 80 cps. Overrange inputs up to 1,500 times full-scale input have no adverse effects.

The Model 2HLA-4 is a portable instrument, weighing 20 pounds and measuring $6\frac{7}{8}$ wide by $6\frac{1}{8}$ high by 14 inches deep. It can be bench or rack mounted.

Transducer

Technology Instrument Corp., 535 Main Street, Acton, Mass., announces a dual element air flow differential pressure transducer. This small instrument is specifically designed for high output to speed-altitude-time computers, telemeter

(Continued on page 100A)



MILLIONS of crystals made to ANY specifications but only ONE standard quality

Midland frequency control units are on the job in two-way communications on land, sea and in the air throughout the world. Now they're playing a leading role in color television. The range of applications Midland serves is wide, but every Midland crystal has one thing in common· a single level of quality.

That one quality is simply the highest that modern methods and machines can produce. It's assured by Midland's system of critical quality control—exacting inspection and test procedures through every step of processing.

Result: Your Midland crystal is going to give you the best possible service in frequency control—with stability, accuracy, and uniformity you can stake your life on . . . as our men in the armed forces and law enforcement do every day.

Whatever your Crystal need, conventional or highly specialized When it has to be exactly right, contact

Midland

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3155 Fiberglas Road, Kansas City, Kansas

WORLD'S LARGEST PRODUCER OF QUARTZ CRYSTALS

News-New Products

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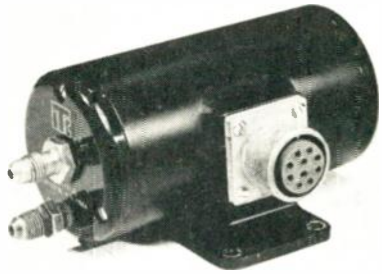
(Continued from page 98A)

systems, electric recorders, servo systems, meter display or alarms.

The pressure response movements of each of two metal bellows are translated into motion of sliders of paired precision potentiometers. One bellows is connected



to the dynamic pick-up line from Pitot or Venturi; the other bellows is exposed to pressure from the pick-up line sampling the static or ambient pressure. The electrical outputs of each of the instrument's four potentiometers are individually brought out to terminals, providing individual access to dynamic or to static data. Through an external electrical combining network, dynamic data may be automatically "normalized" or corrected with respect to static data as required by the application, limited only by the designer's ingenuity.

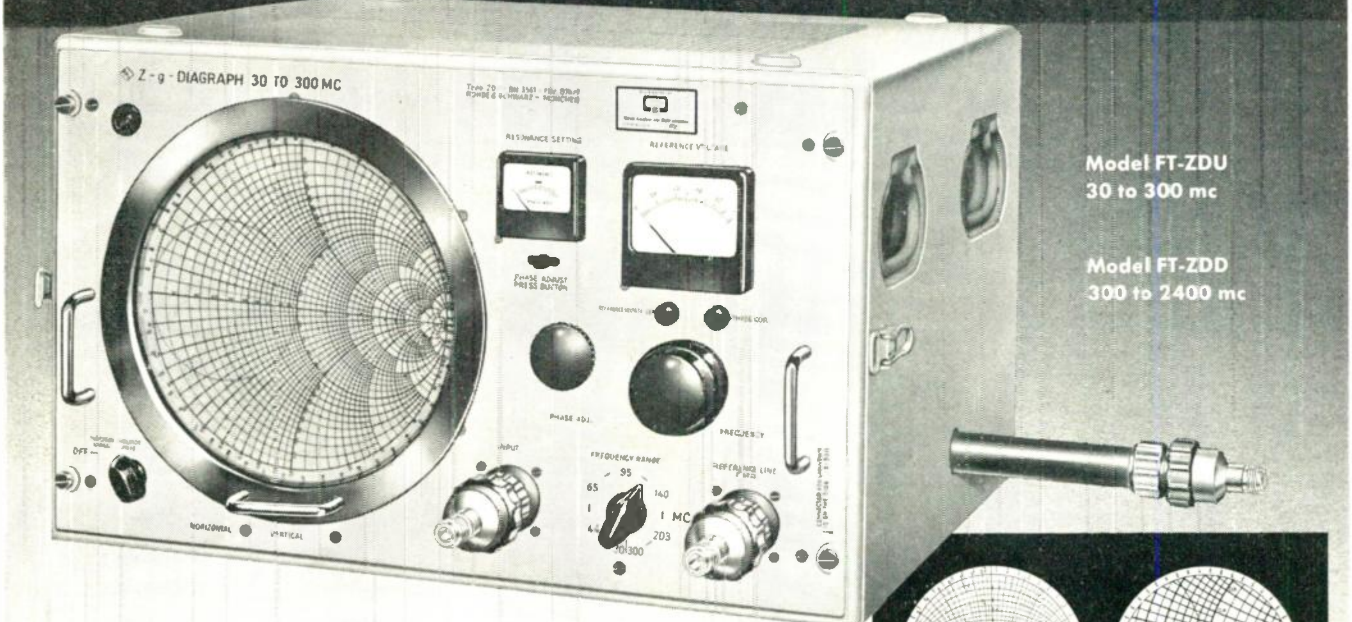
Typical application to speed-altitude measurement provides a range from 0 to 580 knots over an altitude range of −1,000 to 20,000 feet. Potentiometer resistance ratio vs. pressures is linear within 0.5 per cent; hysteresis within 0.5 per cent; change with temperature (−55°C to +71°C) not over 0.5 per cent; resolution increment less than 0.2 per cent of pressure range. Construction withstands 40g shock; departure from static calibration not over 0.7 per cent under 10g vibration up to 200 cps. Size $2\frac{7}{8} \times 4\ 3/88 \times 5\frac{7}{8}$; weight 1.3 lbs.

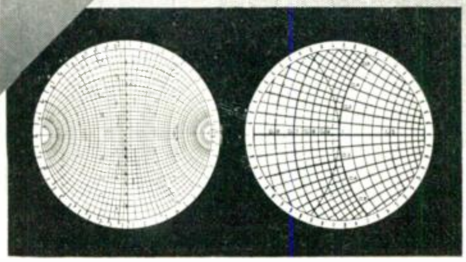
Test Equipment Library

Carl L. Frederick and Associates, 4630 Montgomery Ave., Bethesda 14, Md., is now making available, for the first time commercially, descriptive data on electronics test equipments. This comprehensive compilation (3 volumes; 2,300 pages; approximately 900 test equipments) is the result of a research project originated and monitored by the Wright Air Development Center of the United States Air Force's Air Research and Development Command. The great need for a single reference source on such data with resultant savings to manufacturers was considered to be of such sufficient importance for the Air Force to authorize the contractor to publish and sell these volumes commercially. Further details upon request to firm.

(Continued on page 102A)

DIAGRAPH . . . *for Fast, Accurate* IMPEDANCE *and* ADMITTANCE *Measurements*





Direct reading on interchangeable charts

Combining speed and accuracy, the Diagraphs measure complex impedance and admittance . . . with results presented directly in graphic form. Five different interchangeable charts provide permanent records . . . for files or blueprint reproduction.

Within minutes, complete impedance *vs.* frequency characteristics are obtained. Simplicity of operation facilitates use of the Diagraphs by personnel unskilled in high-frequency techniques. Set-up and operation are reduced to routine adjustments through the use of a self-contained calibration system. Transmission characteristics of four-terminal networks and phase angles between voltages are also measured and presented graphically . . . without accessories.

High speed, accuracy, and elimination of calculations are the features which have made the Diagraphs so well accepted wherever high-frequency measurements are performed regularly . . . wherever reduction of costs and conserving of scarce engineering manpower are important factors.

FEATURES:

- Direct reading on standard charts prevents errors and saves valuable engineering time.
- Rapid measurement of impedance and admittance of antennas, components, coaxial systems, cables, transistors, filters and attenuators.
- Rapid measurement of transmission characteristics of coaxial systems, cables, filters and attenuators.
- Measures phase angle between R-F voltages
- Self-calibrating—no external reference standards required.
- Easily operated by untrained personnel.

SPECIFICATIONS:

Characteristic Impedance: 50 ohms.
Measuring Range: Impedance . . . 1 to 2500 ohms; Phase . . . 0 to 360°; Attenuation . . . 0 to 30 db.
Accuracy Amplitude . . . ±3%; Phase . . . ±1.5°.

Terminals: Type N.
Power Supply: 115 volts (or 220 volts), 50 to 60 cycles.
Dimensions: 22" x 14" x 19"
Weight: 135 pounds.

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News-New Products

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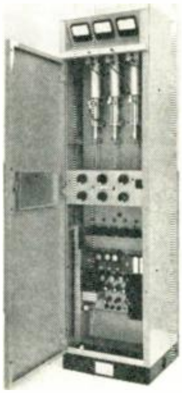
(Continued from page 100A)

High Voltage Dual Power Supply

A new Model D3-300D power supply which provides two high-voltage output ranges, each independent of the other, is being produced by **Dressen-Barnes Corp.,** 250 N. Vinedo Ave., Pasadena 8, Calif. It delivers 0–300 vdc at 300 ma maximum load, 0–300 vdc at 150 ma maximum load, dc variable bias 0–150 volts at 5 ma maximum load (regulated by VR tube), and 0–10 vac, unregulated, at 10 amperes maximum load which is variable with powerstat control. Regulation: for line voltage of 115 vac ±10 per cent, the output voltage change is 0.15 per cent. For dc regulated high voltage, the change from no load to full load is less than 30 mv. Ripple for both high voltage outputs is less than 500 microvolts rms. Both positive or negative of either high-voltage outputs may be grounded. All components in the unit are derated for long service life. Literature on Model D3-300D will be sent on request.

Satellite Transmitter

The **Adler Communications Laboratories,** 1 LeFevre Lane, New Rochelle, N. Y., has announced the shipment of one of its new products, a UST-20 television satellite transmitter, to Manson, Washington. This unit will be used in connection with the experimental Channel 16 satellite station. VHF signals on Channel 4 will be picked up, translated to Channel 16 and then re-radiated into the Manson area after amplification in the UST-20.



This ACL Satellite Transmitter consists of a wideband linear amplifier with a

(Continued on page 104A)

A precision jig is used to check angular alignment of electron guns. This is only one of many inspections to insure absolute accuracy of Du Mont tubes.

CATHODE-RAY TUBES

for precision instrumentation

Du Mont's business is cathode-ray equipment. Only Du Mont, with an intimate background in equipment design and manufacture, can build and recommend the **right** cathode-ray tube to the equipment designer. Through years of consultation on all kinds of tube problems, Du Mont, working with the equipment designer, has been able to lead the way with such developments as the flat face-plate, tight-tolerance glass-rod construction, mono-accelerator design, and many others. Development which leads the way toward the ultimate use of cathode-ray tubes as sensitive, precision measuring devices is our aim. Consult Du Mont for precision tubes to meet your exacting requirements. The best costs no more than the ordinary.

IMPORTANT SPECIFICATIONS OF PRECISION
Cathode-ray Tubes

TYPE	DEFLECTION FACTOR*		DEFLECTION UNIFORMITY	LINE WIDTH*	ANGULAR ALIGNMENT BETWEEN PLATES	P1 LIGHT OUTPUT*
	D1 D2	D3 D4				
3JP-A	150 dcv/in	111 dcv/in	3%	.03" max.	90° ± 1°	25 ft. L. Min.
3WP-	69 dcv/in	47.5 dcv/in	2%	.026" max.	90° ± 1°	7 ft. L. Min.
5ADP-	45 dcv/in	35 dcv/in	2%	.03" max.	90° ± 1°	15 ft. L. Min.
5AMP-	45 dcv/in	22.5 dcv/in	1%	.032" max.	90° ± 1°	15 ft. L. Min.
5AQP-	45 dcv/in	35 dcv/in	1%	.030" max.	90° ± 1°	15 ft. L. Min.
5ARP-	dual beam mono-accelerator — each gun equivalent to Type 5AQP — (see above)					
5ATP-	104 dcv/in	38 dcv/in	1%	.035" max.	90° ± 1°	170 ft. L. Min.

*Under typical operating conditions (7-inch versions of the five inch tubes will be considered on request.)

DU MONT

For further information write to: **TECHNICAL SALES DEPT.**
ALLEN B. DU MONT LABS., INC., 760 BLOOMFIELD AVENUE, CLIFTON, NEW JERSEY

News-New Products

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(Continued from page 102A)

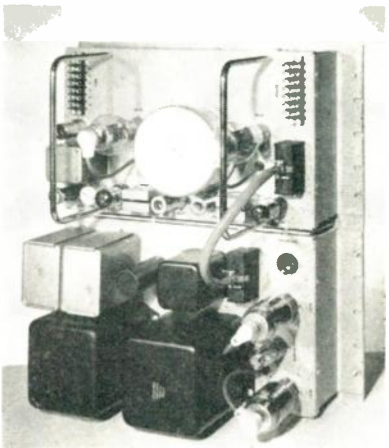
6 mc bandwidth and a gain of 80 db. It operates from a UHF TV signal converted from VHF and delivers 20 watts of peak visual power to a 50 ohm load. Both aural and visual carriers are amplified simultaneously in the same system.

This UST-20 satellite transmitter is identical electrically to the satellite/booster transmitter (KC2XFF) now in experimental operation at station WATR-TV in Waterbury, Connecticut.

The new UST-20 is part of ACL's complete line of low power TV equipment for both UHF and VHF. The UST-150 amplifier operates in conjunction with this unit to produce 150 watts of visual output. However, when used with special ACL UHF antenna systems, up to 5,000 watts ERP is available.

Power Amplifier

Gotham Audio Development Corp., 2 West 46th Street, New York, N. Y. announces development of a new 150-watt power amplifier unit adaptable to many applications.



The PFB-150WD power amplifier is designed for operations requiring high power output over a wide frequency range and at minimum distortion. The unit is suited to laboratory installations, public address systems, stereophonic speaker systems for motion picture projection, program distribution, etc.

The PFB-150WD delivers 150 watts with less than 7/10 per cent RMS harmonic distortion from 40 to 15,000 cps, and 10 watts with less than 0.15 per cent over the range 20–20,000 cps. IM distortion is held below 1 per cent at rated power output, while the damping factor exceeds 10. Noise level is 84 db below 150 watts output. Instantaneous peak power of 400 watts is available with perfect stability. Output ranges from 38 volts at 4 amperes to 125 volts at 1.2 amps.

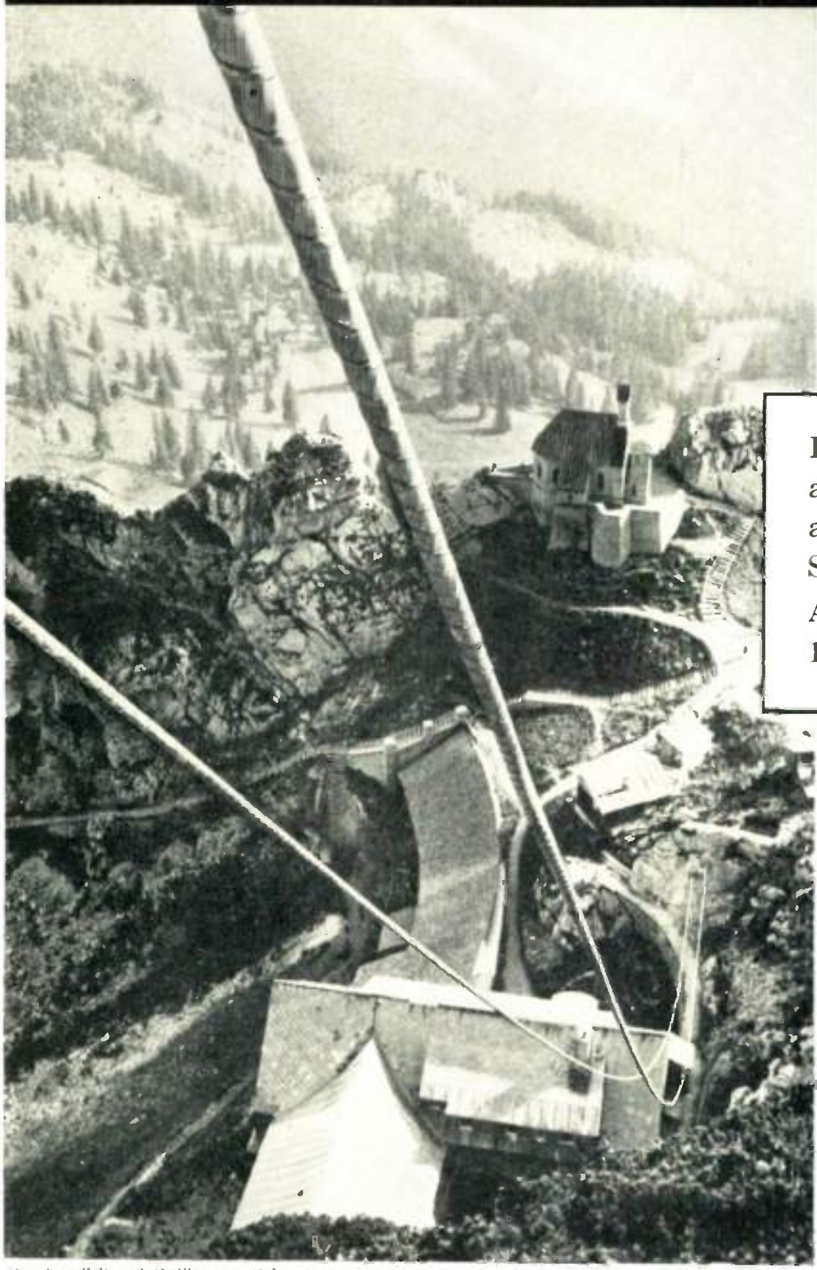
Balanced and unbalanced inputs of from 150 to 10,000 ohms are available.

(Continued on page 106A)

Styroflex Coaxial Cable

FIRST CHOICE FOR

DIFFICULT MOUNTAINOUS CONDITIONS!



Courtesy Felten & Guilleaume, Cologne

After 5 Years of Rigorous Service in the Swiss Alps, the Electrical Characteristics of Styroflex are Unchanged!

FM Installation of Styroflex Cable armored for 25-ton ice load at the Wendelstein Alp near the Swiss border.

Aluminum sheath approximately 1 1/8 inches OD.

Ideal for Remoting Video Circuits for both Television and Radar!

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(Continued from page 104A)

while output impedances range from 8 to 93 ohms. A gain control adjustable in 2 db steps and direct output tube plate metering are provided. Amplifier consists of 6 push-pull stages balanced through use of 1 per cent precision resistors, and features 811-A's as output tubes biased positively for minimum distortion. Output transformer is toroidally wound. 3B28's are used as rectifiers with choke input filter and oil filter capacitors. Warm-up time delay is accomplished through use of a thermistor and thermal delay relay combination.

Unit mounts in standard relay rack, occupying 21 inches of vertical panel space. Weight of both power supply and amplifier together is 95 pounds.

Color Bar Dot Generator

The ChromaDot, a new combined color bar dot generator with vertical sync and requiring one connection to the rf antenna or video amplifier, has been announced by **Kay Electric Co.**, Pine Brook, N. J.

The color bar generator has a horizontal sync pulse signal. It provides a vertical sync pulse and varying pedestals throughout each frame to test linearity of color receiver circuits. The dot section provides signals which contain horizontal pulse, dot pulses, and vertical sync, to display a stationary locked pattern without auxiliary signals.

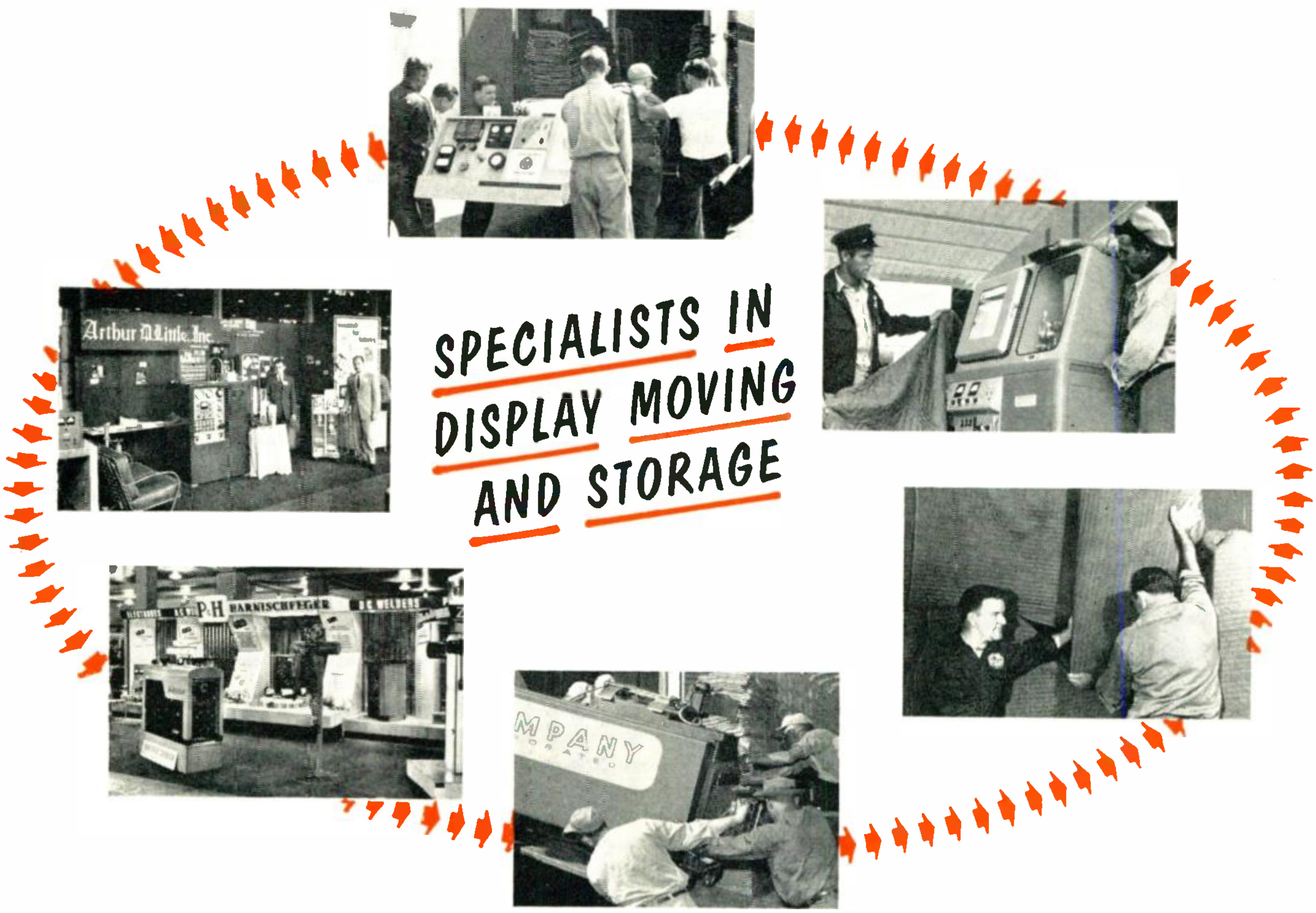
These signals are impressed as modulation on a picture carrier. A specific channel may be selected by customer. An associated sound carrier is provided for tuning.

The color bar pattern produces a display of 10 color bars with progressive 30° phase shifts from the color pulsed signal. This complete signal is suitable for the adjustment of receivers using I, Q, and B-Y, and R-Y Matrix Systems. The output of the color burst and sync pulse is adjustable to the industry standard.

Specifications are as follows: Video output −0.6 volts peak to peak into 75 ohms, 10 volts peak to peak into 5,000 ohms. Positive or negative polarity sync pulses are provided. RF output −0.2 volts into 75 or 300 ohms provided. Price: $395. f.o.b. plant. For complete information, write Kay Electric.

(Continued on page 108A)

Let North American Handle Your ELECTRONICS Exhibit!



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- Minimum Crating
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If you are shipping a display to the Western Electronics Show or the National Electronics Conference, do it the new easy way . . . call your local North American Van Lines agent. His door-to-door service in the same padded van eliminates drayage and minimizes costly crating. Supervision by NAVL's experienced display men at both origin and destination saves you time, reduces over-all exhibit costs. Only North American Van Lines offers this *special departmentalized* service!

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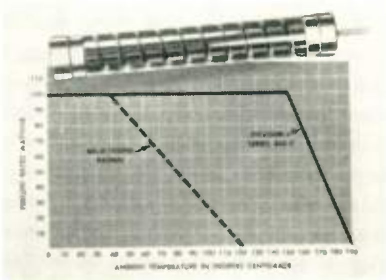
(Continued from page 106A)

Film Type Resistor

The Daven Co., 191 Central Ave., Newark 4, N. J., announces the development and availability of the new film type Davohm Series 850-T Resistor. The new resistors can be used at 150°C at full rated power. They derate linearly to zero power at 190°C.



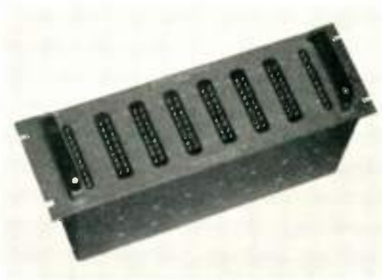
The temperature coefficient of these resistors is unique for a film type resistor. The temperature coefficient is below +400 PPM/°C, is always positive in value, and is independent of resistance value. Any ohmic value of these Series 850-T resistors will track within approximately ±20PPM/°C of the normal temperature coefficient value over the temperature range.

The new hermetically sealed series 850-T resistors offer excellent moisture resistance and load life stability. On a typical MIL cycle 1,000 hour load test of twelve 2 watt type 852-T, 200,000 ohm resistors, one had a maximum deviation of 0.2 per cent and the rest had maximum deviations of 0.1 per cent. The voltage coefficient is below 0.0005 per cent per volt.

The new Series 850-T resistors are available in ½ watt, 1 watt and 2 watts.

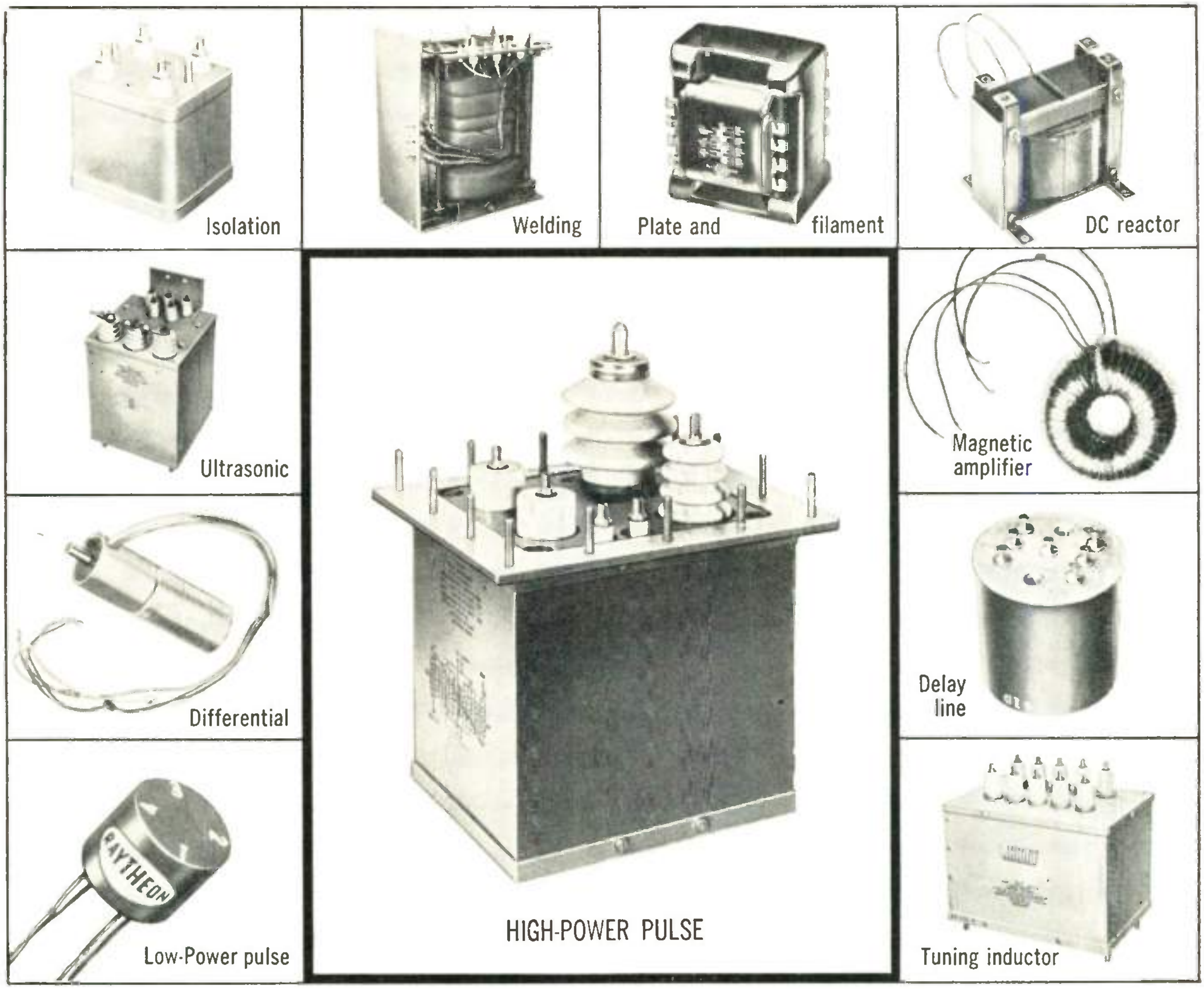
Precision Delay Line

Epsco, Inc., 588 Commonwealth Ave., Boston 15, Mass., has developed a new model DL 0510-400/125 Delay Line—a



precision, low attenuation unit, developed for correlation measurements and wave form analysis covering sub audio and audio frequencies. Over-all delay is 5,000 μs; characteristic impedance 510 ohms; taps are available every 40 μs.; calibration ac-

(Continued on page 110A)



RAYTHEON TRANSFORMERS

designed for your specialized applications

CUSTOM DESIGN

To meet your need for specialized electronic signal and power range transformers, Raytheon offers exceptional standard transformers and custom design facilities. An unusually large and widely experienced engineering staff is at your service to design and develop transformers that best fit your particular applications.

PERSONAL SUPERVISION

Available to you are the resources of Raytheon's entire transformer engineering staff. Yet in order to best satisfy your needs, design, development and production of your transformers are turned over to an individual Raytheon engineer who sees your job through from start to finish.

PRODUCTION AND TESTING

All types of winding, core processing, impregnation and baking equipment are available for model making or full production runs. Raytheon also offers complete facilities for testing.

25 YEARS' EXPERIENCE

Raytheon has successfully custom engineered over 30,000 transformer designs and millions have been produced. Proof of Raytheon quality is this fact: *in 25 years less than ¼ of one percent of all Raytheon transformers have been returned from the field for any reason.*

For full information write Department 6120.
Request catalog 4-100

RAYTHEON MANUFACTURING COMPANY

RAYTHEON

Equipment Marketing Division

Dept. 6120, Waltham 54, Mass.

Excellence in Electronics

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 108A)

curacy at each of 125 taps is ±0.1 μs.; insertion loss is 1.7 db; cut-off frequency 9 kc; phase linearity is ±1 per cent up to 5 kc. The unit size is 19×6×6 inches for relay rack mounting.

Blower Unit

Light weight, long life, miniature tube axial blower units powered by a new one-inch diameter motor have been produced by **Eastern Air Devices, Inc.**, 357 Central Ave., Dover, New Hampshire, for use in spot cooling of electronic equipment and for air changing in small, equipment-filled boxes.

These new tube axial blowers have no brushes, and operate without arcing and without radio interference. At speeds of 11,000 and 20,000 rpm standard 400-cycle blowers produce 32 and 58 cfm (at 0" of static pressure). Variable frequency 360 to 1,600 cycle units produce a minimum of about 25 cfm at 0" of static pressure. All EAD tube axial blowers meet applicable MIL specifications for aircraft use.

Typical EAD blowers measure two inches in diameter and have an over-all length of $3\frac{3}{8}$ inches.

Miniaturized Power Packs

Electronic Research Associates, Inc., 67 E. Centre St., Nutley, N. J., announces a new two-sided catalog sheet covering their line of Transpac miniaturized power packs. This new catalog includes many new models covering both 60 and 400 cps types as well as a new series of models which supply a combination of constant voltage and constant current specifically for transistor application. Other Transpac models include regulated low voltage types for reference and similar applications, constant voltage types for general tube and transistor applications, constant current types, unregulated low, medium, and high voltage models, and high amperage designs.

(Continued on page 114A)

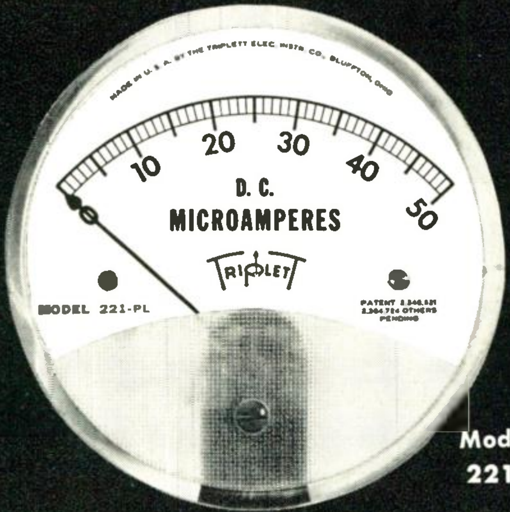
TRIPLETT IS WAY AHEAD IN

PANEL METERS featuring

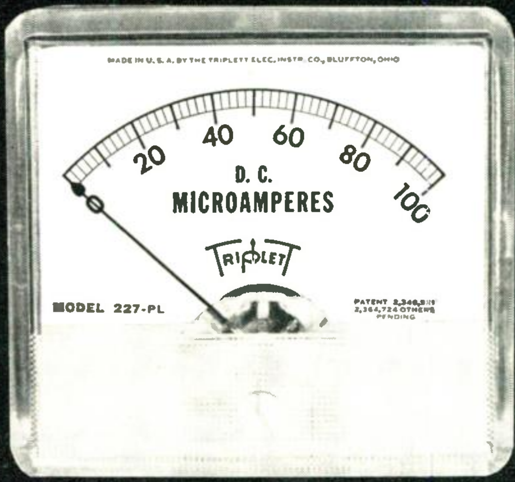
Clear Front Plastic PL Styles with:

1. Much longer scale
2. More light on scale
3. Clear, unbreakable front
4. Regular panel space and mounting
5. Improved panel appearance

Triplett PL meters are now available in 5 styles . . . 2" and 3" round, plus 2", 3" and 4" square. Transparent plastic case front projects over the rim of the instrument, offering longer scale length and easier readability. PL meters are available in D.C. Permanent Magnet Moving Coil and A.C. Iron Vane types.



Model 221-PL



Model 227-PL

TESTING EQUIPMENT

When selecting your next piece of test equipment, look over Triplett's *complete* V-O-M line . . . it will prove Triplett's ability to keep pace with your requirements. Below are examples of the complete V-O-M line. Look to Triplett for leadership.

TRIPLETT ELECTRICAL INSTRUMENT CO.
Bluffton, Ohio

MODEL 630. Famous Volt-Ohm-Milliammeter. Many features making this the favorite in a popular V-O-M line.

MODEL 630NA. The first V-O-M with all-in-one features. Does everything anybody ever wants. 70 ranges, plus other unique features.

MODEL 631. Combination V-O-M and VTVM. 2-in-1 battery operated combination saves money—will do all your work easier at half the price.

BURTON BROWNE ADVERTISING

TRIPLETT HAS SET THE STANDARD IN INSTRUMENTS FOR OVER HALF A CENTURY

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 110A)

White Dot-Cross Hatch-Bar Generator

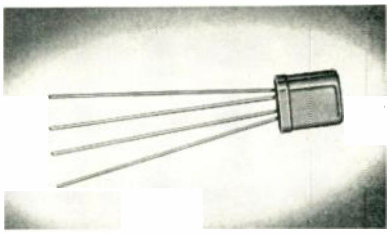
Electronic Measurements Corporation, 280 Lafayette St., New York 12, N. Y., has recently announced the development of the Model 800 signal generator for color and monochrome servicing.

The Model 800 produces white dots on the TV screens of any monochrome or color receiver for the adjustment of color convergency of the tri-gun color tube; it will also produce a cross hatch pattern and a variable number of bars for adjustment of width and height controls. The instrument can also be used for localizing and identifying trouble sections in both monochrome and color TV receivers.

The Model 800 comes completely wired and tested with all necessary leads for $39.90.

Junction Transistor

A general purpose of N-P-N grown junction transistor is now available in quantity from **Bogue Electric Mfg. Co.,** Dept. WNH-2, 52 Iowa Avenue, Paterson 3, N. J.



This transistor is intended for applications at power, audio and radio frequencies up to 1,000 kc. It is particularly adaptable for use in control and switching devices and is capable of power gains up to 32 decibels.

Since this Bogue transistor is hermetically sealed it is impervious to moisture and possesses good temperature characteristics.

(Continued on page 118A)

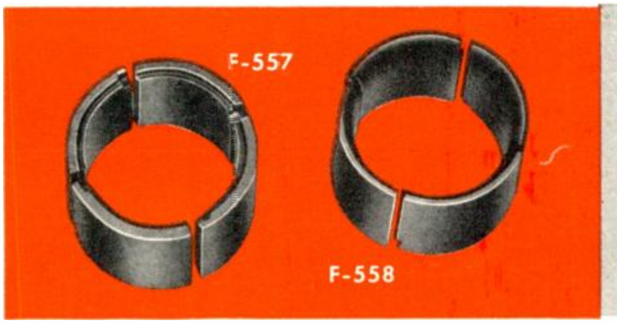
Specify these INDUSTRY PREFERRED TV Components

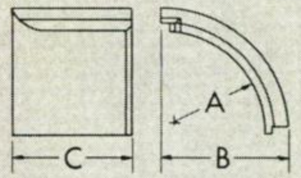
FERRAMIC®

FOR:

- Improved Performance
- Greater Economy
- Faster Delivery

YOKE and FLYBACK CORES by General Ceramics

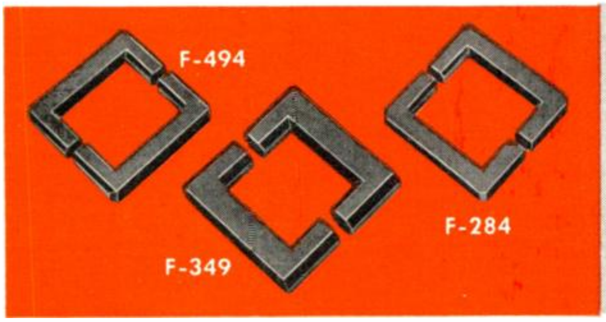


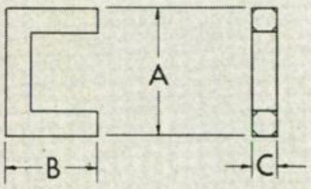


*F-557 conforms to MPA Tentative Standard 18-55T for parts DY-3, DY-4.

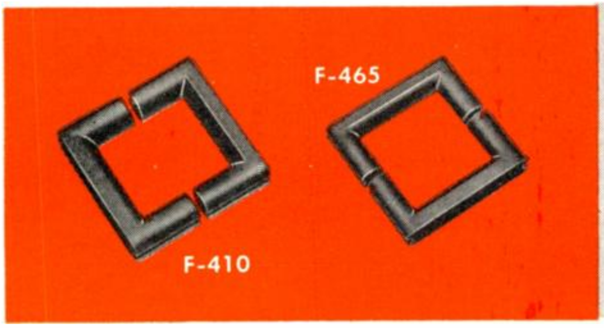
†F-558 conforms to MPA Tentative Standard 18-55T for parts DY-1, DY-2.

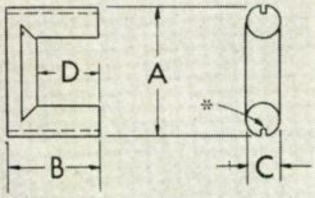
PART NO.	F-557*	F-558†
A	.960″	.950″
B	1.210″	1.215″
C	1.125″ 1.250″	1.250″ 1.375″





PART NO.	F-284	F-349	F-494
A	2.250″	2.500″	2.250″
B	1.057″	1.156″	1.025″
C	.375″	.490″	.375″
X-Sect.	Rect.	Octag.	Rect.





*Longitudinal Screw slots are .156″ deep x .140″ wide.

PART NO.	F-410	F-465
A	2.477″	2.308″
B	1.156″	1.062″
C	.545″	.447″
D	.703″	.687″

OFFERING:

- SOUND MECHANICAL DESIGNS
- HIGH EFFICIENCY
- EXCELLENT STABILITY
- LIGHTWEIGHT
- GREATER UNIFORMITY

Standard Ferramic parts offer a dependable, quick solution to TV design and production problems. These preferred types give you all the electrical advantages of Ferramics, plus faster delivery and maximum economy. Call or write now for complete data on preferred TV Components or Ferramics for other applications.

GENERAL CERAMICS CORPORATION · STEATITE INSULATORS

General CERAMICS CORPORATION

TELEPHONE: VALLEY 6-5100

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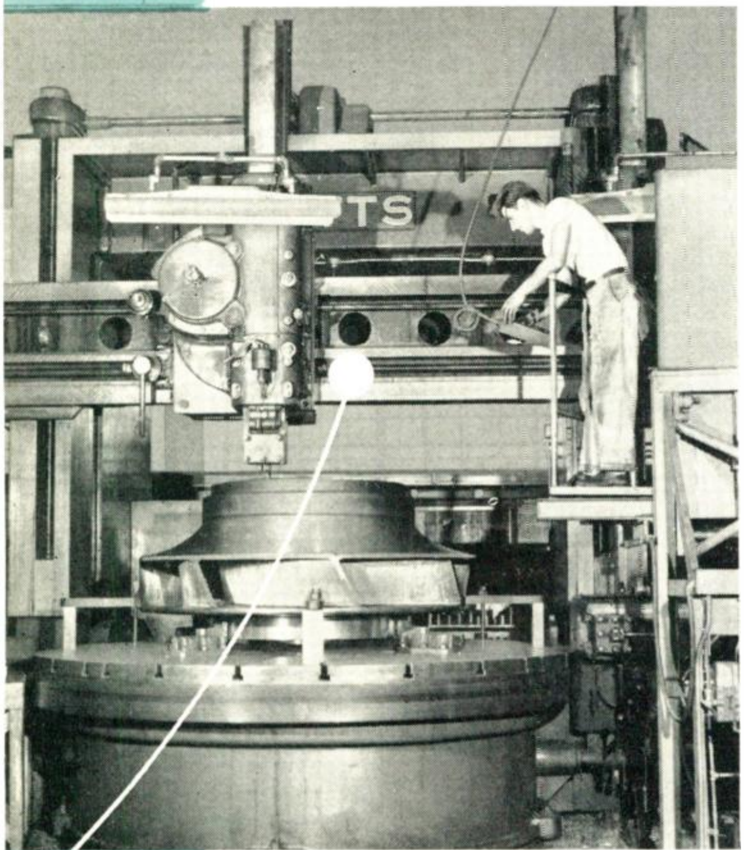
MAKERS OF STEATITE, ALUMINA, ZIRCON, PORCELAIN, SOLDERSEAL TERMINALS, "ADVAC" HIGH TEMPERATURE SEALS, CHEMICAL STONEWARE, IMPERVIOUS GRAPHITE, FERRAMIC MAGNETIC CORES

Byron Jackson Co. *and* The Rollin Co. combine

Now electronic ingenuity combined with heavy industrial facilities



Experienced electronic engineering . . . backed by diversified heavy industrial facilities . . . gives the Byron Jackson Co. Electronic Division an unusual ability to pioneer precision electronics in practical form.

The engineering nucleus of this Electronic Division has, since 1944, been engaged in the development and manufacture of high power signal generators, precision slotted lines, miniature digital type transducers, and other precision electronic equipment. Now these electronic skills are teamed with the metallurgical, hydraulic, mechanical, chemical, nuclear, electrical and petroleum specialists and scientists of Byron Jackson Co. In addition, a multi-million dollar investment in plants and production equipment supports our specialized electronic manufacturing and testing facilities.

This unique combination of engineering talent and production abilities is ready to work for you in the field of measuring and testing instruments, complete systems, or an entire development production project.

BJ SINCE 1872

Byron Jackson Co.

ELECTRONIC DIVISION

2010 LINCOLN AVENUE, PASADENA 3, CALIFORNIA • TELEPHONE: RYAN 1-7134

News-New Products

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

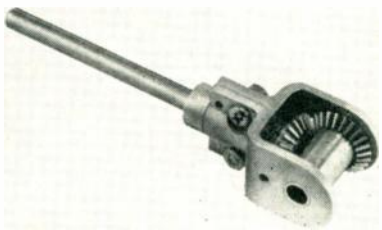
(Continued from page 114A)

Silicon Junction Diodes

A new line of silicon junction diodes has just been announced to the industry. Manufactured by the **Semiconductor Div., Hughes Aircraft Co.,** Florence Ave. at Teale St., Culver City, Calif., the devices are described as having high forward conductance combined with extremely high back resistance. In several of the diode types, this resistance is in the order of 10,000 megohms. In many applications, this is essentially an open circuit. According to the Hughes research engineers, this back resistance makes it possible to use these silicon diodes in many entirely new circuit applications, in addition to those now closed to germanium through temperature limitations.

These diodes include the previously announced Types HD6001, HD6002, and HD6003, plus new Types HD6005 through HD6009. All of these types have an ambient operating temperature range of from −80 to 200°C. Physical features: fusion-sealed in a one-piece glass body, impervious to moisture; flexible dumet leads, tinned for soldering or spot-welding; diode envelope externally coated with black silicone enamel to shield crystal from light. Dimensions, glass body: 0.265 by 0.103 inch, maximum. Specification sheets available upon request.

Right Angle Drive



The National Company, Inc., 61 Sherman St., Malden 48, Mass., has announced production and distribution of a new addition to the National line of precision components. It is a right angle drive, model RAD. This drive features unusually sturdy construction with die cast zinc housing and gears. It is ideal for ganging condensers or potentiometers or other parts located in hard-to-reach locations on a chassis. Total length including shaft is 4½ inches. Shaft is standard ¼ inch diameter. Action is smooth and free from backlash.

(Continued on page 120A)

TUNG-SOL "Magic Mirror" ALUMINIZED PICTURE TUBE

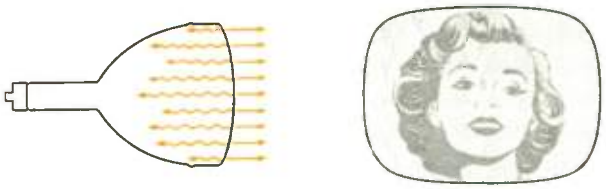
BRIGHTER-SHARPER
MORE DETAIL
MORE CONTRAST

The "Magic-Mirror" Aluminized Picture Tube produces the brightest, most realistic picture ever seen in the American home. The "Magic-Mirror" tube effectively utilizes *all* the light generated by the phosphor screen.

Tung-Sol has developed a unique "fogging" method of backing up the phosphor screen with a mirror-like aluminum reflector. This reflector prevents light radiating uselessly back into the tube. It brings out all the detail of which the receiver circuit is capable. So smooth and true is the Tung-Sol aluminum reflector that mottling, streaks, swirls, "blue-edge", "yellow center" and other objectionable irregularities are eliminated.

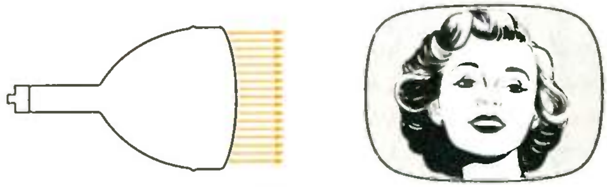
Tung-Sol pin-point-focused electron gun assures a steady, brilliant picture—free from alternate fading and overlighting. Tung-Sol's exacting standards of quality control, manufacture and testing further guarantee the high uniformity and maximum performance of the "Magic-Mirror" TV Picture Tube.

Let the superior qualities of "Magic-Mirror" Picture Tubes add selling advantages to your set.



ORDINARY TUBE—Only *half* the light produced by the phosphor screen is utilized in the picture. Other half radiates wastefully back into tube.

RESULT—A light background within the tube which reduces picture contrast.



MAGIC-MIRROR ALUMINIZED TUBE — Aluminized reflector allows electron beam through. Blocks wasted light from backing up into tube. Reflects *all* the light into picture.

RESULT—Pronounced increase in contrast to make a bright, clear, more realistic picture.

TUNG-SOL ELECTRIC INC., Newark 4, N. J.

Sales Offices: Atlanta, Chicago, Columbus, Culver City (Los Angeles), Dallas, Denver, Detroit, Montreal (Canada), Newark, Seattle.

Tung-Sol makes All-Glass Sealed Beam Lamps, Miniature Lamps, Signal Flashers, Aluminized Picture Tubes, Radio, TV and Special Purpose Electron Tubes and Semiconductor Products.

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 118A)

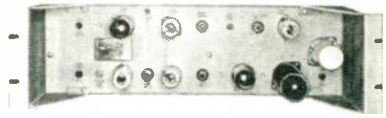
Pease Appointed by Feedback Controls

William M. Pease has recently joined the staff of **Feedback Controls, Inc.**, 1337 N. Henry St., Alexandria, Va., as Vice-President and General Manager.

Pease was formerly Director of the Servomechanisms Laboratory at M.I.T. and more recently Vice-President and Manager of the Electronics Division of Ultrasonics.

Mr. Pease was instrumental in the development of the M.I.T. numerically controlled Milling Machine.

Sync Generator



Standard Sync Generating Equipment is large, expensive, and generally not portable. This condition has complicated the routine checkout of video facilities, especially from remote locations.

The new, small, portable, and inexpensive Model 302-AR Drive Generator developed by **Telechrome, Inc.**, 632 Merrick Road, Amityville, L. I., N. Y., provides blanking horizontal sync, vertical drive, and color burst flag for driving most color and monochrome signal generating equipment where standard sync is not available. The Model 302-AR Drive Generator provides driving pulses for signal generating equipment, such as Multi-Burst, Window, Stairstep, and Linearity Checkers from remote check points, mobile units, and transmitters where studio sync is not available. It is a non-standard generator for test purposes and other functions where an FCC signal is not required.

News-New Products

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

Tube Tester

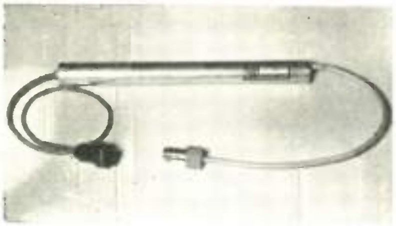
Precise Development Corp., Oceanside, L. I., N. Y., has available in wired form a unit which checks both emission and mutual conductance.

Priced in kit form at $69.95 and factory-wired at $139.95, it will enable the user to check all tubes, including hearing aid, miniatures and CRT's.

In addition, it has new type switches: voltage sapper check; gas check; simplified short check and the latest roll chart. The weight is 30 pounds and the size is 14 by 16 by 6 inches.

X-Band Backward Wave Oscillator

This traveling wave tube developed by **Huggins Laboratories, Inc.**, 711 Hamilton Ave., Menlo Park, Calif., features broadband electronically tuned oscillations from 7 to 14 kmc. This is accomplished by the adjustment of a single voltage without any complementary mechanical adjustments.



This type of oscillator would find its greatest field of use in swept signal generators for automatic testing, swept local oscillators, and wide band transmitter-receiver applications. The tube is capable of being swept across this band in less than one microsecond.

The approximate operating characteristics over this band are 300 to 3,000 volts, 12 ma current, and 10 mw output. X-Band voltage requirements are 450 to 2,000 volts.

(Continued on page 122A)

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 121A)

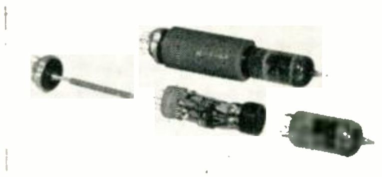
Vitramon Names Cornell Design Chief

Edward S. Cornell was recently appointed by **Vitramon, Inc.,** P.O. Box 544, Bridgeport, Conn. to head up the company's process-design department.

Mr. Cornell brings to Vitramon an outstanding record in product, machine, and tool development. Of special importance to Vitramon is Cornell's experience during World War II, when he was called to Washington as a chief production specialist to make a survey on the production of capacitors.

Embedded Circuit

Alcor Electronics Corp., 180 Lafayette St., New York 13, N. Y., introduces the new Lock-In and Plug-In Encapsor, a durable plastic embedded electronic circuit arranged in the form of a small cylindrical cartridge having means to accept a vacuum tube and which plugs into and is securely locked to a tube socket.



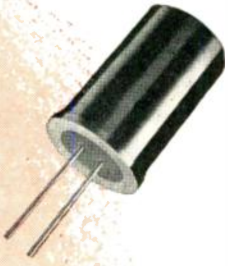
Encapsors are used in place of conventional methods of assembly and wiring components associated with vacuum tubes.

Made of Alcorite plastic, gold plated receptacle contacts, plug pins of phosphor bronze, A-B type resistors, Mylar capacitors, silicon diodes, and other high quality components, it is available for both standard and printed circuits application.

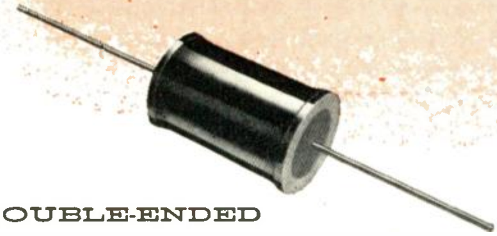
Literature and schematics are available upon request.

(Continued on page 124A)

For *maximum flexibility* in your equipment design and layout



SINGLE-ENDED
with leads
clipped for
socket mounting



DOUBLE-ENDED
for soldered-in
applications

SINGLE-ENDED
for conventional
soldered-in use

Sylvania offers you a choice in Silicon Junction Diodes

featuring

- —high back resistance
- —high rectification ratios
- —sharp breakdown in saturation voltage region
- —ambient operating temperature to 150°C.

Chassis layout call for conventional soldering-in of diodes? Sylvania offers double-ended package design. Or you can choose the single-ended diode package which can be socket-mounted as well as soldered-in. Broad-range diodes and narrow-range diodes are available in both packages.

Broad-Range Diodes—back resistance of 1000 to 10,000 megohms. Low capacitance and moderate forward conductance make them highly useful in general and computer applications as rectifiers, end gates, restorers, and clippers.

Single-ended	Double-ended
IN137	IN409
IN138A	IN410

Narrow-Range Diodes—controlled breakdown voltages from 2 to 600 volts. These general-purpose diodes feature superior rectification ratios and an ability to function at elevated ambient temperatures. They have a very low impedance in the saturation voltage region. Low-voltage types make good voltage regulators for transistor power supplies. All types are applicable in magnetic amplifiers.

Single-ended	Double-ended
IN200 to IN222	IN370 to IN400
IN401 to IN408	

For complete data on all Sylvania Junction Diodes, address Department F32R.

SYLVANIA®

SYLVANIA ELECTRIC PRODUCTS INC.
1740 Broadway, New York 19, N. Y.
In Canada: Sylvania Electric (Canada) Ltd., University Tower Building, Montreal

LIGHTING • RADIO • ELECTRONICS • TELEVISION • ATOMIC ENERGY

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 122A*)

Pulse Generator

Teletronics Laboratory, Inc., 54 Kinkel St., Westbury, L. I., N. Y., announces the availability of a new Pulse Generator and Calibrator Model PC-100A.

The Model PC-100A produces two rectangular pulses having independently controlled amplitudes and polarities; polarities are positive or negative and amplitudes are 0–75 volts, open circuit from a 220 ohm source. Their repetition frequencies are adjustable at 50–5,000 pps, and paired pulse interval is adjustable at 5–5,000 μs. Time durations are 1 μs for sliding-pulse output, and 1 μs for fixed-pulse output synchronized from external source and 1.5 μs when operated self-synchronous. Time markers are provided at 1, 10, and 100 μs, for amplitudes of 0–10 volts, open circuit. Square-wave calibrator output is 60 cps, 0.1 to 100 volts.

Multimeter

Physics Research Laboratories, Inc., 507 Hempstead Turnpike, West Hempstead, L. I., N. Y., is making available the complete line of C. P. Goerz (Vienna) precision electrical instruments. The Universal "HV" is a typical Goerz multimeter which measures voltage, current, resistance and capacitance. A high voltage probe permits measurements of dc voltage to 15,000 volts. Decibel scale for attenuation included. Accessories extend current and voltage ranges beyond scale values. Current transformer available for conversion of meter to ac ammeter.

(*Continued on page 128A*)

Improve quality, cut costs of high-volume TV output with G.E.'s new 600-Series Tubes 6BH8 and 6CN7!

General Electric, originator of 600-Series Tubes for series-string circuits, announces the new 6BH8 and 6CN7, which bring to 50 its line of uniform-warm-up types. Use these new G-E miniatures for higher TV quality—you will have fewer production-line rejects, more reliable performance in set-owners' homes!

Use them for lower TV costs! Each of the new tubes will replace two single-purpose types—is in itself a low-price tube. Use them to reduce plant tube inventory . . . the 6BH8 and 6CN7 can be applied in either series-string or transformer sets!

Use the new G-E tubes for design versatility! The triode sections of both types can do a number of different circuit jobs.

Wire or write for description, ratings, and prices! *Tube Department, General Electric Company, Schenectady 5, New York.*

NEW G-E 6BH8 . . . Medium-mu triode-pentode, for use as general-purpose triode, and video amplifier or i-f amplifier. Has 600-ma heavy-duty heater with series-string warm-up time.

NEW G-E 6CN7... Duplex-diode high-mu triode, with a separate cathode for the two diodes. Triode section is useful in many circuit applications. The tube has a 600-ma heavy-duty heater with series-string warm-up time.

NOW 50 G-E 600-SERIES receiving tubes with uniform series-string warm-up time. And *all* G-E picture tubes have heaters with series-string warm-up! For top performance, specify G-E tubes throughout your circuit!

Progress Is Our Most Important Product

GENERAL ELECTRIC

162 IA3

DIRECT

ONE WAY

STREET TO THE RADIO-ELECTRONICS MARKET!

ONE WAY

IRE DIRECTORY ALONE PROVIDES PRODUCT FACTS ENGINEERS NEED!

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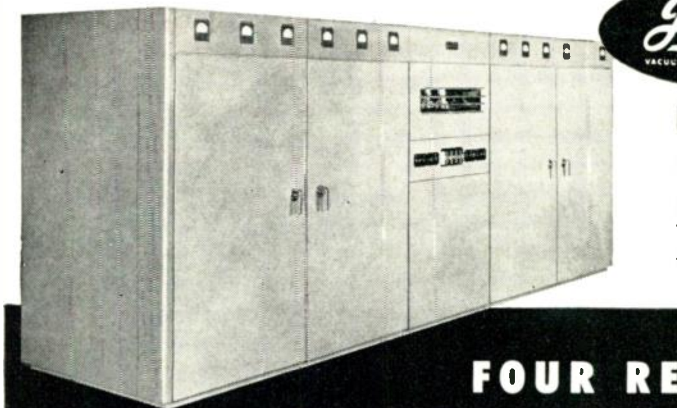
R E

The Institute of Radio Engineers
Advertising Department
1475 Broadway, New York 36, N. Y.

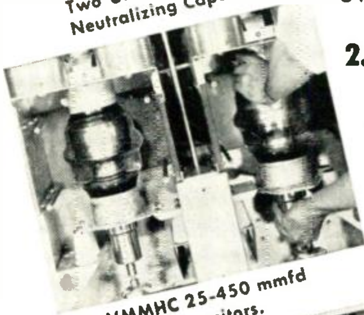
Jennings RADIO
VACUUM ELECTRONIC COMPONENTS



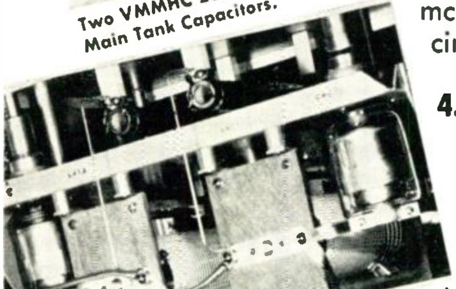
GATES 20 kw, 2 to 22 mc, Short Wave Broadcast Transmitter for Indonesia

FOUR REASONS WHY GATES RADIO USES JENNINGS VACUUM CAPACITORS

Two UH 10-75 mmfd Neutralizing Capacitors.



Two VMMHC 25-450 mmfd Main Tank Capacitors.



Four UCS 25-500 mmfd capacitors in a pi network in the driver stage. Two are also used in a line coupling circuit.

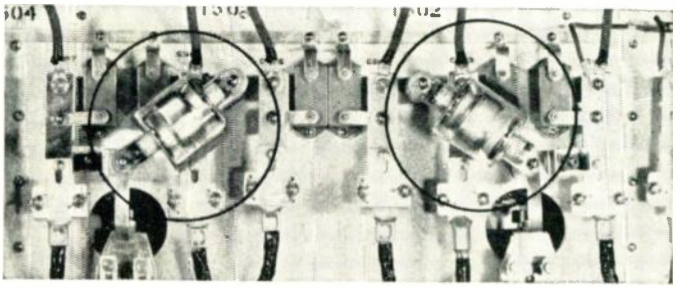
1. TOP PERFORMANCE IN ALL CLIMATES
They are impervious to dust and moisture and are self-healing after moderate overloads.

2. TROUBLE-FREE OPERATION
They require no maintenance and are designed to last the life of the transmitter.

3. WIDE CAPACITY RANGE
They make possible a continuously variable frequency range of 2 to 22 mc without complicated switching circuits.

4. SMALL PHYSICAL SIZE
Their vacuum dielectric has inherent advantages of high voltage and current ratings in a small physical space.

Send for our catalogue summary of vacuum capacitors and switches.



Two JCS 250 mmfd coupling capacitors.

JENNINGS RADIO MFG. CORP. 970 McLAUGHLIN AVE. POST OFFICE BOX 1278 • SAN JOSE 8, CALIFORNIA

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 124A)

Ruehlemann New Chief Engineer for Roller-Smith

Herbert E. Ruehlemann, a leading authority on electric fuses, has been appointed Chief Engineer and Director of Research and Development at the **Instrument Div., Roller-Smith Corp.,** Bethlehem, Pa.

In 1947, Ruehlemann was brought to the United States by the U. S. Navy to become consultant to the Naval Ordnance Laboratory and was responsible for carrying out all of their research and development in the field of electric fuses.

Glass Sealed Crystals

The development by **Rogers Majestic Electronics Ltd.,** 11 Brentcliffe Rd., Toronto, Ont., Canada, of an all-glass, military type quartz crystal which will overcome electronic equipment failures due to crystal aging is announced by M. C. Patterson, manager of the company's Tube and Component Division.

The new crystal now ready for mass production and use in civilian and military communications, was demonstrated recently to members of the Electronic Component Development Committee of the Defence Research Board. Its production in commercial quantities is made possible through the development by Rogers of machinery to evacuate the glass envelope containing the quartz crystal and provide an air-tight glass seal between the envelope and glass base. The quartz crystal is contained in a high vacuum where it maintains its activity and frequency stability under external operating conditions which cause frequent failure to crystals in metal type containers.

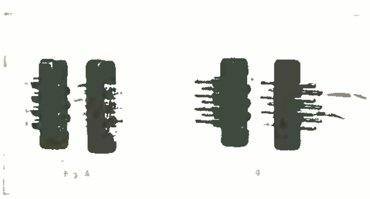
The new crystal is interchangeable mechanically and electrically with the most commonly used types of military crystals. It has the same shape and dimension as the metal HC6/U holder which encloses types CR18/U, CR23/U and CR27/U.

News-New Products

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

Contact Terminals

Two new styles of contact terminals that facilitate assembly wiring or increase the operating current of electrical connectors are now available in the miniature, quick-disconnect and small power connector class from **Winchester Electronics, Inc.,** Dept O, Willard Road, Norwalk, Conn.



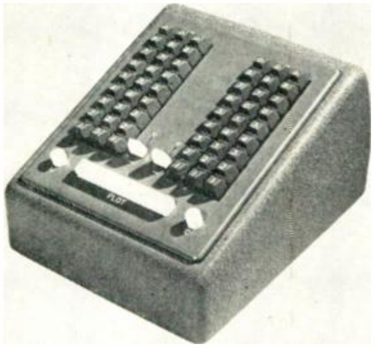
Ideal for higher currents, the turret terminals (Fig. A) enable wrap-around wiring of two or more wires per contact, or, if preferred, allow the use of larger AWG wire than that formerly permitted by conventional solder cup terminals.

Internally-tapered terminals (Fig. B) are designed for use with #53 AMP taper pins to speed individual wire attachment and permit selective engagement or disengagement of wires.

One or both types can be provided on several of this manufacturer's connector series. A complete technical bulletin is available from Winchester.

Manual X-Y Plotter Keyboard

Developed by **Librascope, Inc.,** 1607 Flower St., Glendale 1, Calif., a new manually operated decimal keyboard activates a plotter by supplying excitation for the X and Y input transducers.



Self-wiping, slide-type contacts, linked directly to the keys, completely eliminate the necessity for any adjustments of the switches and contacts.

Librascope's keyboard simulates resistance potentiometers by means of two voltage dividers (one each for the X and Y axes), each consisting of three banks of

(Continued on page 130A)

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 129A)

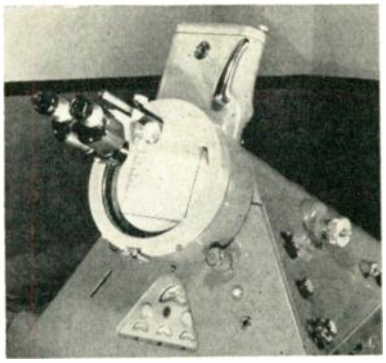
series-connected precision resistors which are switched into the input circuit by pressing the front panel keys.

The quadrant in which an input signal is controlled by the position of two selector switches on the front panel, which reverse the positive and negative connections of the regulated supply across the voltage dividers.

Pressing the plot bar on the keyboard actuates relays which permit motors in the recorder to drive the pen to the correct position on the chart. Circuits in the keyboard prevent the pen from marking the chart until the motors have positioned it in exact correspondence to the numerical value of the depressed switch keys. When the servos in the plotter reach the null, they cause a thyratron in the keyboard to fire, thereby triggering the pen-actuating mechanism in the plotter.

Front-End Camera and Binocular Attachment

A new front-end plate camera and a new binocular attachment, both adaptable to all existing Philips EM-100 electron microscopes, have been announced by the **Research & Control Instruments Div., North American Philips Company, Inc.,** 750 S. Fulton Ave., Mount Vernon, N. Y.



The new camera permits micrographs to be made direct magnifications up to 100,000× direct at the viewing screen. Cassette holds five $3\frac{1}{4}$ by 4 inch or six 4 by 5 inch cut films.

Rapid single lever action removes the cassette from the field and presents fresh film in the automatic cassette. This eliminates double exposures. Camera is equipped with a direct reading exposure counter.

The binocular attachment is available and fits over the flange ring of the Philips 100 Kv electron microscope. Objective lenses provide 3.5 and 7× magnification and aid in determining accurate image focus.

News-New Products

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

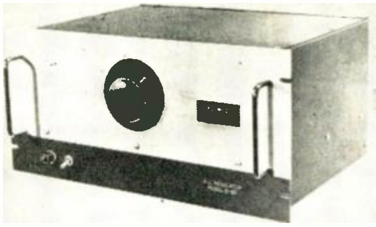
Silicon Rectifier

Transitron Electronic Corp., Melrose 76, Mass., announces the availability of high power silicon rectifiers capable of continuous operation at full rated power at an ambient temperature of 125°C.

These new rectifiers overcome the basic disadvantages of selenium, germanium, and gas filled tubes, and provide reliable operation under wide variations in ambient temperature. Their high forward conductance and low leakage current allow operation at extremely high efficiencies. In most applications, efficiencies of 90 to 99 per cent are easily achieved. The new silicon rectifiers do not exhibit aging effects. Their hermetically sealed construction provides permanent protection against the environment.

Types range in power handling ability from 10 amperes at 50 volts PIV to 5 amperes at 200 volts PIV, all rated at 125°C. Designed for conduction cooling, these rugged rectifiers provide major savings in both size and weight.

AC Voltage Regulator



El Mec Laboratories, Inc., 730 Blvd., Kenilworth, N. J., is announcing their Model E-80 ac voltage regulator, the commercially available package of a new type of ac regulation system, previously in use in precision production testing equipments and automation machines of El Mec design. The unit suppresses line harmonic distortion, maintaining a sine wave under

(Continued on page 132A)

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 131A)

extreme half-wave loading, and it has a response time in the order of 1 ms, with no thermal or magnetic circuit lag. These characteristics, along with its low dynamic impedance, make this unit of value in work involving non-linear loads, such as are encountered in semi-conductor and magnetic circuit development and test applications.

Models are available with fixed 115 volt output, or variable from 50–125 volts; for either 60 or 400 cps; and for various output currents. Further information can be obtained from El-Mec Laboratories.

TV Test Equipment

A new instrument developed by **Kay Electric Co.**, 14 Maple Avenue, Pine Brook, N. J., and named the Kay Keydsweep, eliminates spot frequency checking in Video circuit evaluation.

The Keydsweep provides internal sync pulses, and will operate with an external source of sync and blanking pulses giving pedestals and spacings in accordance with the source characteristics. It may be used with standard video sweep generators, several types of which are manufactured by Kay Electric Co. Also, provision for rack mounting is incorporated into the design.

Requirements for external sync and blanking source are: Sync, 0.5 v. neg. min. Blanking 0.5 neg. min. into 75 ohms. Price is $395.00 F.O.B. Plant. A complete description may be had by writing the manufacturer.

Transistorized Servo Systems

The first successful airplane flights with all-transistor-equipped automatic navigation and landing systems in place of vacuum tubes has been announced by **Bendix Aviation Corp.**, Teterboro, N. J.

The achievement was announced by Roy H. Isaacs, general manager of the Eclipse-Pioneer division, which last November disclosed the first successful flight of a plane controlled by a transistorized automatic pilot. Now the new flights extend transistor-equipped automatic flight to instrument landing approaches and cross-country flights on radio-marked airways.

Developed by Eclipse-Pioneer's automatic flight laboratory under the direction of Paul A. Noxon, chief engineer, the new automatic flight equipment culminates several years' research. The successfu

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

flight tests, which included eight automatic landing approaches as well as cross-country cruising on radio ranges, were made with engineering versions of a new autopilot, designated the PB-20, aboard Eclipse-Pioneer's B-25 Flying Laboratory.

The PB-20 is basically a transistorized autopilot with an integrated electromechanical computer requiring no additional equipment for automatic response to radio flight aids, Isaacs explained. As a result of its successful flight tests, Bendix engineers are working toward equipment of all aircraft flight control systems with transistors.

National Appoints Grant

Announcement of the appointment of Eugene F. Grant as Director of Engineering has been made by Joseph H. Quick, president of the **National Company, Inc.**, Malden & Melrose, Mass. Grant will be responsible for all research, development and engineering activities at National.

Prior to joining the National Company, Grant was an Engineering Manager at the W. L. Maxson Corporation. Before joining W. L. Maxson he was Chief, Computer Branch, Air Defense Group, at the United States Air Force Cambridge Research Center, where he was in charge of the engineering of large-scale digital computers and system design of automatic radars.

Capacitance Decade

Dekapacitors are a new type of precision decade capacitors which have been developed by **Electro-Measurements, Inc.,** 4312 S.E. Stark St., Portland, Ore. Any incremental capacitance value can be selected throughout the 0 to 1 microfarad range with ±1 per cent accuracy by a new style Dekadial featured for the first time on these units. Dekapacitors are completely shielded and are designed for simple installation in computers and other equip-

(Continued on page 134A)

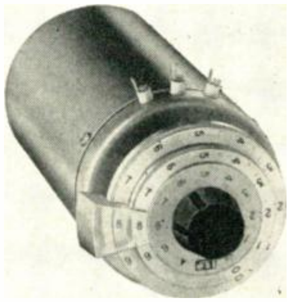
News-New Products

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

(Continued from page 133A)

ment which require wide range RC networks.

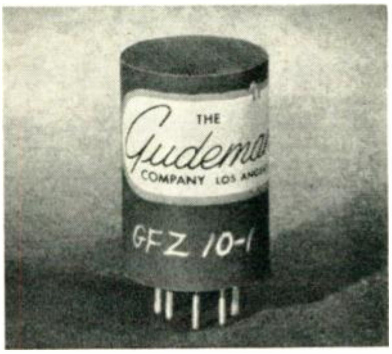
The Dekapacitor-Dekadial combination provides one thousand incremental steps of 0.001 microfarads each; direct



capacitance readings are made on the stacked deck of dials. The Dekadial's outer and middle dials have nine clearly marked positions. Tenths and hundredths of microfarads respectively are selected by these dials. The smallest capacitance increments, thousandths of microfarads, are selected by the inner dial which is provided with ten scale divisions. Solid silver electrical contacts on "dry-filmed" ceramic switches and a detent positioning structure assure accurate capacitance selection.

Ferrite Core Pulse Transformers

The Gudeman Company of California, Inc., 9200 Exposition Blvd., Los Angeles 34, Calif., has in production a new 7-pin impedance-matching ferrite core pulse



transformer with an impedance ratio of 10 to 1 for matching 1,000 ohms to 100 ohms. The unit, GFZ 10-1, is Epoxy resin impregnated and cast, is designed to surpass MIL-T-27, grade 1, class A test specifications, and is useful for coupling a 2 microsecond pulse with less than 5 per cent tilt and overshoot. Rise time is less than 0.07 microsecond. Size: $\frac{23}{32}$ inches diameter by $1\frac{1}{16}$ inches.

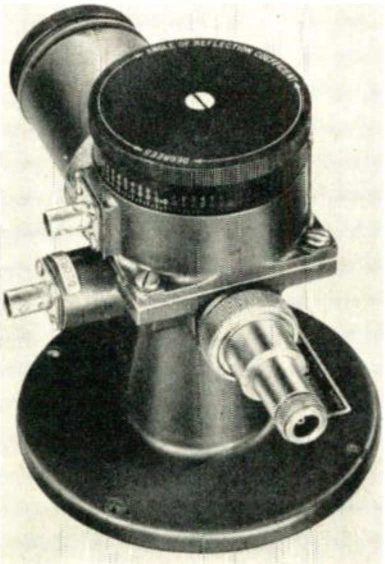
Complete data is available through Donald H. Allen at the company.

News-New Products

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

Standing Wave Detector

The 219 Standing Wave Detector was designated by **Polytechnic Research & Development Co., Inc.,** 202 Tillary St., Brooklyn 1, N. Y., to supersede bulky



slotted sections in the range of 100 to 1,000 mc. It is the small package, low cost solution for making impedance measurements easily and accurately in this region. By connecting the output to a VSWR indicator, such as the PRD Type 277, VSWR may be read directly on the indicator meter. No special detection equipment is required. The reflection coefficient angle is determined by rotating the top drum dial to a minimum indication on the meter and reading the angle on the dial directly in electrical degrees. The probe and crystal detector are self-contained.

For further information, please write to Raymond E. Jacobson, at the firm.

TV Check Tube

Sylvania Electric Products Inc., 1740 Broadway, New York 19, N. Y., has announced a new television receiver check tube, designated the 5AXP4, a 5-inch,

round, magnetically deflected tube using electrostatic self-focusing.

The new check tube, intended primarily for the television serviceman and

(Continued on page 136A)

News-New Products

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

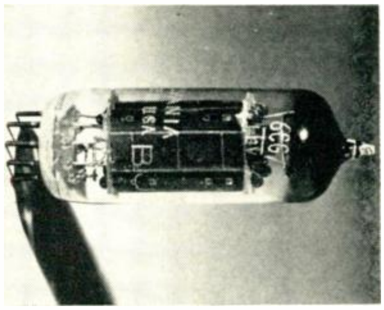
(Continued from page 135A)

for TV equipment manufacturers, will now permit a cabinet-mounted picture tube to be left in the cabinet while the receiver is being serviced in the shop. It is also a universal type of tube which can be inserted into any TV chassis while the set is being serviced.

Since the tube has a focus system built into it, no focus mechanism needs to be used on the tube nor does the ion trap need to be installed while making tests on the receiver. The tube is so light that the yoke of the receiver will very easily support the tube. Only electrical connections required are the high-voltage lead and the picture tube socket of the receiver. The tube may be used in any receiver regardless of the deflection angle. Further details can be obtained by writing the Advertising Department, Sylvania Electric.

Medium-Mu Twin Triode

A new general purpose medium-mu twin triode tube, designated the 6CG7, has been added to the line of 9-pin miniature types, produced by **Sylvania Electric Products Inc.**, 1740 Broadway, New York, N. Y.



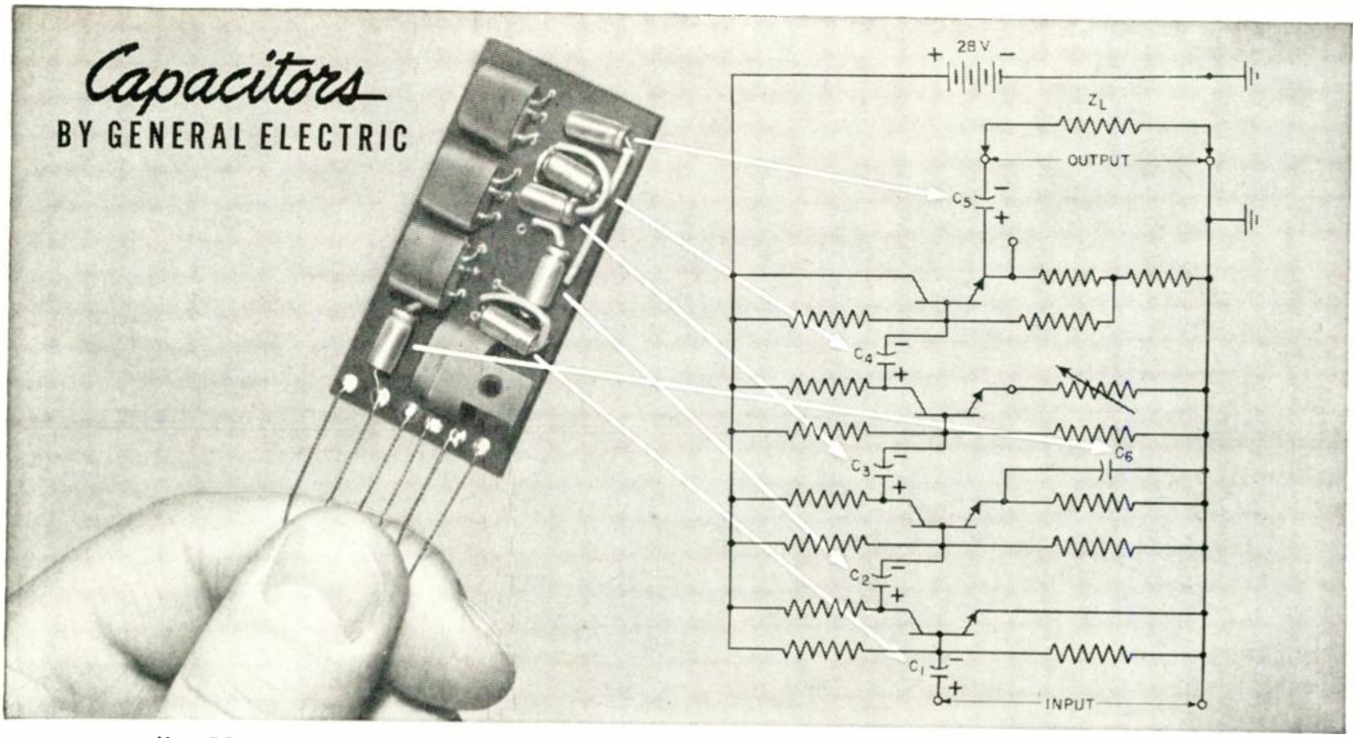
The 6CG7 is intended particularly for use as a vertical deflection oscillator and horizontal deflection oscillator in television receivers. It may also be used as a phase inverter, multivibrator, sync separator and amplifier, and resistance coupled amplifier in electronic equipment.

The new type, similar to the 6SN7GT in characteristic, is in a T6½ envelope.

It is designed with a 600 ma heater, having a controlled warm-up time to insure dependable performance in television receivers employing a single series connected heater string including the heater of the picture tube.

Design features of the 6CG7 include a structure which permits cool operation of the grids, with the result that emission from them is minimized. The structure also incorporates an internal shield which provides effective shielding between the triode units that prevents electrical coupling between them.

(Continued on page 152A)



New RC coupled, silicone transistor servo amplifier was developed for aircraft and guided missile applications. Small size of six G-E Micro-miniature Tantalytic* capacitors permitted size reduction to .68 cubic inches.

COMPACT SIZE, DEPENDABILITY, CREATE . . .

New use for tiny G-E Tantalytic capacitors in subminiature plug-in servo amplifier

Six G-E Micro-miniature capacitors rated at 8 microfarads and at 4 volts are used by the engineers at Plastics and Electronics Corp., Buffalo, N. Y., in their new RC servo amplifier. The 1/8 by 5/16 inch dimensions of the tiny capacitors enable the amplifier to be assembled and encapsulated in plastic in a 1 by 2 by 1/3 inch space.

Five of the capacitors (C1 to C5 above) are used for coupling while the sixth (C6) is for bypassing.

Because the amplifier was designed for critical aircraft and missile applications, capacitors were needed which combined small size, high ratings, and reliability.

"We chose G-E Tantalytic capacitors because they were the smallest, most dependable units with the high capacitance required for low impedance transistor devices," said Plastics and Electronics' chief engineer, Thomas L. Robinson.

If you have a design problem calling for an extremely small, high microfarad capacitor (particularly for transistorized circuits) fill out the coupon below. We will send you complete specification data and descriptive information on G-E Micro-miniature Tantalytic capacitors. For specific application information, contact your nearest G-E Apparatus Sales Office.

**Registered trade-mark of General Electric Co.*

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GENERAL ELECTRIC

General Electric Co.
Section A442-26
Schenectady 5, N. Y.

Please send me Micro-miniature Tantalytic Bulletins GEA-6065 and GET-2405.

Name . . . . Title . . . .

Company . . . .

Address . . . .

City . . . . . . Zone . . . . State . . . . . .

Positions Open

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. . . . .

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
I East 79th St., New York 21, N.Y.

POWER SUPPLY ENGINEERS

Graduate engineers with experience on tubeless regulated power supplies and magnetic amplifiers are needed. Write company complete resume, or phone Philip Diamond, President, Perkin Engineering Corp., 345 Kansas Street, El Segundo, Calif. Oregon 8-7215.

ELECTRONIC ENGINEER

Excellent opportunities with expanding Radiation Instrument Development Group for recent graduates in E.E. or Physics and for 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

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 traveling, principally in the east. The right man can expect an attractive fixed salary and expenses 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.

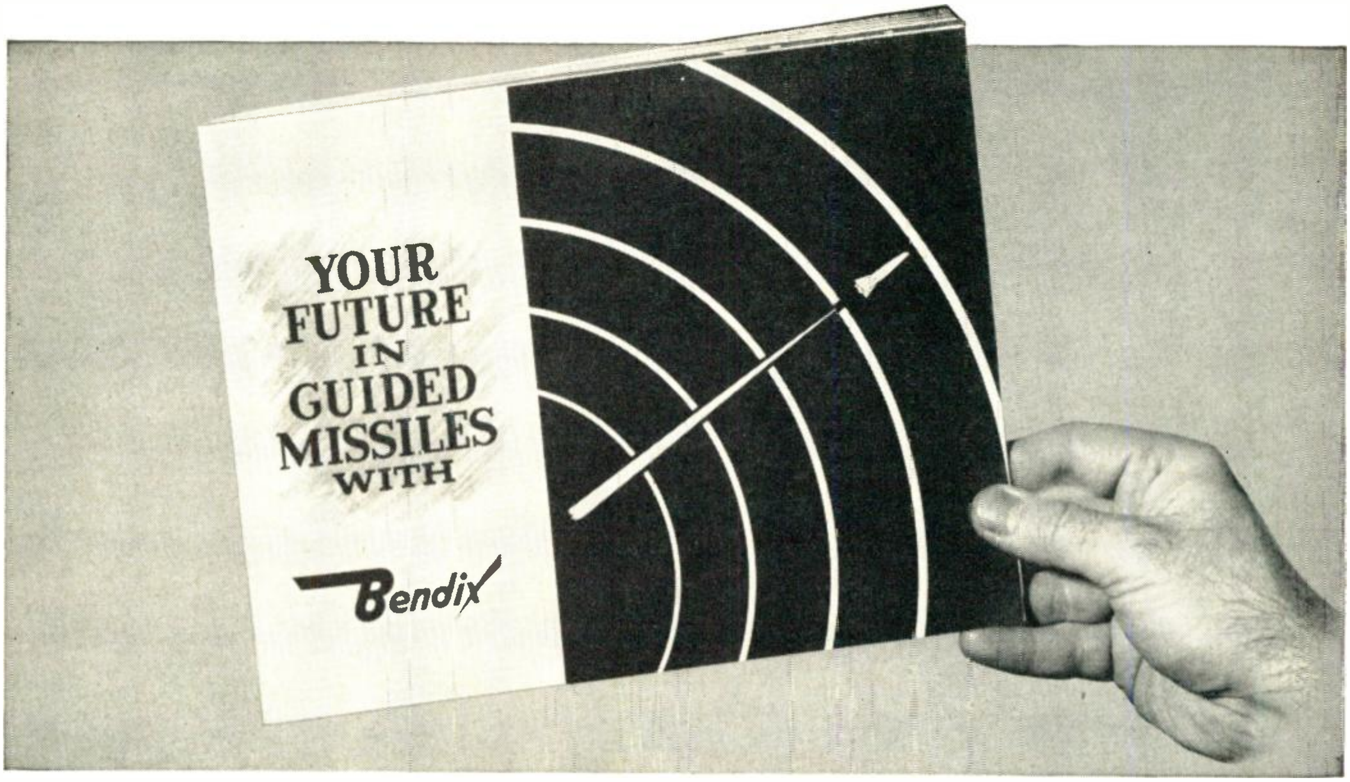
ENGINEERS

Several positions are available for engineers involving the fields of electronics, instrumentation, mechanical engineering, etc. These positions are in the engineering field and are associated with various programs involving special devices, equipment and material for military use. Salary ranges from approximately $4200 to $6000 per year. Education, training and experience will be the significant attributes in determining which position the candidate qualifies for. The above positions are available in upstate New York area and interested applicants should supply a resume of their experience to Box 811.

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 ages of 30-35, willing to travel two-thirds of the time. Contacts cover industrials and distributors. Prospects for partnership interest to qualifying applicant. Send complete resume with photograph. Box 812.

(Continued on page 140A)



If you are interested in guided missiles this book will interest you. Here is one of the most complete guides to job opportunities in the guided missile field yet published. In this book, you will find not only a complete outline of the objectives and accomplishments of the Bendix Guided Missile Section, but also a detailed background of the functions of the various engineering groups such as system analysis, guidance, telemetering, steering intelligence, component evaluation, missile testing, environmental testing, test equipment design, reliability, propulsion, and other important engineering operations. Send for your free copy today.

23 challenging opportunities in the newest and fastest growing branch of the aviation industry are now open

Bendix job opportunities in guided missiles range from top senior engineers to assistant engineers, junior engineers, technicians, and a score of other assignments.

Qualified men are given real job responsibility with Bendix and grow with the development of what is not only the nation's most important weapon system, but a project that will undoubtedly lead to new and important long-range commercial applications.

And at Bendix you will be associated with top missile authorities and have at your command unexcelled engineering and manufacturing facilities.

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Missile Section, Employment Department M
Bendix Products Division, Bendix Aviation Corporation
401 North Bendix Drive, South Bend, Indiana

Please send me a copy of the book "Your Future in Guided Missiles."

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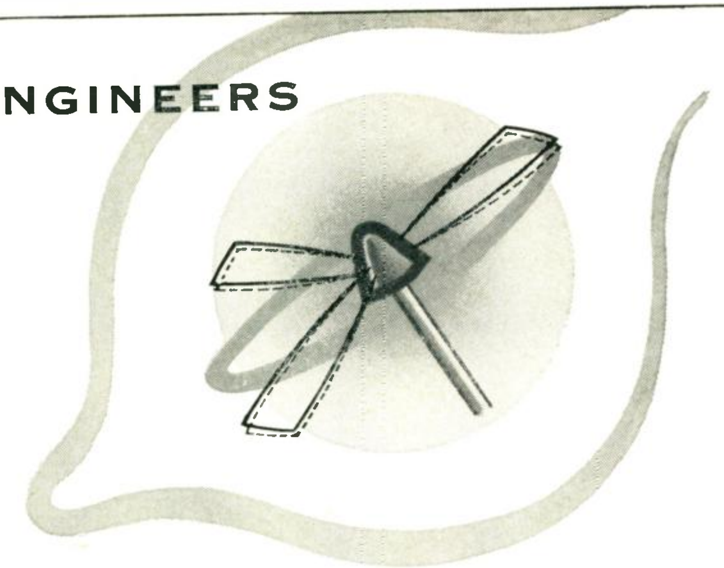
ENGINEERS



How Would You Like to Tackle this Problem?

To instrument a system for testing propellor blades, measuring to 0.1 degree and 0.1 inch respectively, the angular and lateral deflection of a blade rotating at 8000 rpm—with all equipment OUTSIDE of the wind tunnel.

This is just one example of the wide variety of projects at ECA, and it was accomplished by an ingenious application and modification of a well-known device.

To develop those automatic controls, electronic business machines, digital and analog computers that will best meet the increasing need for automation in business and industry requires engineering skill of the highest order...a well developed professional curiosity...a diversity of personal interests.

Men of this calibre come to ECA—and stay—for more than one reason: there's the freedom which encourages individual approach and initiative, the compensation high on the industrial scale, the stability based on the success of many established commercial products.

And as ECA continues to grow, daily exploring new fields and enlarging its interest in old ones, there's room for more exceptional engineers. Please send full details of your experience and education to Mr. W. F. Davis, Dept. 714.

ELECTRONICS CORPORATION OF AMERICA

77 Broadway Cambridge 42, Mass.

Positions Open

(Continued from page 138A)

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.

TECHNICAL WRITERS & PUBLICATIONS ENGINEERS

Electrical engineers to edit and write copy for instruction books and technical manuals. Creative opportunities for professional development and expression in large department. Direct inquiries to Mr. R. T. Hamlett, Publications Mgr., Sperry Gyroscope Co., Great Neck, N.Y.

ELECTRONIC ENGINEER

Electronic engineer with audio experience desired on Consultant basis. Write, giving general information and fee expected. Box 813.

ENGINEERS

The Diamond Ordnance Fuze Laboratories in Washington has vacancies primarily for electronic scientists, electronic and mechanical engineers and physicists, with salaries ranging from $4035 to $8360 per year. These vacancies involve such work as design and development of electronic and electromechanical fuzes, testing of fuzes and related components, electron tubes and transistors, and research and development in the various physical science and engineering fields. Write to L. P. Conners, Civilian Personnel Office, Diamond Ordnance Fuze Labs., Washington 25, D.C.

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 B.S. or advance degree electrical engineers or physicists with a minimum of 3 to 5 years electromagnetic 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 144A)

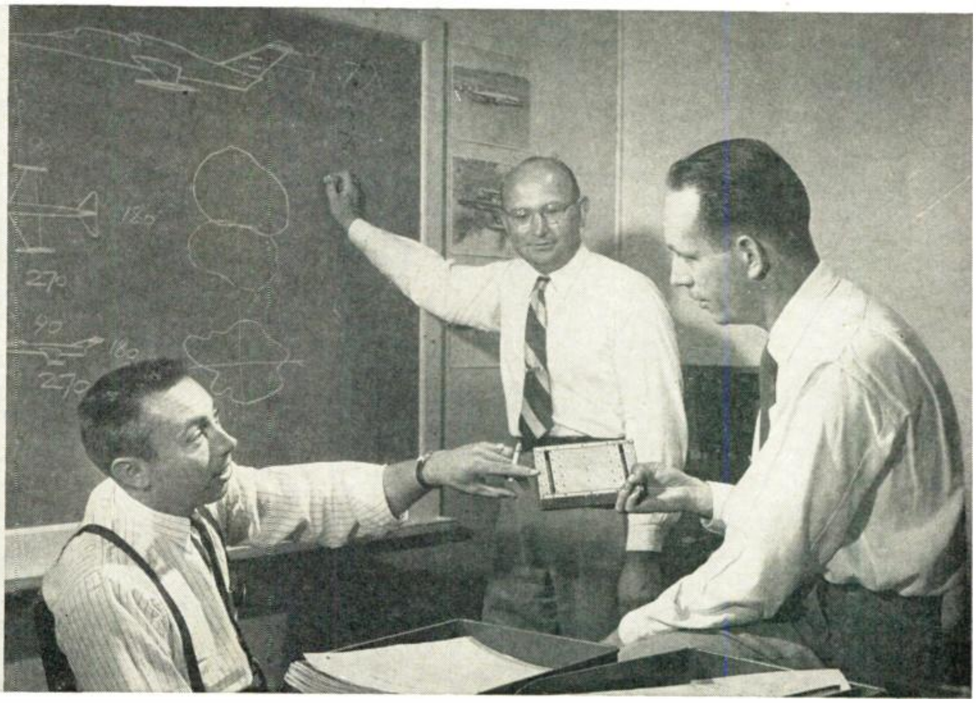
Lockheed antenna program offers wide range of assignments

Airborne Antenna Design is one of the fastest-growing areas of endeavor at Lockheed. Advanced development projects include work on stub, slot, reflector-type, horn and various dipole antennas.

These diverse antenna activities reflect the full scope of Lockheed's expanding development and production program. For with 13 models of aircraft already in production and the largest development program in the company's history underway, the work of Lockheed Antenna Designers covers virtually the entire spectrum of aircraft, commercial and military.

Lockheed's expanding antenna development has created a number of new positions on all levels for qualified antenna designers. Those interested are invited to write E. W. Des Lauriers, Dept. A-8-5.

Research Specialist Edward Lovick (right) discusses application of experimental slot antenna in the vertical stabilizer of a high-speed aircraft with Electronics Research Engineer Fred R. Zboril and Electronics Research Engineer Irving Alne.



Lockheed increases engineers' salaries

Salaries, rate ranges and overtime benefits have been increased. In addition, employe benefits add up to approximately 14% of each engineer's salary in the form of insurance, retirement pension, etc.

Generous travel and moving allowances enable you and your family to join Lockheed at virtually no expense to yourself.

An address by Edward Lovick on "An Electronic Square-Rooter and Pattern Integrator for use with Antenna Range Systems" is available to interested engineers. Address inquiries to Mr. Lovick.

Lockheed AIRCRAFT CORPORATION

CALIFORNIA DIVISION • BURBANK **California**

. . . with *your* future in mind

Check the chart below for positions which interest you most . . .

FIELDS OF ENGINEERING ACTIVITY	TYPE OF DEGREE AND YEARS OF EXPERIENCE PREFERRED											
	Electrical Engineers			Mechanical Engineers			Physical Science			Chemistry Ceramics Glass Technology Metallurgy		
	1-2	2-3	4+	1-2	2-3	4+	1-2	2-3	4+	1-2	2-3	4+
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			

Location Code

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.

O—Overseas—domestic and overseas locations.

W—Waltham, Mass.—near the cultural center of Boston.

X—Los Angeles, Calif.—west coast electronics center.

Please send resume of education and experience, with location preferred, to:

Mr. John R. Weld, Employment Manager
Dept. C-1F, Radio Corporation of America
30 Rockefeller Plaza
New York 20, N.Y.

RCA ®

RADIO CORPORATION of AMERICA

Positions Open

(Continued from page 140A)

PHYSICIST OR ELECTRONICS ENGINEER

Physicist or electronics engineer to design, construct and install setups 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.

ENGINEERS

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.

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 6, 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.

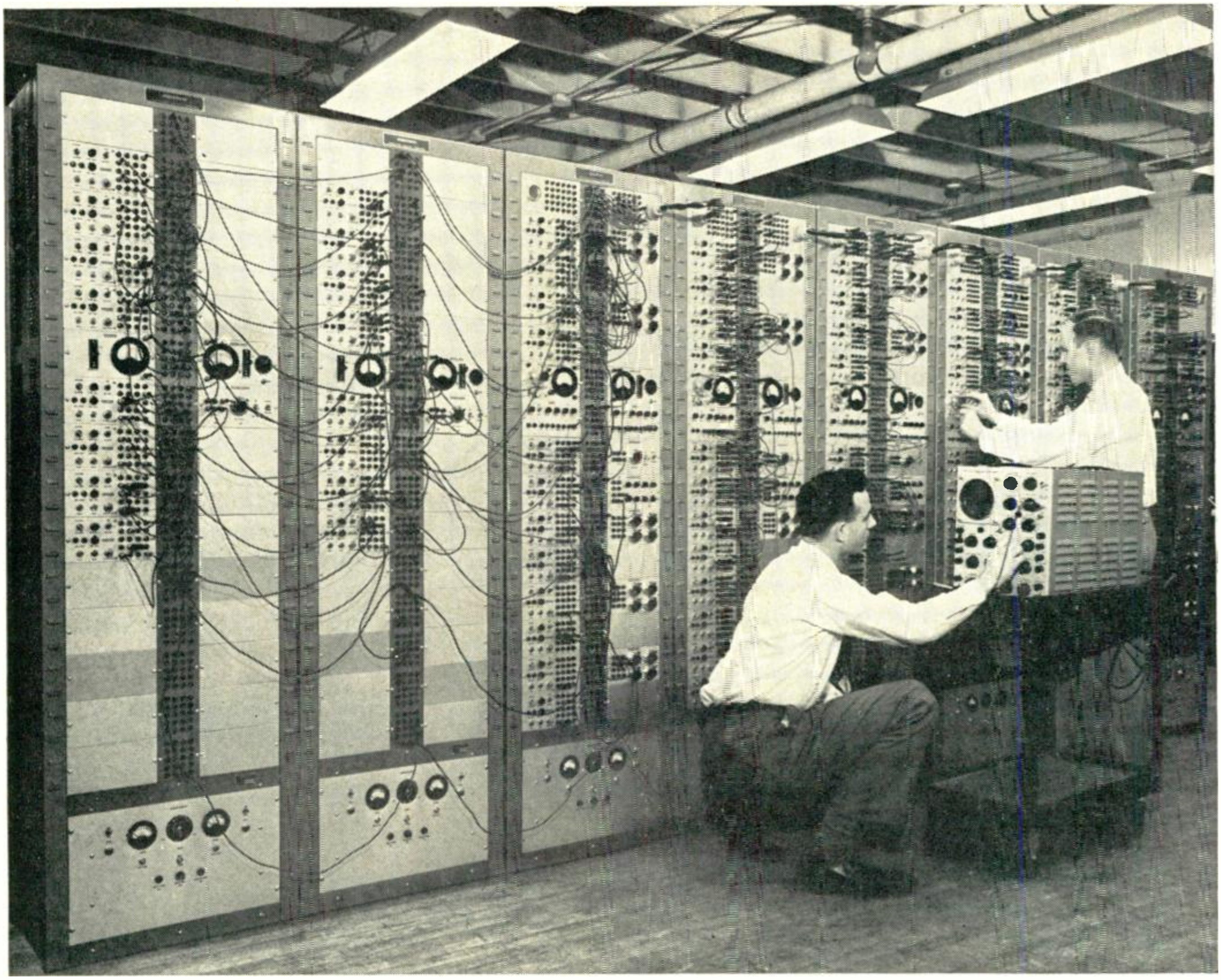
INSTRUCTOR

Instructor in electrical engineering beginning Sept. 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, Nebraska.

PROFESSOR

Position as assistant professor in electronics at Middle Atlantic university. Undergraduate and graduate instruction. Doctorate in E.E. or physics preferred, but will consider experience as alternate. Salary $5000 for nine month session. Box No. 818.

(Continued on page 146A)



Boeing electrical engineers work with superb equipment

Boeing electrical engineers designed this computer—and use it to answer in seconds questions that formerly took weeks. It is one of many advanced facilities that help Boeing electrical engineers solve the challenging problems of tomorrow's aviation, and maintain unsurpassed prestige.

Electrical engineers at Boeing play a vital role in designing and developing high-performance airplanes, guided missiles and components of the future. They find truly creative opportunities in flight test instrumentation, acoustics, radar systems design, autopilot development for guided missiles, and many other fields.

The new multi-million-dollar Boeing Flight Test Center is the largest installation of its type in the country. It houses the latest electronic data reduction equipment, instrumentation laboratories, and a chamber that simulates altitudes up to 100,000 feet. Other facilities include the world's most versatile privately owned wind tunnel, and laboratories for research in rocket, ram jet and nuclear power.

A fourth of Boeing's engineers have been with the company more than ten years. In addition to career stability, Boeing engineers find individual recognition in regular merit reviews and promotions from within the organization.

At Boeing you will work in tightly knit design or project teams with the pick of the country's engineers. You will help design worthy successors to the B-47 and B-52 global jet bombers, the 707 jet tanker-transport, and the Bomarc IM-99 guided missile. You will join a progressive, solidly growing company that now employs twice as many engineers as at the peak of World War II.

JOHN C. SANDERS, Staff Engineer—Personnel
Boeing Airplane Co., Dept. G-40, 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)________ Year(s)____ Degree(s)____

Address____________________

City__________ Zone____ State______

BOEING

Aviation leadership since 1916

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Positions Open

(Continued from page 144A)

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 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.

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.

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BS Engr./Bus. Admn., MBA Marketing. 10 years progressive experience all phases of industrial marketing. Program planning, budgeting and administration, market survey, agency liaison, media evaluation, copywriting and production. Pamphlets & brochures, catalogues and direct mail, trade shows and technical publicity. Media, industry, community and Government relations. Licensed radio operator with background in radio and electronic equipment promotion. Age 35. Desires career position offering greater responsibilities and advancement. Box 811 W.

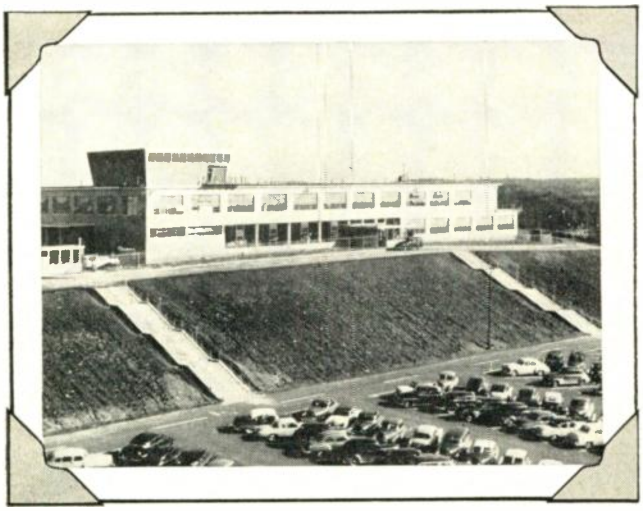
PHYSICIST

MS physics, 1952. Age 30, married. Research and development experience in electronics, ion devices, vacuum systems and instrumentation. Prefer location in southwest or Florida. Box 813 W.

(Continued on page 148A)

Planning a New England vacation?

Visit Raytheon



Raytheon Missile and Radar Division's new laboratory next to Hanscom Air Force Base, Bedford, Mass. Another engineering facility nears completion at Wayland, Mass.

Following your visit with us, be our guest for lunch or dinner at the picturesque Wayside Inn, South Sudbury, Mass., immortalized by Henry Wadsworth Longfellow.

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RAYTHEON MANUFACTURING COMPANY

Waltham 54, Massachusetts

Positions Wanted

By Armed Forces Veterans

(Continued from page 146A)

SYSTEMS ENGINEER

BSTE January 1949. Age 30, married, 1 child. 6 years diversified experience: radar, automatic data reduction systems, digital computers, telemetering and instrumentation systems. Desires similar project or systems engineering position. Box 814 W.

ELECTRONIC ENGINEER

BEE 1950, MEE expected 1955 from New York University. Age 30, married. 5 years experience as electronic circuit and development engineer on automation. Considerable production and mechanical experience. Desires position in New York City area until June 1955 and then will relocate. Box 817 W.

ELECTRONIC ENGINEER

BEE 1955 (Jan.). Age 34, married. 3 years radio coils, 2 years radar, 2 years VHF communications, 2 years radio and TV service. 2 years dial switching ckts., 3 years TV broadcast lab. (color), 2 years sales. Desires responsible position where broad experience can be utilized. Box 818 W.

BROADCAST ENGINEER

RCA graduate. 1st class ticket. Ambitious beginner. Desires position in radio or TV station anywhere, to start at bottom. Anxious to learn all aspects involved in broadcasting. Salary secondary. Box 828 W.

TRANSISTOR ENGINEER

AB, BSEE, MEE, Tau Beta Pi, Eta Kappa Nu, Sigma Xi, pre-doctoral student and EE instructor. 2 years experience design and teaching audio and pulse transistor circuits. Also several years vacuum tube circuit experience. Desires summer job in New York City. Box 829 W.

RADIO ENGINEER

BSEE 1950, Oregon State College, age 28, married. 4½ years experience in airborne electronics and radio aids to navigation, theoretical and practical. Desires position with future with a progressive company in the west, preferably the northwest. Box 830 W.

ENGINEER
(No license)

Age 27, single. 3 years Army. 6 years civilian experience in electronics. Speak, read and write Spanish fluently. Desires position in Latin-America. Box 832 W.

ELECTRONIC ENGINEER

Five years of missile, radar and fire control system study work, and four years of radar and missile component development prior to that. Desires position in southern South America. Box 833 W.

RADIO-TV TECHNICAL DIRECTOR

Six years experience in program production and direction. Education: BA in programming and production. Technical background includes control room operations, equipment design, construction, maintenance. 1st phone license. Age 26, married. Completing Army duty as microwave instructor at the Signal School in June 1955. Prefer Chicago or vicinity. Box 834 W.

(Continued on page 150A)

ELECTRONIC ENGINEERS

ADVANCE YOUR CAREER WITH A LEADER IN WESTERN ELECTRONICS

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- *production*
- *specialized military equipment*
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Special receivers and transmitters, DF and DME, various instruments and Transistor applications—special devices. Studies in noise, radar, miniaturization and test equipment. Relocating expenses, good insurance plan, central location, steady advancement.

Send resume to L. D. Stearns
Engineering Employment Manager

Hoffman
LABORATORIES, INC.
(SUBSIDIARY OF HOFFMAN RADIO CORP.)
3761 S. HILL ST., LOS ANGELES, CALIF.

SUPERVISOR

Receiving-Tube Design

Should have about 10 years' experience in this field. Salary open.

- **Wonderful plant location**
- **Unusual benefits**
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Traveling and moving expenses paid. All replies will be handled in the strictest confidence. Send resume, Box P.R.E. 310, 221 West 41st St., N.Y.C.

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Between 1947 and 1953, the electronics industry grew 24%...Sylvania grew 32%.

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- Mechanical Design
- Miniaturization
- Digital Computer Circuits & Systems
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- Technical Writing
- Missile Analysis

BUFFALO
Engineering

Majors in E.E., M.E., or Physics. Experience in Product Design and Advanced Development in—

- Circuit Design
- Systems Development
- Pulse Techniques
- F.M. Techniques
- Equipment Specifications
- Components
- Microwave Applications
- Servo Mechanisms
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- Mechanical Design
- Shock & Vibration
- Heat Transfer

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Professional Placement Supervisor

SYLVANIA ELECTRIC PRODUCTS INC.

Thomas A. Tierney 100 First St. Waltham, Mass.	Randall A. Kenyon 175 Great Arrow Ave. Buffalo 7, N. Y.

SYLVANIA
SYLVANIA ELECTRIC PRODUCTS INC.
Your inquiries will be answered within two weeks

Positions Wanted

By Armed Forces Veterans

(Continued from page 148A)

ELECTRONICS RESEARCH

BEE 1946, MEE 1950 electronics. Age 29, married, 1 child. 4 years experience hyperbolic radio navigation systems research, creative design, construction, analysis, laboratory and field evaluation. Lieut. Naval Reserve. Member I.R.E. P.G.A.N.E and I.O.N. Presently Unit Head. Desires similar position with advancement opportunity industry or university research program. Box 835 W.

ENGINEER

Instructor of technical electricity and electronics with extensive field and teaching experience. Desires (within 25 miles of Poughkeepsie, N.Y.) a teaching position with some H. S. or college or some phase of technical-engineering work with a private firm. Box 836 W.

ENGINEER

BSEE 1952 communications option. 1½ years graduate work in nuclear engineering and EE. 2 years as electronics technician, Navy. 2 years industrial experience in electronic design. Desires job placement in electronics work, possibly numerical controlled machine tools. Box 837 W.

ENGINEERING PHYSICIST

BS Physics, MS Math. 2 years graduate study in physics. Age 28, married. 4 years research experience in instrumentation, high speed pulse techniques, delay line design, and magnetic resonance phenomena. Patents. Desires research and/or development position with a challenge. Box 845 W.

ELECTRICAL ENGINEER

SB EE (MIT'45), MS, Ph.D.; honors; married; age 29; 3 years teaching, 5 years applied research; PGAP, PGIT, PGCT, PGEM; papers. Desires opportunities for further creative professional development. Brochure available to companies on Eastern seaboard only. Box 846 W.

ELECTRONIC ENGINEER

BS Physics 1950. Age 28, married, one child. 2 years Navy ETM. 4 years additional experience including component research and circuit work. Desires position in connection with long term basic research. Box 847 W.

FIELD ENGINEER

BSEE with broad background in field engineering including communications, radar and navigational aids. Desires overseas position which will make full use of capabilities. Box 848 W.

ENGINEER

BEE 1951, MSEE expected in August 1955, communications option. Age 25. Married, 2 children. 1 year electronic design of radar equipment, 2 years military experience fire control equipment and digital computers. Interested in medical electronics. East preferred. Box 849 W.

ASSISTANT SALES MANAGER

Electrical engineer. Age 30. Good personal sales record, excellent on customer contact, negotiations, product development, market analysis. Box No. 850 W.

ENGINEER

BEE, MBA. Age 33, married. Graduate work in transistors, receiving tube design and applications: 5 years radar tech; 4 years in receiving tube design and applications. Desires challenging opportunity with aggressive company. Box 851 W.

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ENGINEERS

MICROWAVE

Development of microwave instruments & test equipment

ELECTRONIC

Development of electronic instruments

Precision instrument manufacturer requires men with good academic & practical background. At least 2-3 years design & development experience required. Should exhibit qualities of leadership, with the ability to meet & deal with people. Men who fill the bill will be substantially compensated.

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ALL AMERICAN CABLES & RADIO, INC. DOUGLAS AIRCRAFT CO. TRANS-CANADA AIR LINES
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Shortage of Engineering Personnel?

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TODAY'S recruiting efforts hardly help ease today's shortage. But, what about the future? The scarcity shows no signs of letting up.

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CREI's instruction material is prepared and administered in a personalized style by engineers who know how to teach and who know the requirements of the electronics industry.

Whether yours is a large company, or small, let us show you how CREI group training is working for others . . . how to set up a program that works . . . and how CREI can help you prepare for tomorrow's continuing shortages by "developing your own."

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Washington 10, D.C. HObart 2-1520

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 136A)

Direct-Coupled Amplifier

The new Model D-1 direct-coupled amplifier, manufactured by **Southwestern Industrial Electronics Co.,** 2831 Post Oak Rd., Houston, Texas, has a maximum useful over-all gain of 80,000, while frequency response flat from zero to more than 100 kc is available at gain settings up to 10,000.

Excellent signal-to-noise ratio, low drift, and wide dynamic range are achieved through advanced circuitry and careful mechanical design. High input impedance, with either single-ended or differential connection, enable the D-1 to amplify signals from a wide variety of input transducers and other sources for presen-

(Continued on page 154A)

News-New Products

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

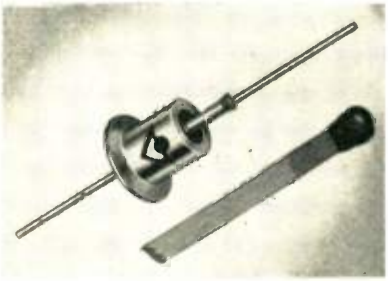
(Continued from page 152A)

tation and recording on cathode-ray tubes, direct-writing recorders, galvanometer oscillographs, and magnetic media.

A fully regulated power supply is in the instrument and the complete unit is designed for rack or bench mounting at the user's option.

Germanium Diodes

International Rectifier Corp., 1521 E. Grand Ave., El Segundo, California, has announced the availability of its new germanium diffused junction power diodes which offer low leakage and high rectification efficiency.



Reliability is featured in the design of these diodes as a result of complete hermetic sealing of the housing, consisting of glass-to-metal and welded metal-to-metal seals throughout.

This technique and construction have been the subject of considerable study in the International Rectifier Research and Development Laboratory. Standard types such as 1N91, 1N92 and 1N93 diodes are now available from production. For special applications, send your requirements to the Semiconductor Division, International Rectifier Corp.

Signal Generator

A new signal generator, Model 3432-A, is now in production by the **Triplett Electrical Instrument Co.**, 286 Harmon Rd., Bluffton, Ohio.

The new generator has complete frequency coverage from 160 kc to 110 mc

(Continued on page 156A)

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 154A)

with no skips in frequency for AM-FM radio, monochrome or color TV servicing. The rf circuits are double shielded with copper steel shields, and a cathode follower output provides good stability by acting as a buffer to the oscillator.

Jacks are provided for either internal modulation or audio output, and both are controlled by audio control to provide variable modulation or af output.

The new generator, which has a step-type, three-position attenuator, weighs 12 pounds, and sells for a dealer net of $99.50.

Comparison Bridge

Southwestern Industrial Electronics Co., 2831 Post Oak Road, Houston, Texas, announces a new model of its Comparison Bridge, to be designated the E-2.

The instrument can now be operated by means of an optional foot switch, as well as by the conventional push button. Improved stability, accuracy and sensitivity enable resistors, capacitors and inductors to be compared over a range of one ohm to five megohms at sixty cycles. Differences as small as 0.1 per cent may be easily measured on the 1 per cent full scale meter range; other scale ranges of 2.5, 5, 10 and 25 per cent are also provided.

In addition to its use as a comparator and go-no-go gauge, the E-2 may be combined with a decade box to make quick and accurate measurements of absolute values.

Grid Brochure

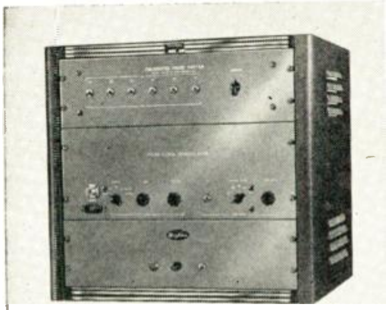
Baker & Co., Inc., 113 Astor St., Newark 5, N. J., has published a new brochure on Platinum Clad Tungsten Wire for high power vacuum tube grids. Two charts illustrate relative sag of platinum clad tungsten at fixed and varying temperatures.

News-New Products

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

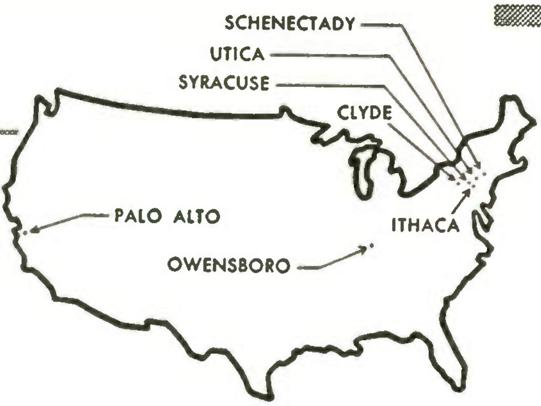
Color Phaze Analyzer

Wickes Engineering and Construction Co., 12th St. & Ferry Ave., Camden, N. J., has designed the CPA-1 Color Phase Analyzer, a new color test instrument to analyze the chrominance components of a composite color video signal.



The equipment compares the phase of the chrominance components with respect to the color synchronizing (reference subcarrier) signal. The phase difference between any two intervals of the composite

(Continued on page 159A)



New

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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.

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Bachelor's or advanced degrees in Electrical or Mechanical Engineering, Physics, and experience in electronics industry necessary.

Please send resume to:
Dept. 6-5-P, Technical Personnel

GENERAL ELECTRIC

ELECTRONICS PARK, SYRACUSE, N. Y.

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EMERSON OFFERS

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IN THE FIELDS OF

FIRE CONTROL	ELECTRONIC COUNTERMEASURES
PULSE TECHNIQUES	MICROWAVE TECHNIQUES
ANTENNAS	NETWORK THEORY
RADAR	OPERATIONS RESEARCH

submit resume to
Dr. R. E. Wilson

Emerson research laboratories

701 LAMONT STREET, N.W. WASHINGTON 10, D.C.

a division of EMERSON RADIO & PHONOGRAPH CORPORATION

Wire Line Carrier Systems Engineer

wanted for high-level position with managerial responsibilities in established electronics manufacturing company. Should have 5 years experience in wireline carrier telephone work with at least 2 years spent in applications, either with manufacturer or large carrier operator. Experience in carrier operation in armed forces is applicable. Bachelor's degree or professional engineer status required. Some travel, about 10 percent of time, will be necessary. The right man for this job will find no quibbling about salary.

Stromberg-Carlson Company has been pioneering in telephone industry for 60 years, and is known for high quality of products. One of fastest growing companies in the industry, offering excellent opportunities for advancement. Liberal bonus. Other benefits.

Located in Rochester, New York, in heart of beautiful Finger Lakes country. Community noted for fine schools, cultural and recreational facilities. Also have openings for several carrier and microwave sales engineers. Must be free to travel in limited territory. Successful sales experience required. There may be an opening in your area. If you qualify for either of above positions, send resume of experience to General Manager, Telephone Division,

Stromberg-Carlson Co.
Rochester 3, New York

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 157A)

signal can be measured also. Phase delay can be measured over the entire range of 0 to 360°. The CPA-1 facilitates the alignment of color coders, and assures accurate signal certification. The equipment also can be used to measure the differential gain of any amplifier or system. A five position function switch permits selection of the desired output signal, for comprehensive analysis of the color signal. The complete equipment includes a CPS-1 Calibrated Phase Shifter, a CSD-1 Color Signal Demodulator, and a PS-7 Regulated Power Supply, plus a cabinet rack and set of interconnecting cables.

Test Speaker

Authorized Manufacturing Co., of 919 Wyckoff Ave., Brooklyn 27, N. Y., announces that their new Model #401 Unispeak universal test speaker is in production.

Unispeak's flexibility is enhanced by a socket providing connections for most sets which are frequently encountered. An

(Continued on page 161A)

News-New Products

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(Continued from page 159A)

adaptor cable is made up, and plugs in to the speaker and/or the set. This feature eliminates the need for pulling a speaker from the cabinet.

Single ended or push-pull out transformer circuits are accommodated, as well as direct voice coils, through the Vari-Z switch. The transformer is a universal type and allows impedance matching of all tubes.

The unit includes a choke and a variable bleeder which permits correct matching for all electro dynamic speakers. These are also available at separate terminals for filter testing.

Power Oscillators

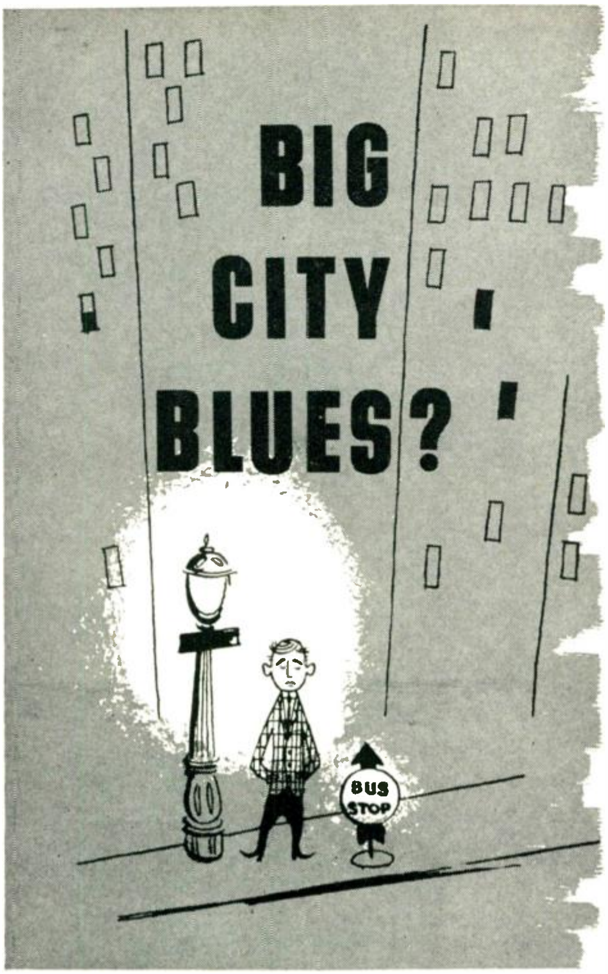
New Model DK-1 low distortion fixed frequency power oscillators which feature essentially pure, highly stabilized sine wave power at 10 volts, 2 watts, 0.03 ohm impedance for general laboratory and production testing, is available from **Neucor, Inc.**, 45 W. Union St., Pasadena 1, Calif.

Frequency is factory fixed to customer's order in range 300 to 10,000 cps; a trimmer provides fine adjustment.

Total harmonic content of sine wave is less than 0.08 per cent. Amplitude shift be-

tween no load and full load or due to line voltage variations of 10 volts is less than 0.18 per cent. Frequency shift between no

(Continued on page 163A)



ENGINEERS — PHYSICISTS

Head for the *friendly* Southwest

. . . if you're looking for a better career opportunity . . . and the kind of life that lets you enjoy it! Here in the Southwest (incidentally, it's the nation's fastest growing industrial frontier) you'll find friendly people (including the boss), a more relaxed attitude toward the business of living, many cultural and recreational advantages, plus ready and waiting job opportunities in your specialized field. And the climate? Well, it's the best one we know . . . for success!

If you're a graduate engineer or physicist (B.S., M.S. or Ph.D) and want to engage in research or development work with transistors, radar, sonar, waveguide and transformer components, or geophysical equipment, send resumê to W. D. Coursey, Personnel Division. We'll be glad to interview qualified applicants to discuss career opportunities at Texas Instruments . . . our Profit Sharing and other benefit programs . . . and to show you our interesting research and manufacturing activities.

TWENTY-FIFTH YEAR OF PLANNED GROWTH

TEXAS INSTRUMENTS
INCORPORATED
6000 LEMMON AVENUE   DALLAS 9, TEXAS

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.

Vacancies exist at the Marine Physical Laboratory for men of originality with sound scientific background in Physics and Mathematics. The Marine Physical Laboratory of the Scripps Institution of Oceanography is a unit of the University of California.

The staff is small, the problems numerous and there is ample opportunity for the conduct of individual research in underwater acoustics, electronics, "noise" backgrounds and other fields falling under the wide classification of Marine Physics. Salaries are based on the University of California scales. Applications with statement of qualification should be addressed to: Director, Marine Physical Laboratory, San Diego 52, California.

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.

News-New Products

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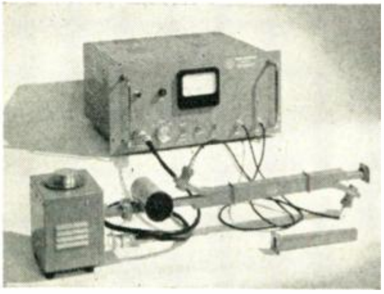
load and full load is less than 0.16 per cent; shift to line fluctuations of 10 volts is nil.

The unit is housed in a ventilated metal case $3\frac{3}{4}$ wide by $4\frac{3}{8}$ high by $8\frac{1}{2}$ inches deep.

The circuit is an L-C, bridge type, incorporating a high-Q toroid, combined with a self-balancing feedback amplifier.

X-band VSWR Measuring System

Retaining the speed and high accuracy of earlier models, the new CTI Model 110B X-Band VSWR Measuring System produced by **Color Television Inc.,** 974 E. San Carlos Ave., San Carlos, Calif., has an added attenuation scale and new VSWR scales reading from 1.02 to 1.20 and 1.1 to 2.50. The system includes a tunable oscillator permitting complete and continuous coverage from 8500 to 9600 mc, an accurate wavemeter to supplement the direct-reading dial of the oscillator, a bi-directional coupler with bolometer detectors for incident and reflected power, and a direct-reading VSWR indicator.



Employing a ratiometer type of indicator, the CTI system is "direct-reading," as no adjusting or zero-setting is necessary when the oscillator is tuned through the X-band or the load is changed.

The overall accuracy of the system is better than 2 per cent which makes it ideal for laboratory measurements. Simplicity of operation facilitates its use by unskilled personnel, making it excellent for production go/no-go tests as well. The accuracy is maintained throughout the frequency range because of the matched characteristics of the bi-directional coupler arms and integral bolometer mounts.

Other specifications are: rf power source V-260 klystron; wavemeter accuracy, ±0.08 per cent; directivity of couplers, greater than 45 db; output waveguide fitting, UG-39/U; indicator cabinet, $8\frac{3}{4}$ inch front panel for bench use or standard 19-inch rack; overall length of waveguide assembly, $31\frac{3}{4}$ inches, primary power, 115 v, 60 cps.

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News-New Products

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(Continued from page 163A)

Tape Recorder

Premier Electronic Laboratories, 382 Lafayette Street, New York 3, N. Y., has developed a recorder which provides three separate heads for monitoring from tape while recording; and the A.B. switch permits comparison between original and recorded program. It is a dual track unit with $4\frac{1}{2}$ inch VU meter. Push button operation, using 7 electrically interlocked dc relays and 1 solenoid. Three speeds are provided: flutter and wow, 0.1 per cent at 15 inches/sec.; 0.2 per cent at $7\frac{1}{2}$ inches/sec.; 0.3 per cent at $3\frac{3}{4}$ inches /sec. It will play 4 hours at $3\frac{3}{4}$ inches/sec.

The sound is high-fidelity, provided by a 10-tube 12 watts push pull amplifier. It has mixing channels for mike, radio or phono inputs. The frequency responses are: 40–16,000 cps at 15 inches/sec.; 40–13,000 cps at $7\frac{1}{2}$ inches/sec.; 40–6,500 cps at $3\frac{3}{4}$ inches/sec. It has an 8 inch extended range speaker. Drive mechanism uses three heavy duty dynamically balanced motors. Fast forward and rewind, less than one minute for $10\frac{1}{2}$ inch reel. Electro dynamic brake action and tape tension never requires adjustment. The recorder measures $16\times23\times11$ inches overall and weighs 62 lbs. It comes in a portable case with 2 covers and is priced at $368.50 net. Additional information may be obtained by writing to S. Miller, at the firm.

50-KV-Isolation Supply

Developed to provide the flexibility of use necessary for experimental work with klystrons, traveling-wave tubes, and other microwave devices, a series of new power supplies is available from **Levinthal Electronic Products, Inc.,** 2979 Fair Oaks Ave., Redwood City, Calif. The unit is 50-kv isolation from ground on both sides of the high-voltage circuit so that several units can be operated in series where desired. Two of the models (illustrated) have continuously-variable output from 0 to 1000 volts and from 0 to 2000 volts at 200 milliamperes. Larger-capacity units of similar design up to 50 kv are also available.

(Continued on page 168A)

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 164A)

Operable from inputs of 105 to 125 volts single-phase 60-cps power, these two supplies exhibit regulation to a maximum of ½ per cent over that input range, and ripple of less than 0.5 volts peak-to-peak at any output.

Units include switches for line power and high voltage as well as continuously-variable controls for coarse voltage and fine voltage. Suitable metering with 1.0 per cent accuracy is included. Units are completely protected by inter locks and automatic shortcircuiting devices.

Push-Button Oscillator

Any of 2000 different frequencies are available on the Model 440-B Precision Push-Button Oscillator, designed and developed by **Krohn-Hite Instrument Co.,** 580 Massachusetts Ave., Cambridge 39, Mass., which covers the frequency range

from 0.5 to 1000 cps in ½ cps steps. Calibration accuracy is ±0.05 per cent and then drift per hour is less than 0.005 per cent.

Distortion and hum are less than 0.1 per cent at any output level. Amplitude varies less than ±0.25 db over the entire

(Continued on page 170A)

Sturtevant TORQUE TESTING FIXTURE

FOR TESTING Screws, thread-cutting and thread-forming screws—all types of threaded fasteners; threaded parts and threaded connections.

FOR MANUFACTURERS
DESIGNERS
INSPECTORS
TOOL ENGINEERS
LABORATORIES and for
PRODUCT CONTROL
in assembly.

Capacities: (0-200 in. lbs.) or (0-150 ft. lbs.)

Write for Bulletin TTF

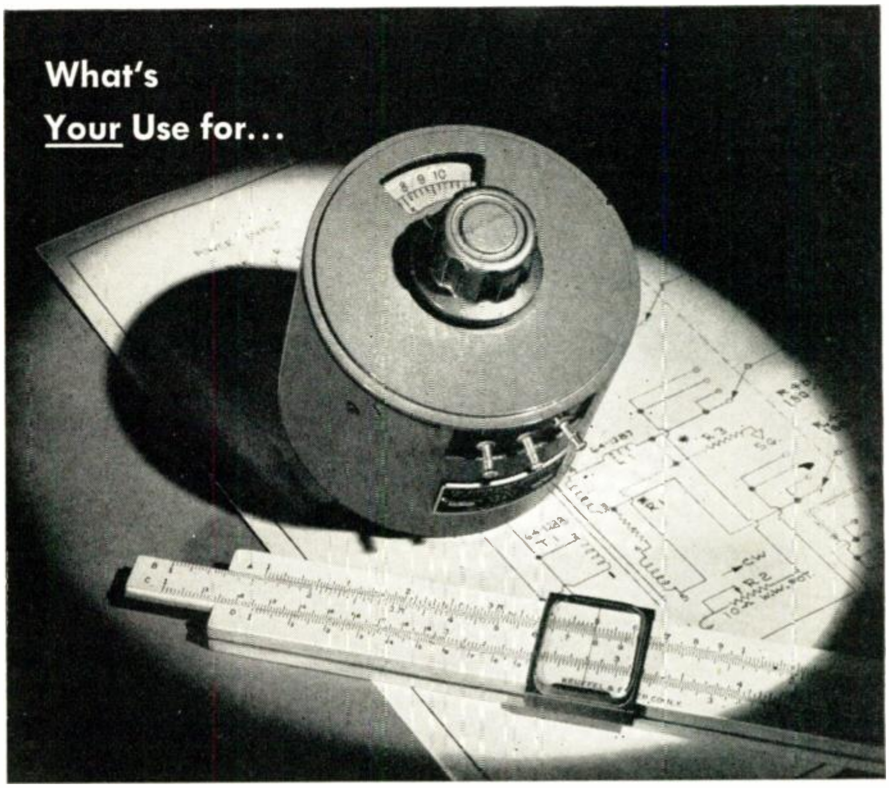
PA STURTEVANT CO
ADDISON QUALITY ILLINOIS

BE SAFE WITH Q-max A-27

LOW-LOSS LACQUER & CEMENT

- Q-Max is widely accepted as the standard for R-F circuit components because it is chemically engineered for this sole purpose.
- Q-Max provides a clear, practically loss-free covering, penetrates deeply, seals out moisture, imparts rigidity and promotes electrical stability.
- Q-Max is easy to apply, dries quickly and adheres to practically all materials. It is useful over a wide temperature range and serves as a mild flux on tinned surfaces.
- Q-Max is an ideal impregnant for "high" Q coils. Coil "Q" remains nearly constant from wet application to dry finish. In 1, 5 and 55 gallon containers.

Communication Products Company, Inc.
MARLBORO, NEW JERSEY
(MONMOUTH COUNTY)
Telephone: FReehold 8-1880



vernistat... *The Revolutionary New Precision Variable-Ratio Transformer*

Analog Computers? Servos? Control Systems? Vernistat is a completely different type of voltage divider combining **low output impedance with an inherently high resolution and linearity** not ordinarily attainable by precision potentiometers.

The Vernistat consists of a tapped auto-transformer which provides the basic division of voltage into several discrete levels. These levels are selected and further sub-divided by a continuous interpolating potentiometer that moves between 30 transformer taps.

Because of its unique operating principles, electrical rotation is held to close tolerances eliminating the need for trim resistors. In many applications there is also no need for impedance matching amplifiers.

Specifications of the standard model Vernistat are shown below. Other versions are under development to meet specific end uses.

What are your requirements for this unique precision voltage divider? Fill in the coupon now.

vernistat division PERKIN-ELMER CORPORATION
NORWALK, CONNECTICUT

SPECIFICATIONS

Linearity Tolerance	better than ±.05%
Resolution	better than .01%
Output Impedance	130 ohms (max.)
Max. Output Current	50 ma
Frequency	50-3000 cps

Other models including a miniaturized 400 cps version will be available in the near future.

vernistat division PEC PERKIN-ELMER CORPORATION
814 Main Avenue, Norwalk, Connecticut

Send me more information on the Vernistat.
The application I have in mind is as follows: ....

NAME ....
TITLE ....
COMPANY ....
ADDRESS ....

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(*Continued from page 168A*)

frequency range. The output amplitude is adjustable continuously by a logarithmical output level control with a scale calibrated in rms volts from 0.01 to 10 maximum. Power output is 100 milliwatts into 1000 ohms.

This stable low distortion oscillator is well suited as a secondary standard in laboratory, production and industrial use. Its uses include numerous special applications, such as narrow band filter alignment and the checking, calibration and adjustment of various instruments.

The unit sells for $950.00, f.o.b., Cambridge, Mass.

Hermetically Sealed Differential Relays

A new line of differential relays which may be used for automatic overload, overvoltage, under-voltage or under-current protection is being offered by **Amperite Co.**, 561 Broadway, New York 12, N. Y. They reset automatically when the abnormal condition is removed.

Since the actuating element is a heater, the relay may be designed for time constants varying from approximately 3 to 30 seconds. It will not respond to transients of shorter duration. Available in SPST—normally open or normally closed. Completely isolated contacts makes many circuit variations possible. The relays can be designed for currents from 10 to 1000 ma., voltages from 1 to 100 v. Higher currents can be handled by an external contactor. Approximately 1 watt is required for the heater. However, the relay will withstand 100 per cent overload indefinitely.

It is available in either the standard octal base or nine-pin miniature. They are pre-set at the factory. Standard tolerance of voltage or current for opening and closing is ±10 per cent. Closer tolerances are available on special order. The relay is compact, light, rugged. Being hermetically sealed, they are not affected by changes in temperature, pressure, or humidity. List price is $4.00 each. In Canada write to: Atlas Radio Corp., 560 King St., W, Toronto 2B, Ontario.

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" paraboloid 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 A1A, 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/2A, 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 881	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 LO-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-345	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 CL-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

NFI-I Noise Filters, Rated 100A/35 volts DC. AAF #3300-398857005, Mallory ....79¢

POWER SUPPLY and modulator, MP-28, For TA-12 transmitter. Dynamotor rated: Input: 28 VDC/14.5A. 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 111, 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 MI-2. A lattice of 3 crystals in a disk-like structure; 17 37 Kc. High imp. $27.50

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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

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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

SA1A/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

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TYPE	INPUT VOLTS	INPUT AMPS	OUTPUT VOLTS	OUTPUT AMPS	Price
35X.059	19	3.8	405	.095	$4.35
POSX-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. † Used, Excellent. * Replacement for PE 94.

PE 94-, Brand New 5.95

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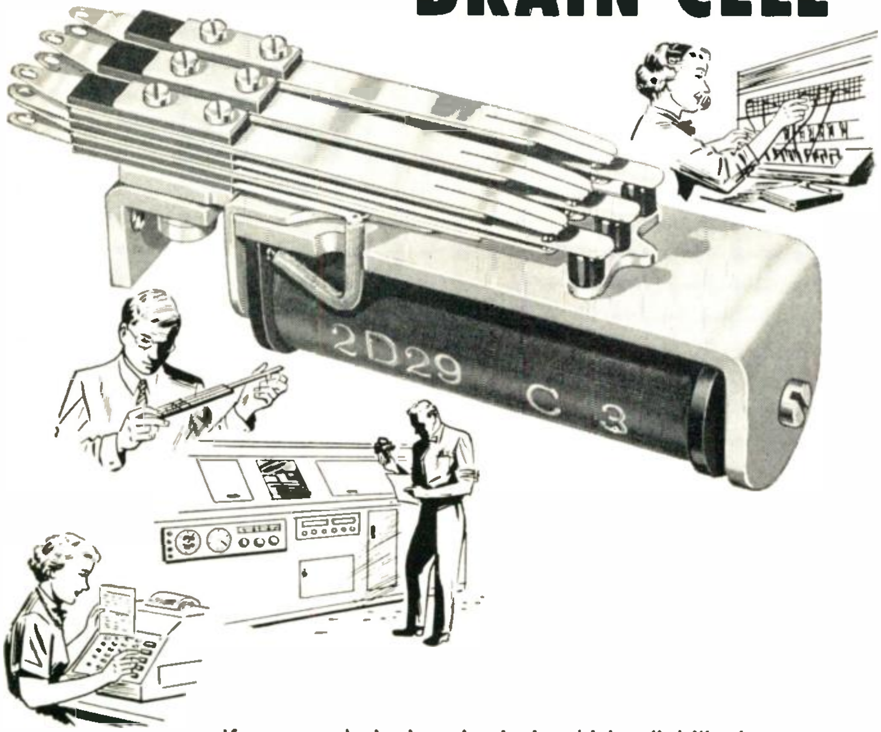

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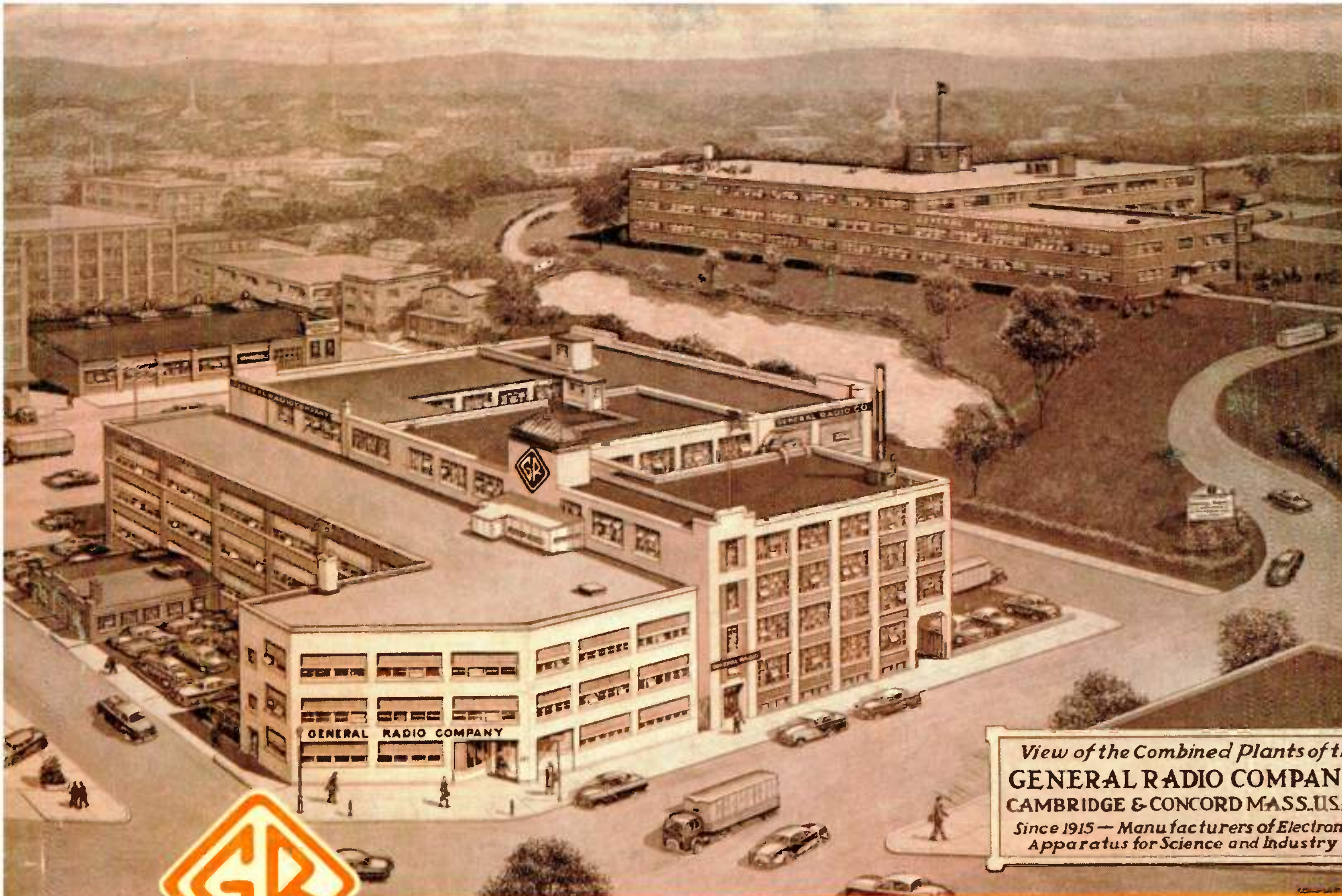

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