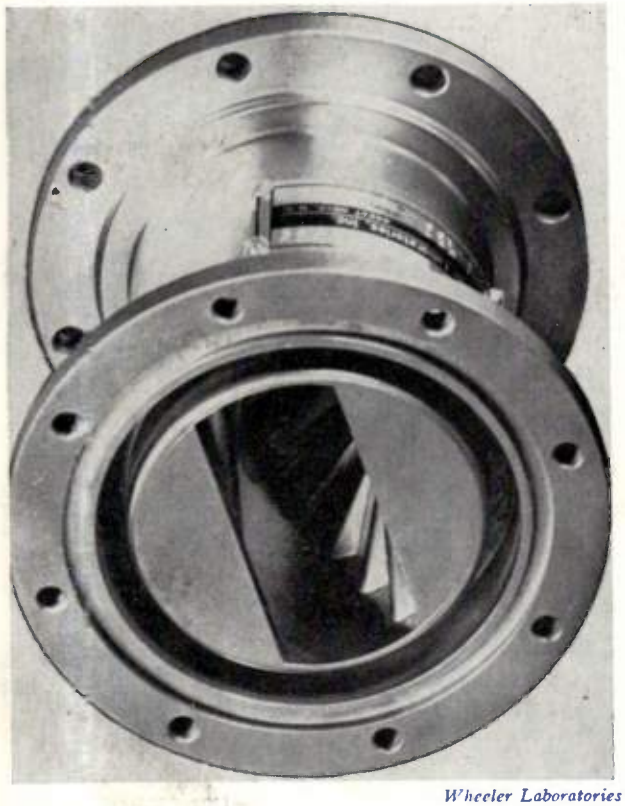
FEBRUARY · 1954

Proceedings of the I·R·E

A Journal of Communications and Electronic Engineering

STEP-TWIST WAVEGUIDE



Wheeler Laboratories

One unit of a complete series of short 90-degree step-twist waveguide components designed for Signal Corps Engineering Laboratories in their program of component development.

Volume 42 Number 2

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The Institute of Radio Engineers

NEW "M" TYPE TOROIDS Maximum Q Minimum Size

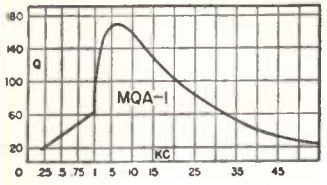
UTC Permalloy Dust Toroids have been the standard of the industry for over 15 years. The **MQ series** of coils provide the highest Q factor in their class (see curves below), with miniaturized dimensions. **All units are hermetically sealed** to MIL-T-27 Specifications.

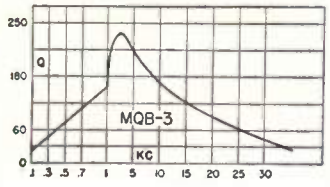
The **stability** is excellent. For the MQE-7 the inductance change is less than 1% for voltages from .1 to 3 volts. The MQA-13 change is less than 1% for applied voltages from .1 to 20 volts. The MQB-5 change is less than 1% for applied voltages from .1 to 50 volts. DC is permissible through the coil (values listed below). Inductance is virtually independent of frequency temperature and vibration.

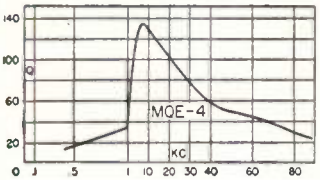
Hum pickup is extremely low due to the toroidal winding structure, with windings uniformly spread over the core. The case is of high permeability, affording additional shielding such that close spacing of units can be effected, the coupling attenuation being approximately 80 DB.

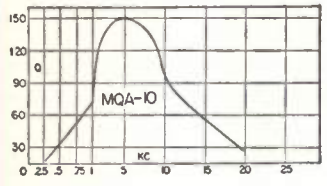
Other values of inductance than those listed are available on special order at the price of the next higher listed value.

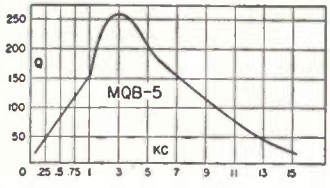
TYPICAL Q CURVES

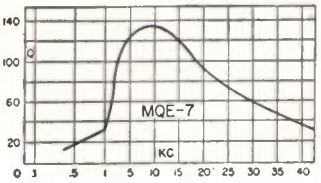


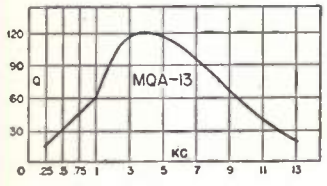


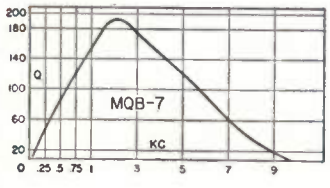


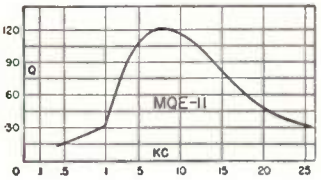


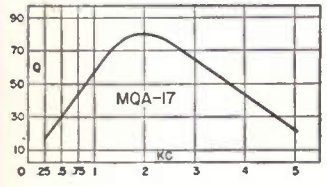


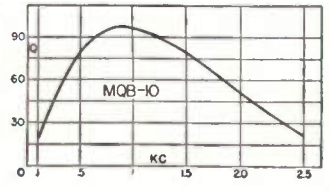


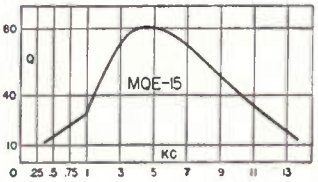












MQA TYPES

Type No.	Inductance	*DC Max.
MQA-1	7 mhy.	250
MQA-2	12 mhy.	200
MQA-3	20 mhy.	150
MQA-4	30 mhy.	125
MQA-5	50 mhy.	100
MQA-6	70 mhy.	80
MQA-7	120 mhy.	60
MQA-8	.2 hy.	50
MQA-9	.3 hy.	40
MQA-10	.5 hy.	30
MQA-11	.7 hy.	25
MQA-12	1 hy.	20
MQA-13	1.5 hy.	17
MQA-14	2.5 hy.	13
MQA-15	4 hy.	10
MQA-16	6 hy.	9
MQA-17	10 hy.	7
MQA-18	15 hy.	5
MQA-19	22 hy.	4

*This value of D.C. (MA) will drop the coil inductance 5%. Values of D.C. below this will show proportionately (linear) less inductance drop. For example, MQE-1 will drop ½% in L with 13.5 MA.

MQB TYPES

Type No.	Inductance	*DC Max.
MQB-1	10 mhy.	400
MQB-2	30 mhy.	250
MQB-3	70 mhy.	170
MQB-4	120 mhy.	120
MQB-5	.5 hy.	60
MQB-6	1 hy.	40
MQB-7	2 hy.	30
MQB-8	3.5 hy.	22
MQB-9	7.5 hy.	16
MQB-10	12 hy.	11
MQB-11	18 hy.	9
MQB-12	25 hy.	8

MQE TYPES

Type No.	Inductance	*DC Max.
MQE-1	7 mhy.	135
MQE-2	12 mhy.	100
MQE-3	20 mhy.	80
MQE-4	30 mhy.	65
MQE-5	50 mhy.	50
MQE-6	70 mhy.	40
MQE-7	100 mhy.	35
MQE-8	150 mhy.	30
MQE-9	.25 hy.	22
MQE-10	.4 hy.	17
MQE-11	.6 hy.	14
MQE-12	.9 hy.	12
MQE-13	1.5 hy.	9
MQE-14	2 hy.	8
MQE-15	2.8 hy.	7.2

MQE CASE

Length ..... 1 1/16"
Width ..... ½"
Height ..... 1 1/32"
Unit Weight ..... 5 oz.

MQA CASE

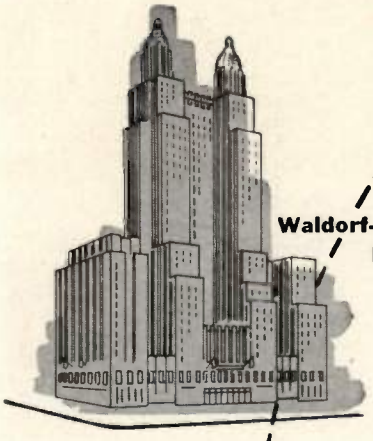
Length ..... 1 1/32"
Width ..... 1/16"
Height ..... 1 1/32"
Unit Weight ..... 1 oz.

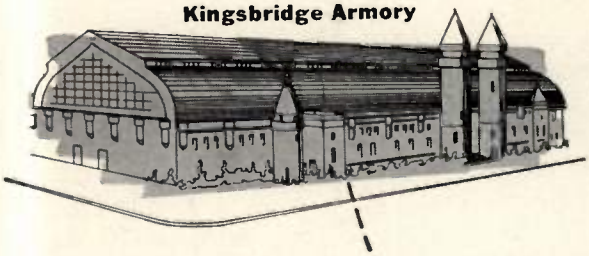
MQB CASE

Length ..... 2 7/16"
Width ..... 1 3/16"
Height ..... 2 3/16"
Unit Weight ..... 2 oz.

United Transformer Co.
150 VARICK STREET NEW YORK 13, N. Y.
EXPORT DIVISION: 13 EAST 40th STREET, NEW YORK 16, N. Y.. CABLES: "ARLAB"

The eye-opening event of the radio-electronic new year!





For the past 12 months the vast, fast-growing radio-electronic industry has been preparing for 4 great days — March 22-25. This is when the IRE National Convention and Radio Engineering Show — the biggest and best ever — will take place in New York City. Be sure to join the other radio-electronic men — nearly 40,000 are expected — who will come, see and appraise the show at which all that is new will be unveiled.

A practical summary of radio-electronic progress will be unfolded at 54 technical sessions during the four-day period. 243 scientific and engineering papers, grouped by related interests, will be presented during these sessions, more than half of which are organized by IRE professional groups. Actually, you will be attending 21 conventions fused into one. New York's finest meeting facilities are provided — the Waldorf-Astoria Hotel plus 3 huge halls in Kingsbridge Armory. Transportation between the two locations is quick, easy — by subway and bus service.

At the show you will find over 600 firms **"spotlighting the new"** in their high-interest product exhibits. These will extend over a mile and a half along avenues appropriately named for radio elements: "Instruments," "Components," "Airborne," "Radar," "Transistor," "Audio," "Microwave," etc. These exhibits, an education and revelation in themselves, fill the four-acre space of the great Kingsbridge Armory ... and can be viewed throughout any one or all of the four days.

Admission is by registration only, and serves for the four-day period. For IRE members the cost is only $1.00. For non-members it is a low $3.00, covering sessions and exhibits. Social events have been carefully planned. These are priced separately.

March 22-25, 1954

is the date! New York is the city where the radio-electronic event of the year will take place. Come! See! Enjoy!

THE 1954
IRE NATIONAL CONVENTION
AND
RADIO ENGINEERING SHOW

THE INSTITUTE
OF RADIO ENGINEERS

1 EAST 79th STREET, NEW YORK CITY

Proceedings of the I.R.E. October, 1953, Vol. 41, No. 10. 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 $1.00; non-members $2.25. Yearly subscription price: to members $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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NOW I-T-E quality
I.F. and R.F. transformers and coils

Custom built to your specifications

Long noted for top-quality wire-wound components—precision resistors, power resistors, deflection yokes, and focus coils—I-T-E now adds I.F. and R.F. transformers and coils to its line.

Coils or complete transformers—the simplest to the most complex—are precisely fabricated to your specifications. Versatile coil-winding machinery plus latest-type testing equipment assure close electrical and mechanical tolerances. Components are sturdy—built to "take it". They're stable over time, temperature variation, and in humid atmospheres.

Take advantage of I-T-E engineering skill, coil-winding experience, and modern facilities. We will build to your particular specifications . . . and, at a competitive price!

all types of I.F. and R.F. coils • R.F. output transformers • antenna transformers • any stage tuners • buffers • doublers • mixers • R.F. chokes • linearity coils • peaking coils

Get your copy of Catalog R-200 P. It gives complete information about I-T-E wire-wound products.

Write to *Resistor Division, Advt. Dept., I-T-E Circuit Breaker Co., 1924 Hamilton St., Phila. 30, Pa.*

I-T-E **WIRE-WOUND PRODUCTS**

Meetings with Exhibits

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

Δ

February 4, 5 & 6, 1954
Sixth Southwestern I.R.E. Conference & Electronics Show, Tulsa, Oklahoma
Exhibits: John H. Hidy, P.O. Box 5121, Tulsa, Okla.

Δ

February 11, 12, 1954
Western Computer Conference and Exhibit, Ambassador Hotel, Los Angeles, California
Exhibits: Norman Loretz, Magnavox Research Laboratories, 2255 Carmelina Ave., West Los Angeles, Calif.

Δ

March 22, 23, 24 & 25, 1954
Radio Engineering Show and I.R.E. National Convention, New York City. Exhibits at Kingsbridge Armory and headquarters hotel Waldorf Astoria. Direct connections by subway and by private express busses.
Exhibits Manager: Wm. C. Copp, 1175 Broadway, New York 36, N.Y.

Δ

April 24, 1954
Eighth Spring Technical Conference, Cincinnati, Ohio
Exhibits: Mr. Alonzo M. Boothe, 1338 Grace Ave., Cincinnati 8, Ohio

Δ

May 5, 6 & 7, 1954
Technical Conference with Electronic Exhibits, I.R.E. Region Seven, Multnomah Hotel, Portland, Ore.
Exhibits: W. K. Dallas, P.O. Box 831, Portland 7, Ore.

Δ

May 7, 8, 1954
NEREM, New England Radio Engineering Meeting, Sheraton-Plaza Hotel, Boston, Mass.
Exhibits Chairman: Robert A. Waters, 4 Gordon St., Waltham 54, Mass.

Δ

May 10, 11, 12, 1954
Airborne Electronics Conference, Dayton Biltmore Hotel, Dayton, Ohio
Exhibits: R. J. McIlrath, 410 West First Street, Dayton 2, Ohio

Δ

August 25, 26 & 27, 1954
Western Electronic Show & Convention, Pan-Pacific Auditorium, Los Angeles, Calif.
Business Manager: Mr. Mal Mobley, Jr., 344 North La Brea Ave., Los Angeles, Calif.

Δ

October 4, 5, 6, 1954
National Electronics Conference, Sherman Hotel, Chicago, Ill.
Exhibits: Mr. George H. Wise, c/o DeVry Technical Institute, 4141 West Belmont, Chicago 41, Ill.

Δ

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 such course listings are free to IRE Professional Groups.

HIGH VOLTAGE

molded ceramic filter CAPACITORS

Specifically engineered for reliable service in the high voltage supply filter circuits of modern television receivers and cathode ray instruments are Sprague's new molded jacket "doorknob" capacitors.

These moderately priced units incorporate an improved ceramic dielectric element encased in a thermosetting, non-flammable housing for maximum protection. Fifteen different terminal combinations are standard to meet practically every mounting requirement.

Standard capacitance rating is 500 mmf. Voltages are 30,000, 25,000, and 20,000 volts d-c to fit all applications in television receivers from 27-inch down to 17-inch screen size.

Complete engineering information on these capacitors is contained in Bulletin 606A, available on letterhead request to Sprague Electric Company, 235 Marshall Street, North Adams, Massachusetts.

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

SPRAGUE

WORLD'S LARGEST CAPACITOR MANUFACTURER

EXPORT FOR THE AMERICAS: SPRAGUE ELECTRIC INTERNATIONAL LTD., NORTH ADAMS, MASS. CABLE: SPREXINT

This rubber "tire" talks to you

Bell Laboratories engineers have developed a new and highly economical way to record sound magnetically.

Instead of tape or wire they use a mixture of rubber and iron oxides which is formed into a band and mounted on a wheel. This simple and very rugged "talking rubber" can play back messages clearly millions of times.

Talking rubber is already at work for the Bell System announcing weather and answering customers who call vacant or disconnected numbers. It promises to have many other uses. In a new machine, it answers your telephone in your own voice when you are away—and takes a message for you in the voice of your caller.

Many businesses, too, other than telephone, are expected to find a variety of ways to use talking rubber—especially whenever a message must be given quickly to many people.

Talking rubber proves again the downright practicality of Bell Laboratories' research to improve telephone service.

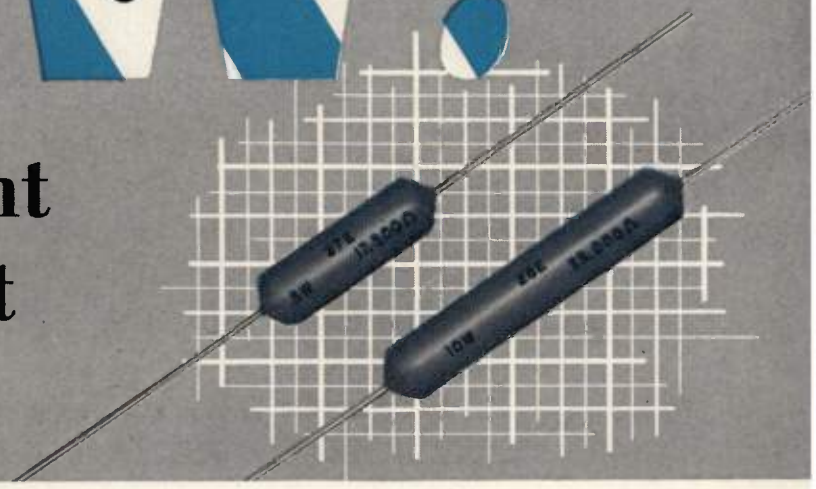
One of a bank of recorder-reproducers operated by the New York Telephone Company for the New York Stock Exchange. They give instant stock quotations to brokers who dial a code number. Recording and pickup heads are shown above wheel.

BELL TELEPHONE LABORATORIES

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

NEW!

for convenient point-to-point wiring . . .



MINIATURIZED 5 AND 10 WATT AXIAL LEAD WIRE-WOUND RESISTORS

Here are two *truly miniaturized* self-mounting wire-wound power resistors to simplify your TV and industrial electronic production where space is a factor. They're ideal for point-to-point wiring, terminal board mounting, and processed wiring boards, where they fit in admirably in dip-soldered subassemblies.

Axial lead Blue Jackets are rugged vitreous enamel power resistors built to withstand the severe humidity performance requirements of RETMA and MIL Specifications. They have been field-tested for a full year—*successfully!* As for *economy*, these newest members of the Sprague Blue Jacket family are low in cost . . . eliminate need for extra hardware . . . save time and labor in mounting!

You can get these outstanding new Blue Jacket Resistors without delay in any quantity you require. Sprague Engineering Bulletin 111 gives full data on these and all other commercial Blue Jacket Resistors. Send for your copy.

SPRAGUE ELECTRIC COMPANY
235 Marshall Street, North Adams, Mass.

SPRAGUE TYPE NO.	WATTAGE RATING	DIMENSIONS L (inches)	D	MAXIMUM RESISTANCE
27E	5	1⅛	5/16	17,500 Ω
28E	10	1⅞	5/16	35,000 Ω

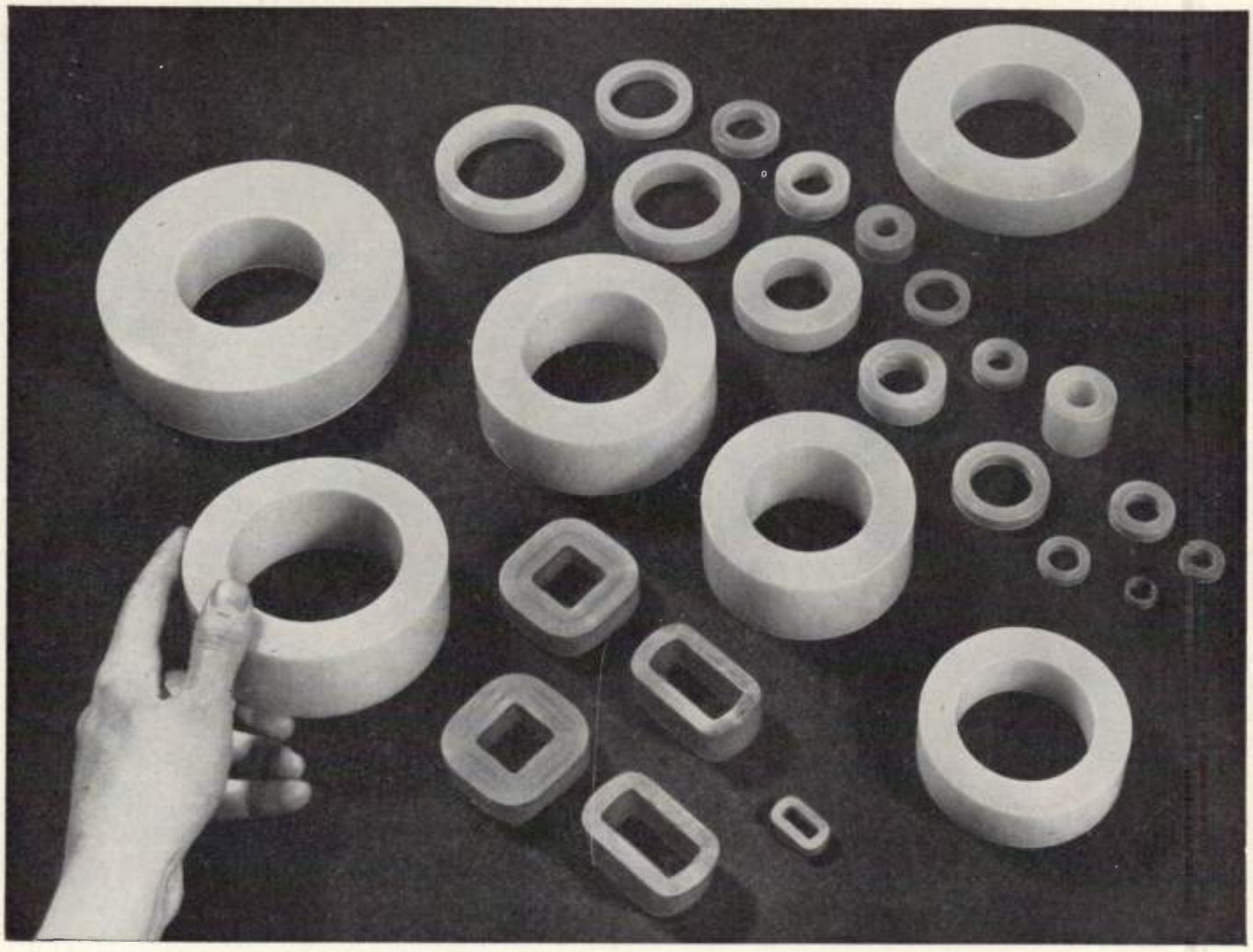
Standard Resistance Tolerance: ±5%

SPRAGUE Blue Jackets®

PIONEERS IN ELECTRIC AND ELECTRONIC DEVELOPMENT

NORTH ADAMS, MASSACHUSETTS

EXPORT FOR THE AMERICAS: SPRAGUE ELECTRIC INTERNATIONAL LTD., NORTH ADAMS, MASS. CABLE: SPREXINT



IN TAPE-WOUND CORES

JUST NAME YOUR REQUIREMENTS!

RANGE OF MATERIALS

Depending upon the specific properties required by the application, Arnold Tape-Wound Cores are available made of DELTAMAX . . . 4-79 MO-PERMALLOY . . . SUPERMALLOY . . . MUMETAL . . . 4750 ELECTRICAL METAL . . . or SILECTRON (grain-oriented silicon steel).

RANGE OF SIZES

Practically any size Tape-Wound Core can be supplied, from a fraction of a gram to several hundred pounds in weight. Toroidal cores are made in twenty-two standard sizes with protective nylon cases. Special sizes of toroidal cores—and all cut cores, square or rectangular cores—are manufactured to meet your individual requirements.

RANGE OF TYPES

In each of the magnetic materials named, Arnold Tape-Wound Cores are produced in the following standard tape thicknesses: .012", .008", .004", .002", .001", .0005", or .00025", as required.

Applications

Let us help with your problems of cores for Magnetic Amplifiers, Pulse Transformers, Current Transformers, Wide-Band Transformers, Non-Linear Retard Coils, Peaking Strips, Reactors, etc.

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Announcing..

one compact instrument that measures...

Frequency from 10 cps to 200 mc

Time interval from 1 μsec to 100 days

Period from 0 cps to 10 kc

instantly
automatically
directly

the completely new, completely different...

-*hp*- 525A Frequency Converter
10 to 100 mc

-*hp*- 525B Frequency Converter
100 to 200 mc

-*hp*- 526A Video Amplifier
sensitivity 10 mv

-*hp*- 526B Time Interval Unit
Interval 1 μsec to 100 days

...524B ELECTRONIC COUNTER

with plug-in units to fit your exact measuring need.

Direct, instantaneous, automatic readings

Easily used by non-technical personnel

Resolution 0.1 μsec; accuracy 1/1,000,000 ± 1 count

High sensitivity, impedance, extreme reliability

No calculation, interpolation or complex setup

Quality components; compact, military design

Automatic illuminated decimal point

Why pay for more instrumentation than you need? The revolutionary new all-purpose -*hp*- 524B Electronic Counter gives you *exactly* the frequency, time interval or period measuring coverage you want now. Later, you can add inexpensive -*hp*- Plug-In Units, and double or triple the usefulness of this outstanding instrument.

-*hp*- 524B Counter gives you more range, simplicity, usefulness, convenience and reliability, smaller size and lower cost than any comparable commercial instrumentation ever offered. With this one complete instrument, you can measure transmitter and crystal oscillator frequencies, electronic, electrical and mechanical time intervals, pulse lengths and repetition rates, frequency drift; make high accuracy ballistics time measurements or high resolution tachometry measurements. The instrument is also an ideal precision frequency standard, giving convenience and flexibility not provided in the usual primary standard. It is simple to operate and readily used by non-technical personnel.

hp INSTRUMENTS | COMPLETE COVERAGE

Frequency from 10 cps to 10 mc is read over 5 selected periods—0.001, 0.01, 0.1, 1, and 10 seconds. Display time is variable at will, counts are automatically reset, and action is repetitive. When set for low frequency period measurement, the duration of 1 low frequency cycle is measured in microseconds. A 10 cycle sample is taken to determine this period. Periods may be displayed repetitively or "held" (as in frequency counting). Results are measured instantly and automatically, and presented in direct-reading form with automatic illuminated decimal point.

DETAILS OF COUNTER PLUG-IN UNITS

Addition of -*hp*- Plug-In Units will extend the Counter's frequency range to 200 mc, provide increased sensitivity and a high-impedance pickup probe, and make available uniquely flexible time interval circuits that may be started and stopped by any electrical impulse.

-*hp*- 525A FREQUENCY CONVERTER. Extends Counter's 10 mc direct-reading range in decade steps to 100 mc. Maintains Counter accuracy throughout extended range. Provides additional amplification to increase video sensitivity to 0.1 volt through Counter's basic 10 cps to 10 mc range.

-*hp*- 525B FREQUENCY CONVERTER. Similar to 525A Converter but extends Counter's range from 100 mc to 200 mc in 10 mc steps. Maintains same high accuracy throughout range, and provides high sensitivity for measurement of low level circuits.

-*hp*- 526A VIDEO AMPLIFIER. Increases 10 cps to 10 mc sensitivity of Counter to 10 millivolts for frequency measurement at low level. Special probe assembly simplifies remote pick-up at high impedance levels.

-*hp*- 526B TIME INTERVAL UNIT. Measures interval 1.0 microseconds to 100 days with accuracy of 0.1 μsec $\pm$ 0.001%. Intervals are read direct in seconds, milliseconds or microseconds. Start and stop triggering is performed in either common or separate channels, and may be accomplished through the use of positive or negative going waves. Trigger voltage levels are continuously adjustable from —192 to +192 volts.

SPECIFICATIONS

BASIC 10 MC COUNTER WITHOUT PLUG-IN UNITS

FREQUENCY MEASUREMENT

Range: 10 cps to 10 mc.
Gate Time: 0.001, 0.01, 0.1, 1, 10 seconds or manual control.
Accuracy: $\pm$ 1 count $\pm$ stability (see below).
Reads In: kc with automatic decimal point.

PERIOD MEASUREMENT

Range: 0 cps to 10 kc.
Gate Time: 1 or 10 cycles of unknown.
Accuracy: $\pm$ 0.3% (measurement one period).
$\pm$ 0.03% (ten period average).
Standard Frequency Counted: 10 cps; 1 or 100 kc; 10 mc; or externally applied frequency.
Reads In: Seconds, milliseconds or microseconds with automatic decimal point.

GENERAL

Registration: 99,999,999; 8 places. First 6 places on neon lamp decades; last 2 meters.
Stability: 1/1,000,000 short-term; 2/1,000,000 per week. May be standardized against WWV; or used with external 100 kc primary standard for higher accuracy.
Display Time: Variable 0.1 to 10 seconds in steps of gate time selected. Display can be held indefinitely.
Output Frequencies: Secondary standard frequencies available at front panel: 10 cps, 1 kc rectangular; 100 kc positive pulse; 10 mc sine wave. (Stability as above.)
Self-Check: Panel control provides automatic count of internal standard 100 kc and 10 mc frequencies to insure accuracy of gate and proper operation of counters.
Input Voltage: 1 v R.M.S. minimum.
Input Impedance: Approx. 1 megohm, 40 $\mu\mu$fd shunt.
Connectors: BNC type.
Power Supply: 115 $\pm$ 10%, 50-1,000 cps, approx. 500 watts.
Cabinet Size: 19" wide, 19¼" high, 17" deep. Supplied for rack mounting.
Weight: 112 lbs. net, 175 lbs. packed.
Accessories Furnished: -*hp*- AC-16D cable assembly, 42" RG-58/U cable terminated one end with UG-88/U BNC connector; -*hp*- 61B-16H power cable.
Price: $1,890.00 f.o.b. factory.

-*hp*- 526A VIDEO AMPLIFIER

plugged into -*hp*- 524B Counter

Range: 10 cps to 10 mc.
Input Voltage: 10 mv R.M.S. minimum.
Level Control: Meter indicates signal level and correct adjustment.
Output Terminal: BNC connector provides 10 times input voltage into 50 ohm or higher load. Allows oscilloscope monitoring of input signal without loading circuit.
Accessories Furnished: -*hp*- 526D-16A probe assembly—input impedance 15 $\mu\mu$f and 10 megohms; remote signal pickup at 0.1 R.M.S. minimum. Cables and connectors.
Size: Supplied in aluminum storage case, with carrying handle, case 12" wide, 9" high, 8" deep. (See photo.)
Weight: 10 lbs. net, 19 lbs. packed.
Price: $125.00 f.o.b. factory.

-*hp*- 525A FREQUENCY CONVERTER

plugged into -*hp*- 524B Counter

Range: As amplifier: 10 cps to 10 mc. As converter: 10 mc to 100 mc.
Accuracy: $\pm$ 1 cps $\pm$ stability (see General).
Registration: 9 places: first place indicated on panel selector switch labeled 0, 10, 20 . . . 90; next 8 as indicated under General.
Input Voltage: 0.1 v R.M.S. minimum, 10 cps to 10 mc; 10 mv R.M.S. minimum, 10 mc to 100 mc.
Input Impedance: Approx. 1 megohm shunted by 40 $\mu\mu$fd, 10 cps to 10 mc; approx. 50 ohms, 10 mc to 100 mc.
Level Control: Tuning eye aids frequency selection, indicates correct voltage level adjustment.
Size, Weight: Same as 526A Amplifier.
Price: $225.00 f.o.b. factory.

-*hp*- 525B FREQUENCY CONVERTER

plugged into -*hp*- 524B Counter

Range: 100 mc to 200 mc.
Accuracy: $\pm$ 1 cps $\pm$ stability (see General).
Registration: 10 places; first two places indicated on panel selector switch labeled 100, 110, 120 . . . 190, next eight as indicated under General.
Input Voltage: 0.25 R.M.S. minimum.
Input Impedance: Approximately 50 ohms.
Level Control, Size, Weight: Same as 525A above.
Price: $225.00 f.o.b. factory.

-*hp*- 526B TIME INTERVAL UNIT

plugged into -*hp*- 524B Counter

Range: 1 μsec to 10^7 seconds.
Accuracy: $\pm$ 1/standard frequency counted $\pm$ stability (see General).
Registration: Same as indicated under General.
Input Voltage: 1 v peak minimum, direct-coupled input.
Input Impedance: Approx. 1 megohm, 40 $\mu\mu$fd shunt.
Start and Stop: Independent or common channels.
Trigger Slope: Positive or negative on start and/or stop channels.
Trigger Amplitude: Continuously adjustable, both channels, —192 to +192.
Standard Frequency Counted: 10 cps; 1 or 100 kc; 10 mc or externally applied frequency.
Reads In: Seconds, milliseconds or microseconds with automatic decimal point.
Accessories Furnished: -*hp*- AC-16D cable assembly, 42" RG-58/U cable terminated on one end with UG-88/U BNC connector.
Size, Weight: Same as 525A and 525B.
Price: $150.00 f.o.b. factory.

Data subject to change without notice.

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The "skin" we love to watch

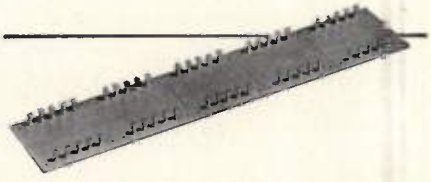
The "skin," or plated coating, on CTC terminals gets extremely close scrutiny from our quality control engineers. And we take pleasure in this careful watching because —

We know, as a result, that you can depend on CTC terminals for electroplated coatings of guaranteed minimum thickness — whether to government specifications or your own.

Our "watching" of these coatings includes periodic bend tests for adhesion, and periodic microscopic inspection of cross sections for coating thickness. These are but two of many examples of quality control that enable us to offer customers *guaranteed* electronic components . . . custom or standard.

Besides terminals, we pay close attention to the production of CTC terminal boards, capacitors, swagers, hardware, insulated terminals, coil forms and coils. For all specifications and prices, write to Cambridge Thermionic Corporation, 456 Concord Avenue, Cambridge 38, Mass. West Coast Manufacturers contact: E. V. Roberts, 5068 West Washington Blvd., Los Angeles 16 and 988 Market St., San Francisco, California.

Terminal Data: Our standard terminal line includes 30 types, each in varied shank lengths. Made of silver plated brass, coated with water dip lacquer to keep them chemically clean for soldering. Also available: combination screw and solder terminals in 3 sizes, and a complete line of phenolic and ceramic insulated terminals. All materials, processes and finishes meet applicable government specifications. Special order finishes include hot tin, electrotin, cadmium plate or gold plate.



Standard CTC Terminal Boards as well as those made to your own specifications by CTC are available. Standard in cotton fabric phenolic, rayon phenolic or grade L-5 silicone impregnated ceramic. Custom made in cloth, paper phenolic, melamine, or silicone fibreglas laminates, imprinted as required and lacquered or varnished to specifications MIL-V-173 and JAN-T-152.

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Length 49 inches
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Eimac 3K50,000L Klystrons in typical CW operation give 10KW power output with only 10 watts drive

High power, high efficiency, ultra-high frequency Eimac type 3K50,000L klystrons, widely heralded for UHF-TV, are proving outstanding for CW. Typical CW operation of these versatile klystrons shows 40% efficiency while delivering 10 kw output with only 10 watts drive—a power gain of 30 db., or 1000 times. Furthermore service at frequencies above and below the UHF-TV band is being obtained through flexibility provided by the externally tuned cavities of Eimac klystrons.

● ***For information about Eimac type 3K50,000L klystrons contact our Application Engineering department.***

TYPICAL OPERATION
3K50,000L Klystrons

	CW	TV
D-C Beam Current	1.65	2.15 amps
D-C Beam Voltage	15	17.2 kv
Driving Power	10	55* watts
Power Output	10	12* kw
Efficiency	40%	32%*

*Peak synchronizing level (80% of saturation power)

3K50,000L KLYSTRONS FOR UHF-TV

TYPE	FREQUENCIES
3K50,000LA	470-580 mc
3K50,000LF	580-720 mc
3K50,000LK	720-890 mc

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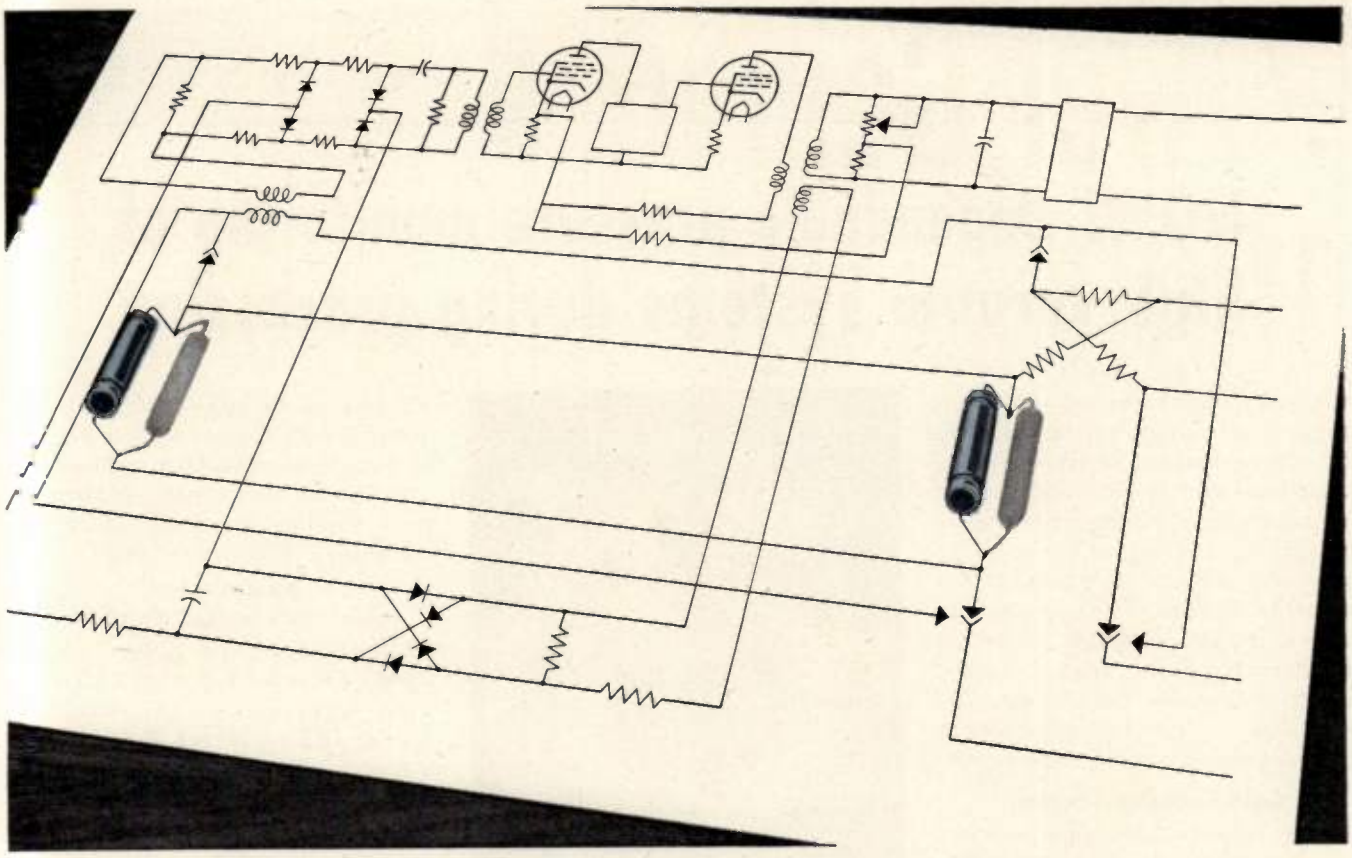
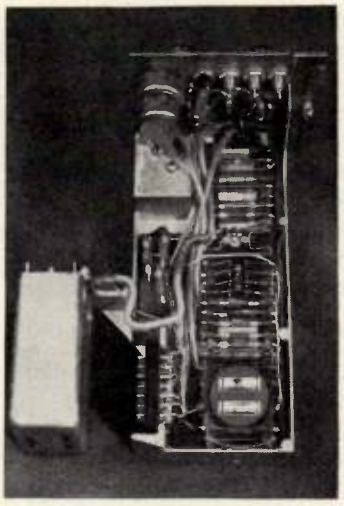


Small G-E Tantalytic Capacitors do big job in new Bell System carrier circuit

The new Bell System N Carrier System is a 12-channel, double-sideband system for single cable application . . . providing low loss, stable, high velocity service for toll and exchange circuits in the range from 15 to 200 miles . . . at a minimum manufactured, installed and maintenance cost. This system requires the use of miniaturized components which will yield large reductions in size and weight yet still give maximum service.

G-E Tantalytic capacitors are a "natural" for the system to handle the job of series d-c blocking, r-c timing and d-c power noise filtering. Recently developed, these polar and non-polar electrolytic capacitors are recommended for virtually all low-voltage d-c applications (ratings from 175 muf at 5 vdc to 12 muf at 150 vdc) where small size, large capacitance, long operating life and long shelf life are major considerations. And since they offer greater capacitance per unit volume than aluminum electrolytics and paper capacitors, they are ideally suited for miniaturized equipment. In some short-time applications, i.e. guided missiles, it is now possible to operate these capacitors in a temperature range from −55 to +110C with proper voltage and life derating.

If your application calls for a small size capacitor with superior performance, it will pay you to investigate the new G-E Tantalytic capacitor. For further information on Tantalytic and other General Electric specialty capacitors for a-c and d-c applications, see your local G-E representative or write for "Tantalytic Capacitors" Bulletin GEC-808 to General Electric Co., Section 442-5, Schenectady 5, New York.



G-E TANTALYTIC capacitors installed in telephone carrier amplifier.

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NOW! Dependable pressure monitoring of high vacuum systems during processing

The new Litton Ionization Gauge is a rugged and completely dependable production tool for monitoring pressures from 10^{-4} to 10^{-7} mm Hg. The instrument is a Philips-type gauge*, specifically engineered for constant production monitoring of high vacuum pressures. It eliminates annoyance and costs of burned-out gauges, activation of poisoned cathodes, heating of grids, etc. Even in steady, day-after-day use, it requires no attention other than a chemical cleaning about twice a year.

Cold Cathode Emitter

The Type L-3032 gauge was developed within Litton Engineering Laboratories to facilitate our own manufacturing of vacuum tubes. It utilizes crossed electric and magnetic fields which enhance collision probability in a small volume so that a cold cathode emitter can be used. Thus operation, even at atmospheric pressure, will not damage the tube. (In normal use, the tube is not operated until black-out of the vacuum system is reached. Good relative pressure readings are available throughout the range of 10^{-4} to 10^{-7} mm Hg.) Type L-3032 tubes have been tested during the past two years on Litton vacuum tube production lines. They are now installed on every exhaust station in our plant.

Monel-Encased

The Ion Gauge Tube is composed of a monel-encased interaction space with the case near ground potential. A nichrome wire anode at 2,500 volts is centered within the case. An outgassing 6.3 volt heater is mounted near the

Type L-3032 Ionization Gauge (above) with adapter for glass systems

monel case, but insulated from it. A ¾" diameter kovar tube, insulated from the monel case by a glass seal, is supplied for connection to the vacuum line. The magnetic field is provided by permanent magnets mounted in a sheet steel shell. This shell also serves as a return magnetic path, connection block, package envelope and oven for the outgassing heater. Electrical connections are made to binding posts on the steel case. The tube weighs but 22 oz. and measures 7" x 5" x 3½". $60.00.

Model 4301 Ionization Gauge Amplifier

This amplifier is a companion instrument for Type L-3032 Ionization Gauge Tube. It includes a range switch for measuring from 10^{-4} to 10^{-8} mm Hg, a special leak-check control providing full scale deflection at any pressure, a zero adjustment control, and a gauge heater supply switch.

Model 4301 Amplifier

It consists of a high voltage rf power supply, a vacuum tube voltmeter circuit with current-sampling resistors, a 6.3-volt transformer (to provide current for the outgassing heater in Type L-3032 Ionization Gauge Tube) and a self-regulating low voltage power supply providing wide input voltage variation without affecting performance. Electrical connection is by cable with banana plugs to Type L-3032 Ion Gauge. Power supply requirements are 110 volts, 60 cps. The instrument measures 10" x 8" x 8". Weight is 17½ lbs. $255.00.

*Licensed under Philips Laboratories, Inc. Patent No. 2197079 Data subject to change without notice. Prices f.o.b. factory.

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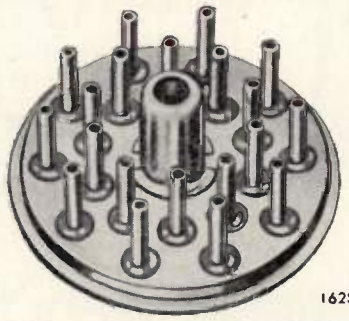

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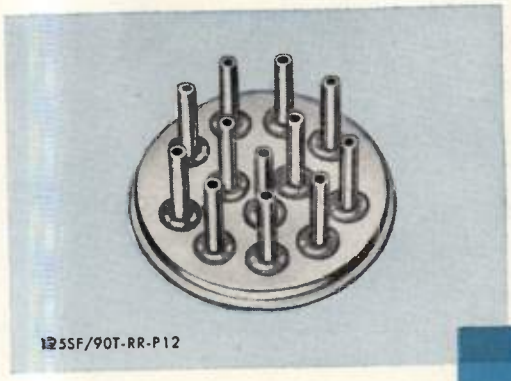

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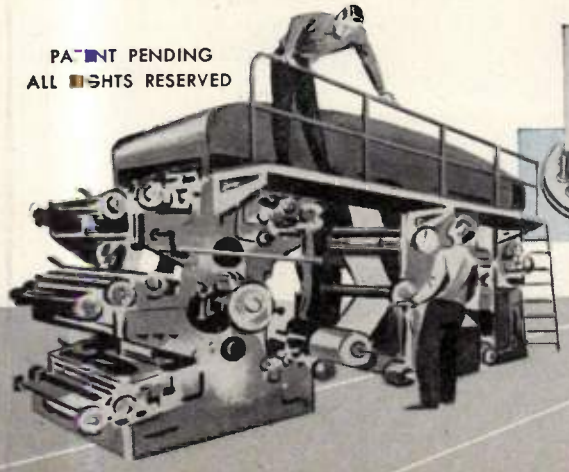

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The only grid-dip meter covering the frequency range of—

2.2 Mc. to 400 Mc.

A multi-purpose instrument for determining the resonant frequency of tuned circuits, antennas, transmission lines. For the measurement of capacitance, inductance, relative "Q"; as an auxiliary signal generator; for signal tracing; as a marker for use with a sweep-frequency generator, and many other applications.

FEATURES:

- **Compact oscillator unit for coupling to circuits in small spaces.**
- **Individually calibrated, direct reading frequency dial; accurate to ±2%.**
- **Internal modulation.**
- **May be battery operated.**

Model 80

STANDARD SIGNAL GENERATOR

Frequency Range: 2 Mc. to 400 Mc.

CRYSTAL CALIBRATOR

Model 111

For the calibration and frequency checking of receivers, transmitters, grid-dip meters, signal generators and other equipment where a high degree of frequency accuracy is required.

250 Kc. to 1000 Mc.

(To within .25 Mc.)

Frequency Accuracy: ±0.002%

The Model 111 is a dual-purpose calibrator. It provides a test signal of crystal controlled frequency and has a self-contained detector of 2 microwatts sensitivity.

Model 78-FM

FM SIGNAL GENERATOR

Frequency Range:
86 Mc. to 108 Mc.

INTERMODULATION METER

Model 31

This instrument will enable you to get the best performance from all audio systems; for the correct adjustment and maintenance of AM and FM receivers and transmitters; for checking linearity of film and disc recordings and reproductions; checking phonograph pickups and recording styli; adjusting bias in tape recordings, etc.

The generator section of the Model 31 produces the mixed high and low frequency signal required for intermodulation testing. A direct reading meter in the analyzer section indicates the percentage of intermodulation.

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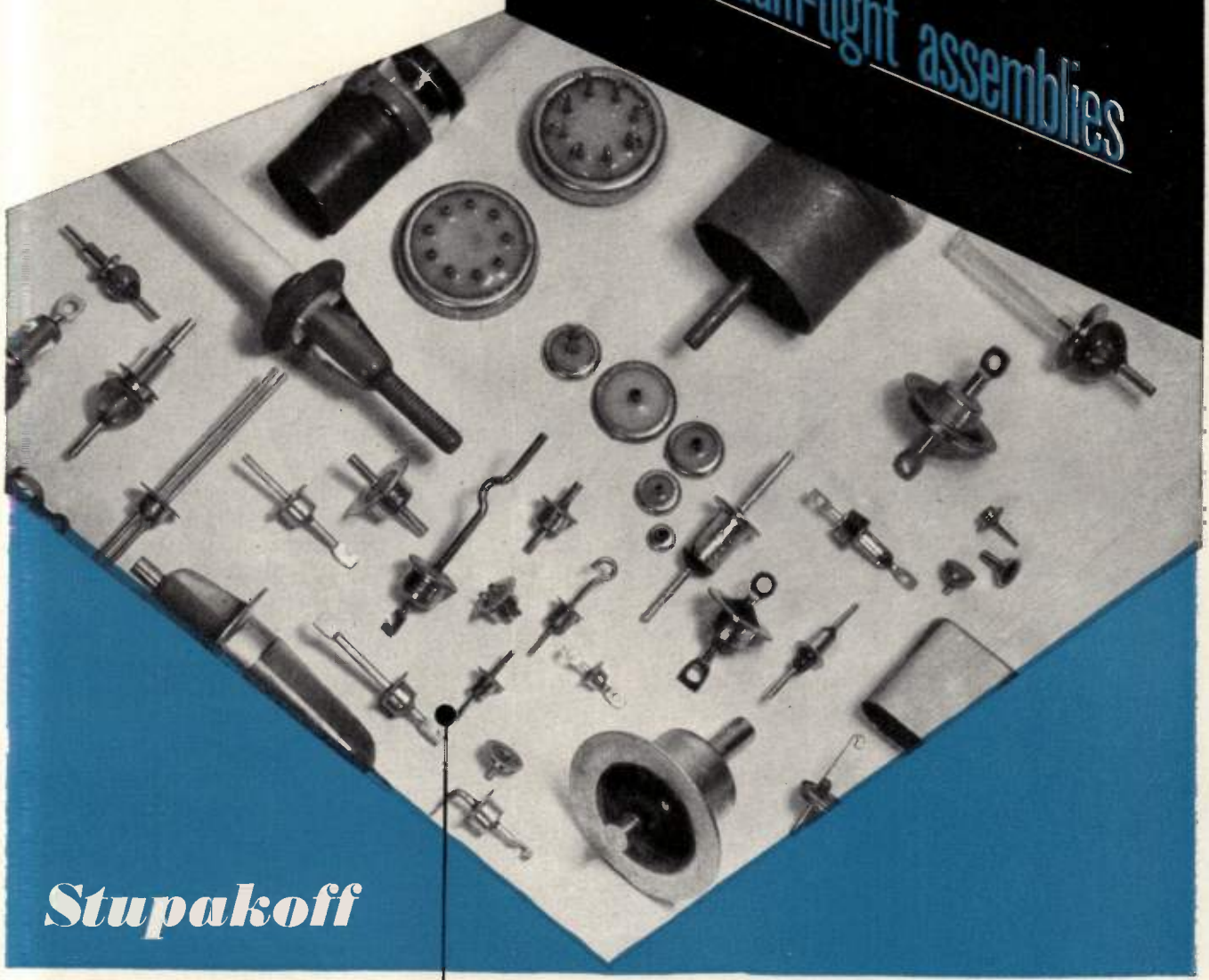

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Industrial Engineering Notes*

Federal Communications Commission

Records of the Federal Communications Commission through November 1953 show that since the freeze was lifted, special temporary authorizations have been issued to 224 outlets. Of these, 206 have been issued since the first of the year, and added to the 108 stations operating at the time the "freeze" was lifted, brings the number of stations authorized to go on the air to 332. The only state now without at least one TV station is Vermont, although service is received from neighboring areas. Of the 224 STAs issued since the lifting of the "freeze" 113 have been for UHF stations and 111 have been for VHF outlets. Four STAs have gone to educational stations. Despite the remarkable pace at which new stations have started operating, less than half of those which received construction permits now are on the air. Since the middle of last year, the FCC reports that it has issued CPs for 493 outlets, including 26 for noncommercial educational stations. Of the total, CPs have gone to 294 UHF stations and 199 VHF stations. Only 25 of these construction permits have been returned—19 UHF and six VHF. As a preview of the demands yet to be made on set and equipment manufacturers, the FCC still has on file 416 applications for TV stations. Some of these are, of course, mutually exclusive and will be eliminated in hearings. Of the pending applications, 97 are for UHF facilities compared with 319 for VHF outlets. Twenty of the applications are for educational stations. . . . **On the international front, regular television service recently was inaugurated in one country and several new stations were added to existing services.** At present there are 17 foreign

(Continued on page 19A)

* The data on which these notes are based were selected by permission from Industry Reports issues of November 27, and December 4, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

Industrial Engineering Notes

(Continued from page 18A)

countries with regular television service. Television broadcasts by the Belgian National Radio started recently in Brussels, according to reports reaching the Department of Commerce. A dual system is being used with an 819-line transmission for French programs and a 625-line transmission for Flemish programs. Two hours of programming per night, six nights a week, are presented in each language. Receivers operating on the dual standard reportedly are being manufactured locally. It is planned to add transmitters at Liége and Antwerp during the early part of this year, and when they are operating, 90 to 95 per cent of the country's population reportedly will be covered. In Japan, the Nippon Television Network Corp. (NTV) now is operating the nation's first commercial station in Tokyo. The station, JOAX-TV, is on the air five hours a day, using a 10-kw transmitter and 510-foot tower. There are an estimated 2,000 TV receivers in the Tokyo area. The government-owned Japanese Broadcasting Corp. (NHK) also has a station, JOAK-TV, on the air in Tokyo and Radio Tokyo is expected to open a station early next year. Service in South America was expanded recently when "Radio Caracas TV" put a signal on the air in Venezuela. The station operates on Channel 7 as the third station in that area. The government-owned "Televisora Naçional" has been operating on Channel 12 for a year, while the commercial outlet "Televisa" has been on the air on Channel 4 since June, 1953. The Swiss government, which has been testing the practicality of TV for about a year, has concluded that service will be possible but faces the problem of financing the proposed stations. This question was slated to be discussed during December when the financial structure of the government's Posts and Telegraph system was to be reviewed. . . . The

(Continued on page 34A)

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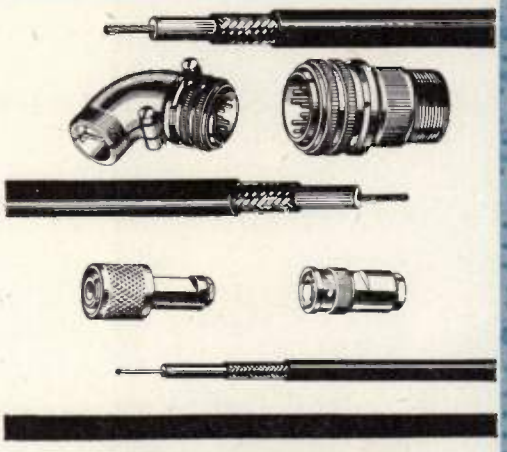

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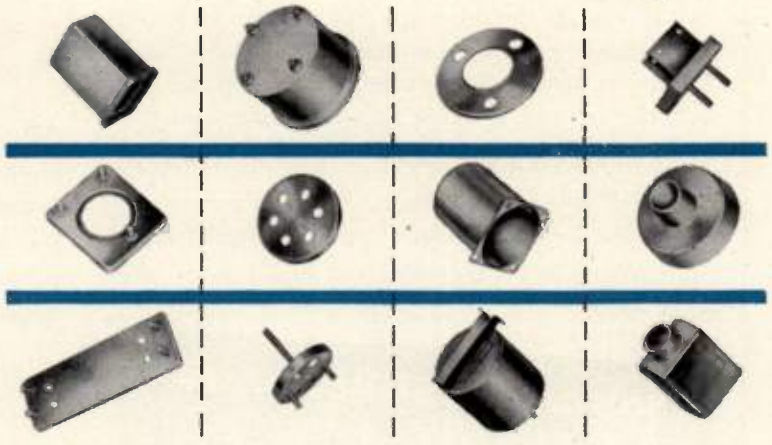

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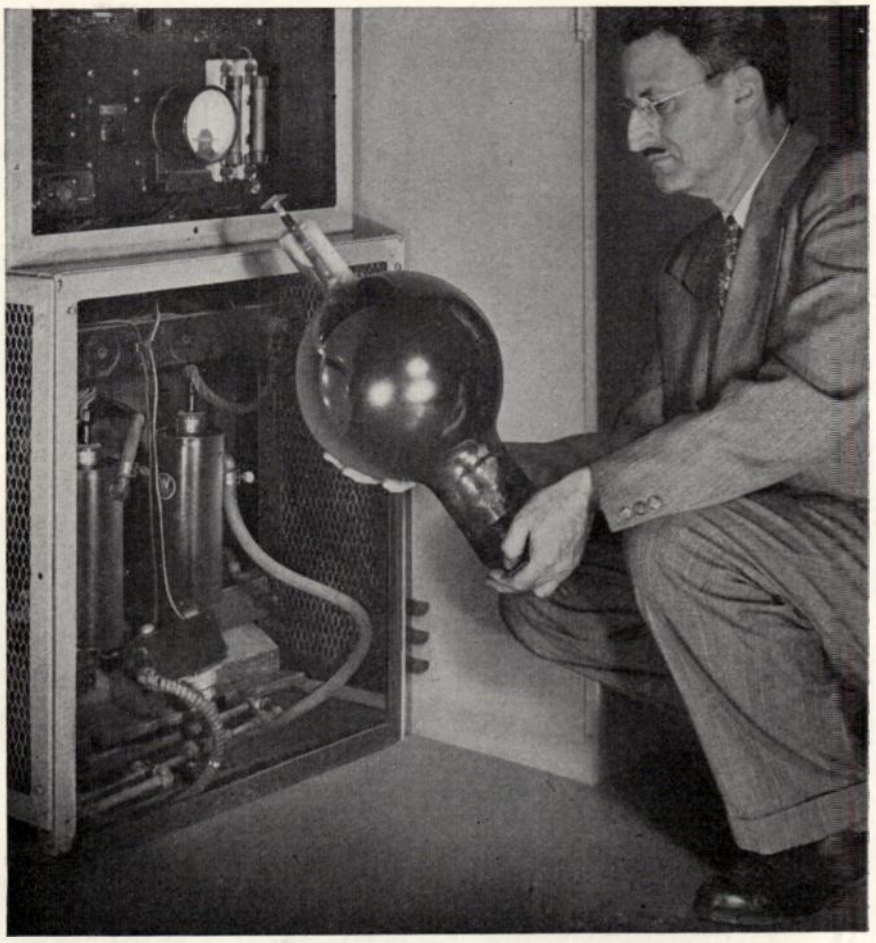
Westinghouse Ignitron Gives 16 Years' Trouble-Free Service



In the Westinghouse plant in Elmira, N. Y., Donald E. Marshall, Section Manager of Gas and Industrial Tube Development, holds a KU-671 Ignitron retired after more than 32,000 hours service.

This rugged old Westinghouse Ignitron was finally retired after 16 years' service only because re-design of resistance welding equipment made a modern Westinghouse Ignitron simpler to use; it is still operable. An exclusive Westinghouse design, the KU-671 was developed in 1937 a few years after Westinghouse engineers first invented the Ignitron —it was the first Ignitron type to be sealed, its predecessors being continuously pumped while in service. (Even some of those earlier Westinghouse tubes are still in active service—pumps and all!)

To an electronic equipment designer such performance means two things:

First, a Westinghouse Ignitron designed into a circuit means reliability and long service without repair, replacement, or down-time due to failure—just as when Westinghouse invented and developed the tube.

Second, the wide variety of modern metal-encased Ignitrons by Westinghouse offers designers of resistance welding and power rectification equipment and inverters possibilities not yet fully explored.

Since the first Ignitron, Westinghouse continued to forge ahead and adapt its Ignitrons to new, more difficult and more exacting applications. Through continuing research and development, Westinghouse has maintained leadership through the years.

For the latest application data, or for design aids or suggestions, write to Department C-1024 at the address below, or call your nearest Westinghouse Electronic Tube sales office.

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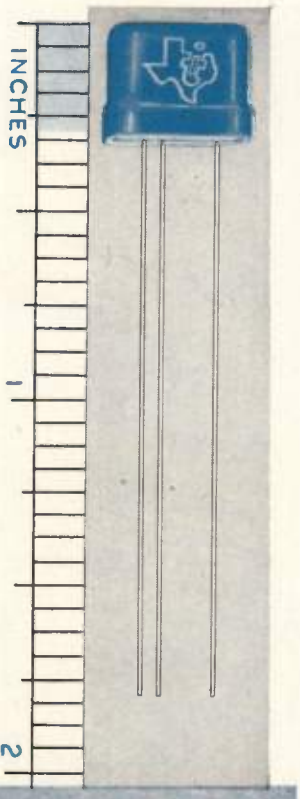
WESTINGHOUSE ELECTRIC CORPORATION, ELECTRONIC TUBE DIVISION, ELMIRA, N. Y.

TI transistors will FIT in Your Future!



FORTY PERCENT SIZE REDUCTION has been made in Texas Instruments junction transistor cases. With case length now less than a third of an inch, TI transistors offer you a major opportunity for equipment miniaturization.

This important size reduction was achieved without reducing the quality or changing the construction of the proven TI junction transistor. Transistor fabrication methods, materials, and moisture-proof glass-to-metal hermetic sealing remain the same.

Texas Instruments junction transistors undergo an exhaustive testing procedure to insure their close adherence to published specifications (see distribution curves below). And not only must all transistors pass more than 20 test procedures—in addition to continual visual checks—but also they are aged over 48 *hours* at rated output. They are then completely re-inspected, as a positive operating double-check.

If *you* want transistors combining small size with high quality, they are now in production and available in five types from Texas Instruments Incorporated. Write for bulletins DL-S 310 (junction) and DL-S 312 (point-contact). Custom-built units also are available.

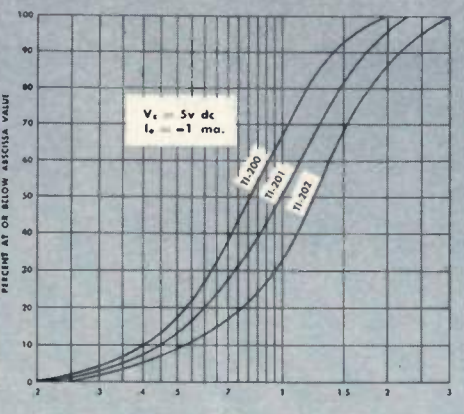
ELECTRICAL DATA:

	n-p-n junction transistors type 200	type 201	type 202	
RATINGS, RECOMMENDED MAXIMUM:				
Collector Voltage	30	30	30	volts
Collector Current	5	5	5	ma.
Collector Dissipation (at 25°C)	50	50	50	mw.
Ambient Temperature	50°C	50°C	50°C	
AVERAGE CHARACTERSTICS (AT 25°C.):				
Collector Voltage	5	5	5	volts
Emitter Current	−1	−1	−1	ma.
Collector Resistance (Minimum)	.4	.4	.4	megohms
Base Resistance	150	170	200	ohms
Emitter Resistance	22	22	35	ohms
Current Amplification Factor* (Minimum)	9	19	49	
Collector Cutoff Current (Maximum)	10	10	10	μa.
Collector Capacitance	15	17	19	$\mu\mu$fd.
Noise Factor** (V_C = 2.5 V., I_C = −.5 ma)	26	23	20	db
Frequency Cutoff** (α_{co})	.90	1.10	1.30	m.c.

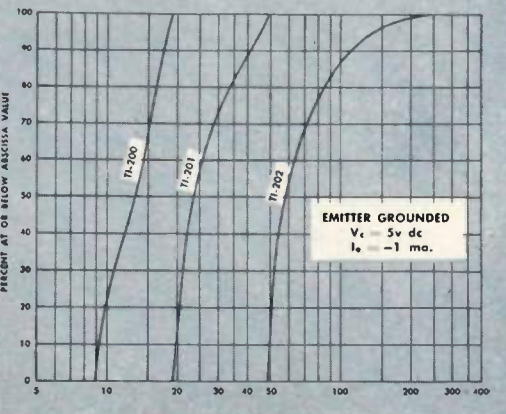
*Emitter Grounded.

**Noise Factor and Frequency Cutoff are average and individual units may vary.

STATISTICAL DISTRIBUTION CURVES
Based on 100 transistors of each type



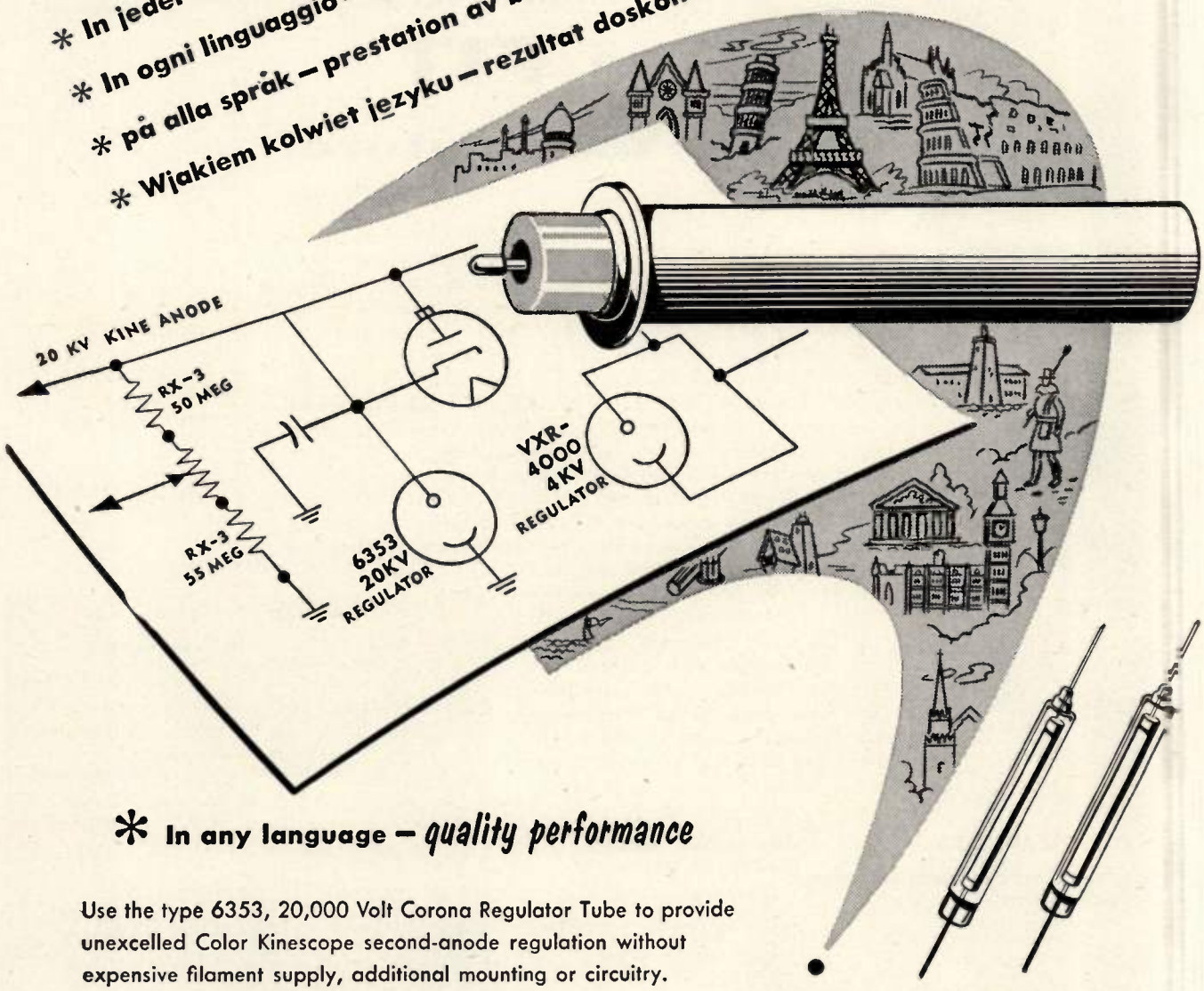
α_{co} — Alpha Cutoff Frequency — megacycles



K — Current Amplification Factor

TEXAS INSTRUMENTS
INCORPORATED
6000 LEMMON AVENUE DALLAS 9, TEXAS

* In jeder sprache — qualitaets leistung
* In ogni linguaggio — qualitád funzionamento
* på alla språk — prestation av bästa kvalitet
* Wjakiem kolwiet języku — rezultat doskonatości



* In any language — *quality performance*

Use the type 6353, 20,000 Volt Corona Regulator Tube to provide unexcelled Color Kinescope second-anode regulation without expensive filament supply, additional mounting or circuitry. To these add unlimited use—life, protection with positive safety and low cost . . . the result is*

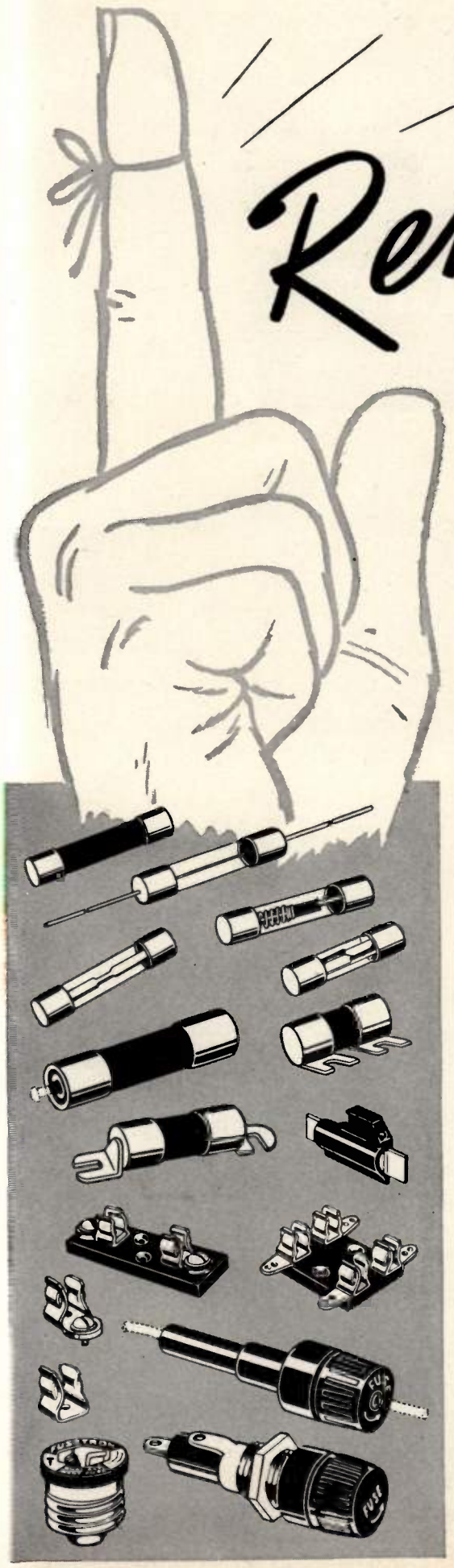
Use the VXR-4000 type tube to regulate dependably the focus bleeder against line voltage variations. This regulator has the same desirable characteristics obtained in the 6353 tube . . . the result is*

These three components, expressly designed for color television will assure quality performance.

Use an RX-3 Series Resistor, 50 or 55 megohm—or select any resistor from 20 to 100 megohm in the kine convergence bleeder to assure stability. These hi-meg resistors are hermetically sealed, rated at 10KV, have a negligible voltage coefficient and .08%/°C temperature coefficient—they are designed for extremely long life . . . The result is*

The Victoreen Instrument Co.

COMPONENTS DIVISION: 3800 PERKINS AVE. • CLEVELAND 14, OHIO



Remember!

For "Trouble-Free" Fuses Use BUSS FUSES

There is a reason manufacturers and service organizations have learned to rely on BUSS fuses to operate properly under all service conditions. Every BUSS fuse normally used by the Electronic Industries is tested in a sensitive electronic device that rejects any fuse that is not correctly calibrated, properly constructed and right in all physical dimensions.

Once properly installed, a BUSS fuse will blow only to protect. If a BUSS fuse does blow, you can be sure there is trouble in the circuit. When the trouble is corrected and a new BUSS fuse installed the job is finished. The user won't be annoyed and your equipment won't be out of service by the fuse failing to operate properly... because a BUSS fuse will carry its rated current and it is properly constructed to prevent poor contact heating causing needless blows.

And by standardizing on BUSS fuses, you can fill your exact fuse needs from one source. The line is complete-dual-element (slow-blowing), renewable and one time types . . . in sizes from 1/500 ampere up.

Should you have a special problem in electrical protection —

Our fuse research laboratory and its staff of engineers are at your service to help you select the right fuse for the job — and if possible a fuse already available in wholesalers stocks.

FUSETRON
TRUSTWORTHY NAMES IN ELECTRICAL PROTECTION
BUSS

For More Information *Mail this Coupon*

BUSSMANN Mfg. Co. (Division of McGraw Electric Co.)
University at Jefferson, St. Louis 7, Mo.

Please send me bulletin SFB containing facts on BUSS small dimension fuses and fuse holders.

*Name*_____

*Title*_____

*Company*_____

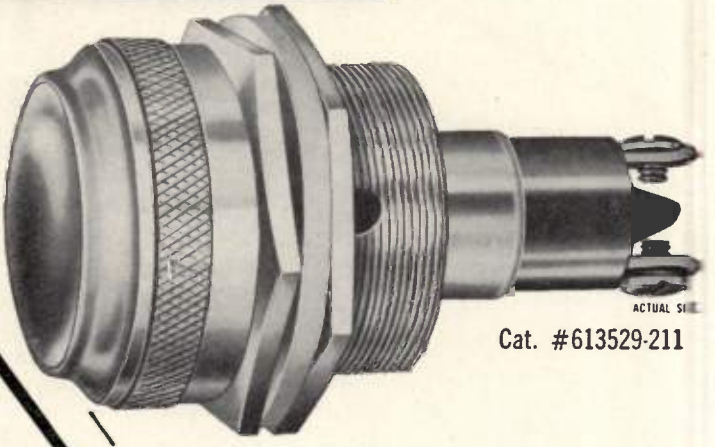
*Address*_____

*City & Zone*_____ *State*_____ IRE-254

for **YOUR** *product*

WHICH PILOT LIGHT DO YOU NEED?

DIALCO



ACTUAL SIZE

Cat. #613529-211

THE BIG ONE

This Pilot Light Assembly was first made to accommodate the *S-11 lamp* and was intended for use in the cabs of great diesel locomotives.

This BIG one

or

this LITTLE one

Dialco HAS THE COMPLETE LINE OF INDICATOR and PANEL LIGHTS

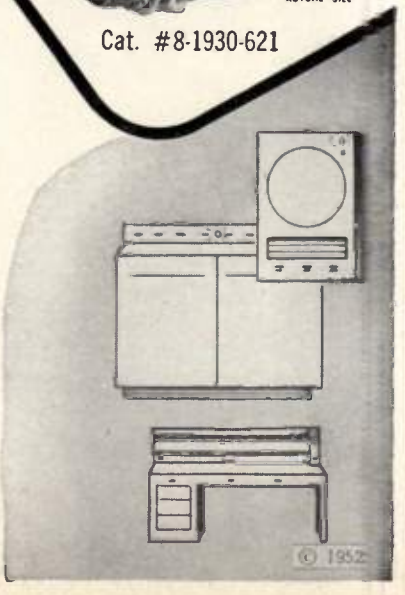
THE LITTLE ONE

The miniaturization program on defense products required the development of this *sub-miniature* light. It is used on communication equipment and aircraft. Midget flanged base bulbs to fit are rated 1.3, 6, 12, and 28 volts.

ACTUAL SIZE

Cat. #8-1930-621

Samples to suit your own special conditions and requirements will be sent promptly and *without cost*. Just outline your needs. Let our engineering department assist in selecting the *right lamp* and the *best pilot light* for YOU.



Write for the Dialco HANDBOOK of PILOT LIGHTS

Foremost Manufacturer of Pilot Lights

DIALIGHT CORPORATION

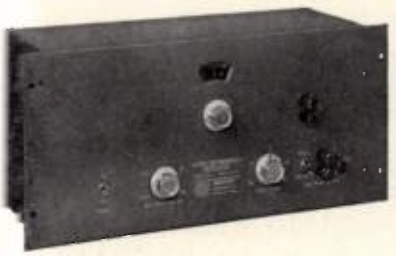
60 STEWART AVE., BROOKLYN 37, N. Y. HYACINTH 7-7600

News—News Products

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

Low Frequency Oscillator

A new low-frequency oscillator, Model 420-C, designed by **Krohn-Hite Instrument Co.**, 580 Massachusetts Ave., Cambridge 39, Mass., is intended for applications requiring a source of sine-wave and square-wave voltages in the frequency range from 0.33 to 52,000 cps. Special circuits have been installed to eliminate transients after tuning or band-switching. Maximum output is 30 volts peak to peak across a 1,000 ohm load.



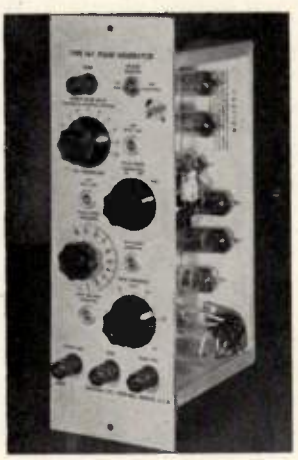
A standard rack panel construction is used, with the unit enclosed in a steel cabinet.

The frequency range, extending from sub-audio to above the audio range, makes the 420-C useful for servo-mechanism, geophysical, and seismological work, for vibration checks and medical research, and for work with audio frequency circuits.

The dimensions of the unit are 19×8¾×7½ inches, and the price is $325. A descriptive pamphlet is available on request.

Fast-Rise Pulse Generator

A new fast-rise pulse generator, Type 163, has been added to the Type 160 series of special wave-form generators manufactured by **Tektronix, Inc.**, P. O. Box 831, Portland 7, Oregon.



The Type 163 is designed to supply rectangular pulses of less than 0.2-μsec rise-time when triggered by either a positive pulse of calibrated continuously-variable amplitude from 0 to 25 volts peak to peak, and a positive gate of 25 volts amplitude are supplied. The pulse and gate are identical in other characteristics. Duration is

(Continued on page 31A)

FAIRCHILD

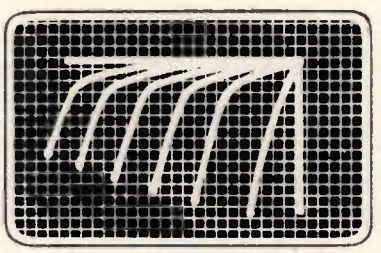
Transistor Analyzer

Complete—all calibrating circuits built-in.
Needs only standard DC oscilloscope.

Developed by Fairchild in its program for transistorizing Missile Guidance Systems and other intricate electronic circuits.

Rapidly Plots Static and Dynamic Characteristics of ALL Transistors ...point contact and junction. Designed on basic principles, to meet future transistor needs.

Complete Families of Curves obtainable in 10 incremental steps for each of 5 ranges. Sweeping technique shows up anomalies.



Presents on the Scope:

Alpha vs. Emitter Current

•

Collector, Emitter and Transfer Characteristics

•

Collector Characteristics in Grounded Emitter Connection

ENGINE AND AIRPLANE CORPORATION

FAIRCHILD

Guided Missiles Division

Wyandanch, L. I., N. Y.

Other Divisions: Aircraft Division, Hagerstown, Md. • Engine Division, Farmingdale, N. Y.

FAIRCHILD ENGINE & AIRPLANE CORPORATION
GUIDED MISSILES DIVISION, WYANDANCH, L. I., N. Y.

Please send me your Detailed Transistor Analyzer Technical Bulletin.

Name and Company_____

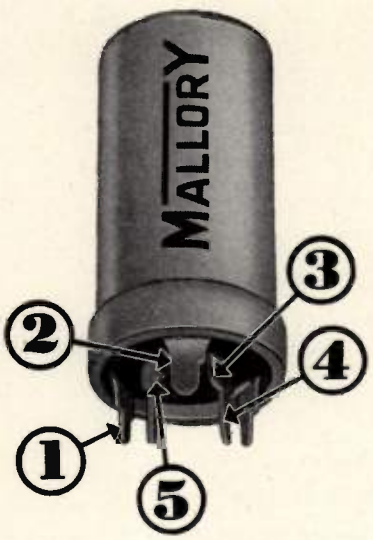
Address_____

City_____ Zone_____ State_____

P

Mallory Capacitors for printed circuit applications

Mallory FP Electrolytic Capacitors are now available in a construction that is specifically designed for printed electronic circuits. Their mounting prongs and terminals have been adapted for quick, foolproof production line assembly.



1 *Prongs and terminals are smaller . . . take less space . . . hold securely with a small amount of solder. No need for large copper areas in the printed chassis.*

2 *Self-positioning. One mounting prong is wider. The capacitor can fit only in its correct position.*

3 *Shoulders on prongs hold capacitor clear of chassis, permitting use of printed circuitry on both sides.*

4 *Prongs can be automatically spread on insertion by simple jig, insuring strong mechanical mounting prior to soldering.*

5 *Positive soldering. Aluminum risers do not extend through chassis. No danger of contaminating solder.*

These refinements of design are one more step in the capacitor developments which Mallory has pioneered, and which have made Mallory FP Capacitors the leader in their field. Write or call us for detailed literature, or for engineering service on your capacitor problems by one of our field specialists.

Expect more . . . Get more from P.R. MALLORY & CO. Inc. MALLORY

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

Serving Industry with These Products:

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

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

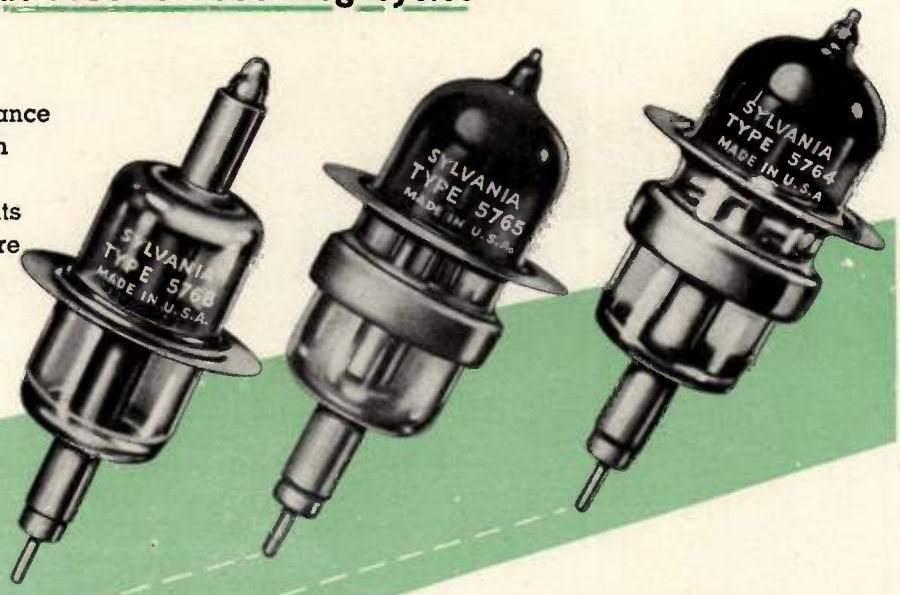
SYLVANIA

ROCKET TUBES

(PLANAR TRIODES)

The Answer to Efficient Power at 1000 to 4000 Megacycles

Low Lead Inductance
Low Interelectrode Capacitance
Rugged Planar Construction
Small Size
Moderate Input requirements
Good frequency-temperature Characteristics



TYPICAL APPLICATIONS

TYPE	SERVICE	FREQUENCY MC	OUTPUT
2C36	Pulsed oscillator	1000	125 Watts
2C37	CW oscillator	1000-3300	450 MW.
5764	Pulsed oscillator	2900	200 Watts
5765	CW Tunable oscillator	900-2900	250 MW. Av over the band
5768	CW Tunable amplifier	1000-3000	10 db. av gain
RT434	CW oscillator for butterfly type circuits	1000-3000	400 MW.

Still more reasons why it pays to SPECIFY SYLVANIA

For complete information concerning Sylvania Rocket Tubes or other electronic tubes write to Sylvania, Dept. 4E-3102.

SYLVANIA

Sylvania Electric Products Inc. 1740 Broadway, New York 19, N. Y.

LIGHTING · RADIO · ELECTRONICS · TELEVISION

In Canada: Sylvania Electric (Canada) Ltd., University Tower Bldg., St. Catherine St., Montreal, P. Q.

OHMITE®

FOR *Dependability*

RHEOSTATS

All-ceramic and metal, close control rheostats for unsurpassed dependability and smoothness of operation. Ten stock sizes, 25 to 1,000 watts.

TAP SWITCHES

Five compact models, 10 to 100 amperes, AC, up to 12 taps. All-ceramic and metal construction. Silver-to-silver contacts, with self-cleaning rotor.

RESISTORS

A wide range of dependable, fixed, adjustable, tapped, and non-inductive, power wire-wound resistors. Also a wide range of precision resistors.

R. F. CHOKES

Single layer R. F. plate chokes and power line chokes, on steatite or plastic cores. Protected by a special moisture-resistant coating.

Be Right with **OHMITE**

OHMITE MANUFACTURING COMPANY

3617 Howard St., Skokie, Illinois (Suburb of Chicago)

WRITE on company letterhead for Catalog and Engineering Manual No. 40

News—New Products

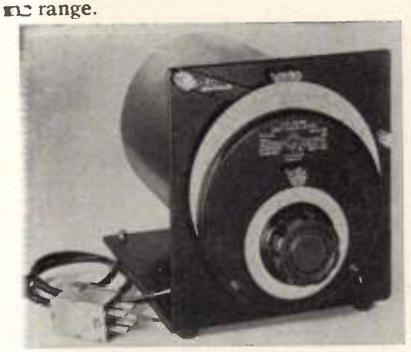
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 27A)

calibrated and is continuously variable from 1 μsec to 10,000 μsec. When triggered by a saw-tooth voltage, the output may be delayed a calibrated interval from 0 to 100 per cent of the duration of the sawtooth. Decay time is 0.2 μsec, and overshoot can be adjusted to zero.

Unit Oscillator

The General Radio Co., 275 Massachusetts Ave., Cambridge 39, Mass., has a new Type 1211-A Unit Oscillator, the latest span of the oscillator is 0.5 to 50 mc, which is covered in two 10-to-1 logarithmic ranges. Frequency is read directly from a 6-inch dial, with a slow-motion-drive dial indicating frequency increments of 0.2 per cent per division. Output power is well over one watt over the 0.5- to 5-mc range and is at least 0.2 watt over the 5- to 50-mc range.

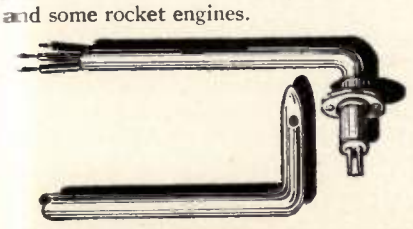


The construction of the Type 1200-A Unit Oscillator provides effective shielding so that the instrument can be used as a power source in bridge measurements. The Type 874 Coaxial Output Connector permits extension of the shield system to the bridge. The Type 1203-A Unit Power Supply is available for ac operation. Batteries can be used for field applications. Direct amplitude modulation over the audio-frequency range is possible with an external audio oscillator.

The Type 1211-A Unit Oscillator is priced at $295.00, and the Type 1203-A Power Supply is available at $40.00, both net f.o.b. Cambridge, Mass.

Fluid Flow Instrumentation

The Aero Research Instrument Co., 1901 N. Hermitage Ave., Chicago 22, Ill., announces the manufacture of new and improved probes for sensing gas temperature, pressure, direction, velocity, and turbulence in and about turbo jets, ram jets and some rocket engines.



The company has an engineering staff that is capable of solving probe problems

(Continued on page 32A)

Now! Measure reflection coefficient or VSWR quickly; match loads to line

Model 136A Reflection Coefficient Meter

New Model 136A Reflection Coefficient Meter is a compact, moderately-priced instrument designed for quick, simple measurement of transmission line reflection coefficient or VSWR. It also provides a highly convenient method of matching loads to line to minimize reflected power, and may be used as a wide range laboratory receiver. The instrument is very simple to operate and particularly designed for rapid, accurate use by non-technical personnel.

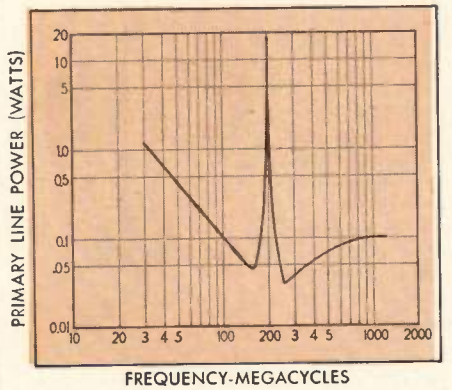


Figure 1. Sensitivity, Model 136A. Primary line CW power required to read reflection coefficient .02 as function of frequency. (Values for Sierra Models 138 and 138A Couplers).

Data subject to change without notice.

2960

Model 136A Meters use standard Sierra Wideband Directional Couplers to sample incident and reflected voltage in the transmission line under test. A superheterodyne vacuum tube voltmeter may be switched to either the incident or reflected coupler output. In the incident voltage position, a precision attenuator (calibrated directly in reflection coefficient and VSWR) is inserted in the IF amplifier circuit. The Meter's frequency range is 32 to 1125 mc. Sufficient sensitivity is provided for use with available signal sources.

See these and other new Sierra instruments at the I.R.E. Show, Booth 711.

sierra

Sierra Electronic Corporation

San Carlos 2, California, U. S. A.

Sales representatives in major cities

Manufacturers of Carrier Frequency Voltmeters, Wave Analyzers, Line Fault Analyzers, Directional Couplers, Wideband RF Transformers, Custom Radio Transmitters, VHF-UHF Detectors, Variable Impedance Wattmeters, Reflection Coefficient Meters.

IN RELAY KLYSTRONS,

Quality counts most...

For performance that measures up to the exacting demands of high frequency relay systems, there can be no compromise with quality.

System designers and equipment buyers know that high performance, high frequency systems depend upon klystrons having sufficient power to override noise, excellent frequency stability and long life. Varian klystrons are **designed** and **built** to meet these exacting demands.

VARIAN MEANS **PROVED** PERFORMANCE...

In the 6000 to 8000 megacycle band, Varian X-26 klystrons outperform all others. Here are five reasons why leading system designers insist on these klystrons for top performance in relay applications:

- **Greater Power** — X-26 high power klystrons are conservatively rated. They will deliver **more** than rated power **without** failure.
- **Greater Frequency Stability** — X-26 klystrons have negligible short term drift — long term drift is less than 5 megacycles.
- **Greater Uniformity** — Varian mass production techniques assure uniformity — every klystron is as reliable as a nut and bolt.
- **Longer Life** — X-26 klystrons can be operated at full power for thousands of hours, at low power for **years**.
- **Less Distortion, Less Noise** — FM distortion and inherent noise are negligible — 60 db below a 1-megacycle deviation.

IN EVERY KLYSTRON APPLICATION, VARIAN GIVES YOU:

- *Advanced Design*
- *Proved Performance*
- *Operating Economy*
- *Structural Integrity*

IN KLYSTRONS, THE MARK OF LEADERSHIP IS

VARIAN associates

PALO ALTO 2, CALIFORNIA

Representatives in all principal cities

News—New Products

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

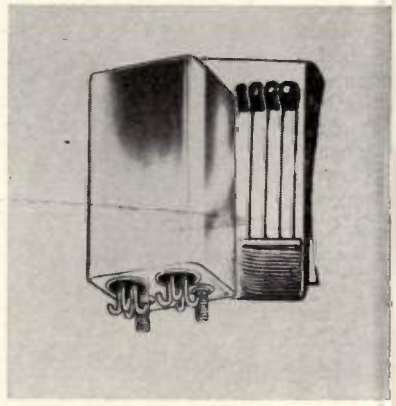
(Continued from page 31A)

and can supply aerodynamic performance characteristics with their probes. Aero Research can also manufacture probes in accord with your specific design.

Illustrated bulletin will be sent upon request to the company.

Miniature Relays

Signal Engineering & Mfg., 154 W. 14 St., New York 11, N. Y., is offering a new line of miniature sensitive relays.



These small, lightweight sensitive relays (Series 100–105–106) measure 1×1×1¾ inches and are sensitive to low input. They are hermetically sealed with headers for solder connections and have high shock immunity. This series is available in either single-pole, double-throw or double-pole, and double-throw. Standard operating power is 40 milliwatts for dpdt with other contact arrangement and conditions permitting lower operating power; 10g vibration immunity in accordance with MIL-R-5757B; silver contacts rated 2 amperes, 28 volts, dc resistive; life is over 100,000 operations at rated resistive load; double-coil magnetic circuit of high efficiency type with windings up to 8000 ohms and 1 watt maximum dissipation. Request Bulle in 100.

JAN Connectors

Coaxial Connector Co., 35 N. Second Ave., Mt. Vernon, N. Y., is now in production on PL-294 and PL-279 connectors, made in accordance with JAN specifications. These connectors are used on the cable assemblies CD-1086 and CX-75TRC, which form part of AN/GRC-9 and SCR-694 equipment.

These heavy duty 9-pin connectors have a body and coupling nut of half-hard brass, QQ-B-611, and melamine insulation. The contacts are silver-plated brass and spring phosphor bronze. The finish is olive drab chromate on a cadmium base. The cable clamp assemblies are precision aluminum castings.

The company has a complete line of connectors, and is ready to supply special designs, engineering and production. Inquiries are invited. A condensed catalog will be sent on request.

(Continued on page 41A)

Check these:

Stabilizing and Regulating Ability

Correction Speed

Waveform Distortion

Input Range

Output Range

and you will specify

STABILINE TYPE IE

Instantaneous Electronic

AUTOMATIC VOLTAGE REGULATORS

Here's how the Stabiline type IE measures up:

Stabilizing and regulating ability — For all conditions maximum variation less than ±.25 of 1%. For input voltage changes, variation less than ±0.1 of 1%. Load current change or power factor change from lagging .5 to leading .9 will vary output voltage less than ±.15 of 1%.

Correction speed — Comparatively instantaneous — 3 to 10 cycles.

Waveform distortion — Never exceeds 3%. Is generally under 2%.

Input Range — For nominal 115 volts output, input range is 95 to 135 volts. For nominal 230 volts output, input range is 195 to 255 volts.

Output Range — Output voltage on 115 volt units can be adjusted from 110 to 120 volts; on 230 volt units from 220 to 240 volts.

Furthermore, the Stabiline type IE has a circuit simplicity and mechanical ruggedness that minimizes maintenance.

Check all these characteristics against all other automatic voltage regulators and you will find Stabiline type IE is *superior* in design, construction and performance.

Stabiline automatic voltage regulators type IE are available in ratings from .25 to 5.0 KVA. Special types will be application engineered to meet specific requirements.

Send Coupon Today for Bulletin S351

THE SUPERIOR ELECTRIC COMPANY

BRISTOL, CONNECTICUT

Manufacturers of

POWERSTAT VARIABLE TRANSFORMERS
STABILINE AUTOMATIC VOLTAGE REGULATORS
VOLTBOX A-C POWER SUPPLIES
POWERSTAT LIGHT DIMMING EQUIPMENT
VARICELL D-C POWER SUPPLIES
SUPERIOR 5-WAY BINDING POSTS

Send Coupon Today

THE SUPERIOR ELECTRIC CO.
1102 Clarke Avenue, Bristol, Conn.
Please send my copy of Bulletin S351

Name .....
Position .....
Co. Name .....
Co. Address .....
City ..... *Zone* ..... *State* .....

BALLANTINE Sensitive, Wide Band Electronic Voltmeter

MODEL 314
Price $285

To measure	**1 millivolt to 1000 volts**
from	**15 cycles to 6 megacycles**
with accuracy of	**3% to 3 mc; 5% above**
with input impedance	**7.5 mmfds shunted by 11 megs**

When used without probe, sensitivity is increased to 100 MICRO-VOLTS but impedance is reduced to 25 mmfds and 1 megohm

Featuring customary Ballantine

SENSITIVITY — ACCURACY — STABILITY

- Same accuracy at *ALL* points on a logarithmic voltage scale and uniform DB scale.
- Only *ONE* voltage scale to read with decade range switching.
- No "turnover" discrepancy on unsymmetrical waves.
- Easy-to-use probe with self-holding connector tip and unique supporting clamp.
- Low impedance ground return provided by supporting clamp.
- Stabilized by generous use of negative feedback.
- Can be used as 60 DB high fidelity video pre-amplifier.

Write for catalog for more information about this and other BALLANTINE voltmeters, amplifiers, and accessories.

BALLANTINE LABORATORIES, INC.
102 Fanny Road, Boonton, N.J.

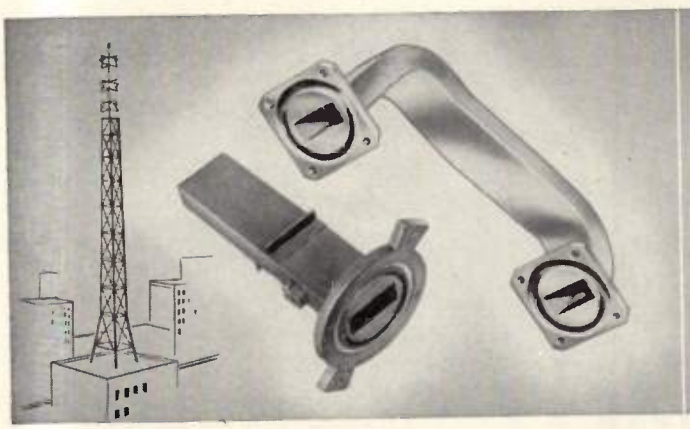
Industrial Engineering Notes

(Continued from page 19A)

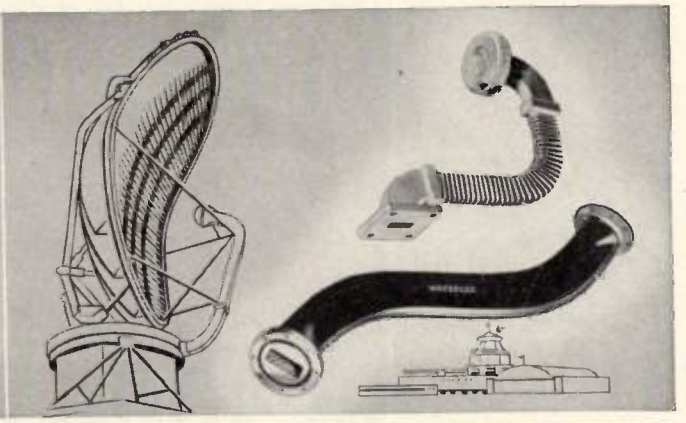
Federal Communications Commission recently, by public notice, has stated that no necessity exists for providing for the licensing of passive antenna systems which merely divert or reflect signals of radio or television broadcasting stations. Persons venturing into the use of passive antenna systems were cautioned, however, against the use of any electrical equipment or any other apparatus in connection with such systems which would result in the amplification of the signal strength of the station whose signal is being re-radiated or reflected by the passive antenna system. . . . **The FCC recently issued a Notice of Proposed Rule Making looking toward amending Part 2 of the rules to reallocate the radio spectrum above 890 mc to meet the existing service needs and foreseeable developments in the use of microwave by various fixed and mobile services.** The proposed amendments, it was reported, are the result of Commission study of current frequency problems and look toward additional exclusive spectrum space for TV pickup purposes; exclusive space for common carrier mobile functions including additional TV pickup service, and other mobile non-broadcast video pickup requirements; additional exclusive broadcast studio-transmitter link frequencies for AM, FM and FM (sound only); additional common carrier fixed service spectrum space, particularly with respect to the needs of areas where the use of the common carrier bands above 3700 mc is not practicable; and additional exclusive, as compared to some present shared, allocations for the non-government fixed services now operating in the vicinity of 900 mc. The proposal would continue to allocate the band 890–940 mc to the ISM (Industrial, Scientific, Medical) service primarily, would allocate the band 890–936 mc to common carrier fixed service, with

(Continued on page 36A)

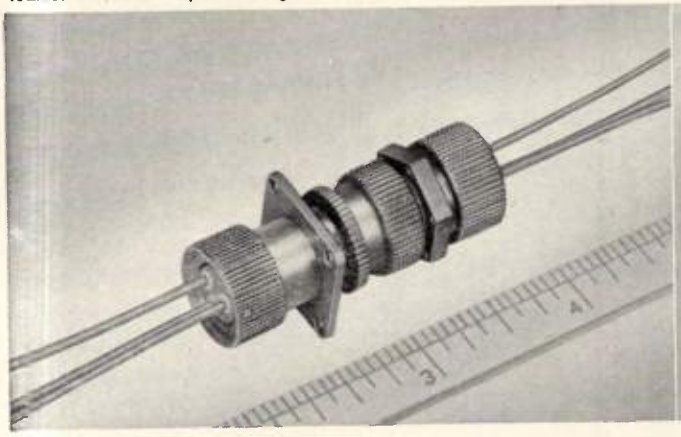
All these from one experienced source



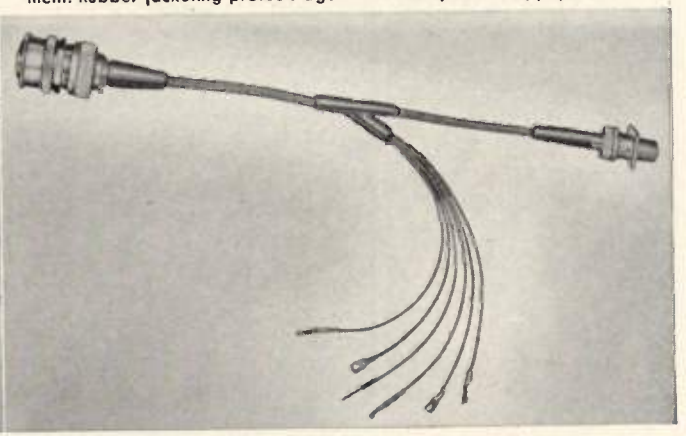
TITEFLEX DESIGNS AND MANUFACTURES — to customer specifications — rigid waveguides and combinations of rigid and flexible waveguides. Where there is, or should be, no movement, or where complicated accessories must be connected, Titeflex *rigid* waveguides are specially recommended.



WAVEFLEX® FLEXIBLE WAVEGUIDES are fabricated to retain critical dimensions — regardless of twisting or bending. Waveflex waveguides make assembly easy, improve design, compensate for expansion or movement. Rubber jacketing protects against weather, corrosion, physical abuse.



TITEFLEX CONNECTOR—lightweight, corrosion and moisture resistant with temperature ranges of —65°F. to +400°F. This connector's insulation properties will permit 3500 volts at sea level, 1200 volts at 50,000 feet altitude. Connector is available with 2 or 3 pins. 7 amperes. Weight ¼ of ounce. Size 2" in length.



TITEFLEX CUSTOM WIRING SYSTEMS are corrosion resistant, moisture proof, pressure-tight and efficient at temperatures of — 65°F. to +400°F. Can be furnished with Titeflex or standard AN connectors for a wide range of service requirements. Can be sheathed with metal braids, fiber glass or nylon—and jacketed with silicone or other compounds.

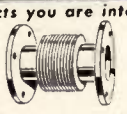
MORE THAN 37 YEARS of developmental experience make Titeflex a logical source of the components pictured on this page. We are currently in a position to supply connectors and wiring systems to makers of aviation and electronic equipment. If you have a problem requiring our unusual combination of products and engineering, let us quote on your requirements. The coupon will bring you information on our products.

Let Our Family of Products Help Yours

Titeflex

✓ Check products you are interested in.

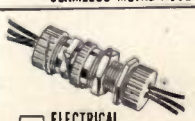
☐ SEAMED AND SEAMLESS METAL HOSE



☐ PRECISION BELLOWS

☐ IGNITION HARNESS

☐ IGNITION SHIELDING



☐ ELECTRICAL CONNECTORS

☐ RIGID AND FLEXIBLE WAVE GUIDES

☐ WIRING SYSTEMS

☐ FUSES

TITEFLEX, INC.
511 Frelinghuysen Ave.
Newark 5, N.J.

MAIL COUPON TODAY

Please send me without cost information about the products checked at the left.

NAME_____

TITLE_____

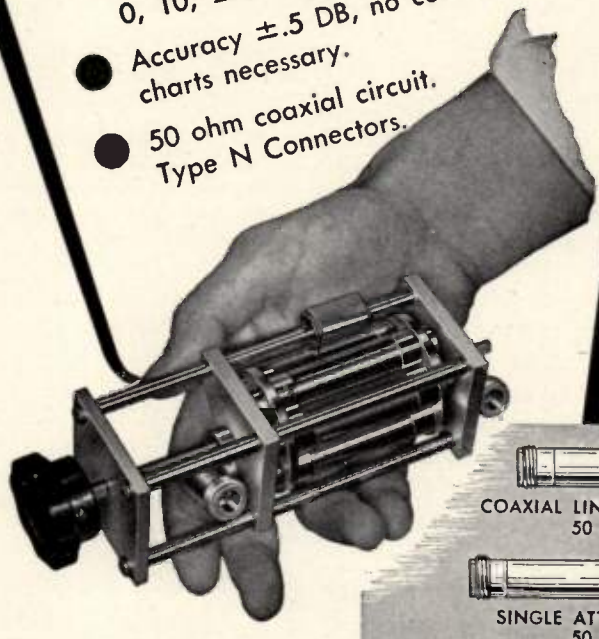
FIRM_____

ADDRESS_____

CITY_____ ZONE____ STATE_____

Precision ATTENUATION to 3000 mc!

- VSWR less than 1.2 at all frequencies to **3000 mc.**
- **TURRET ATTENUATOR** featuring **"Pull - Turn - Push"** action with 0, 10, 20, 30, 40, 50 DB steps.
- Accuracy ±.5 DB, no correction charts necessary.
- 50 ohm coaxial circuit. Type N Connectors.



COAXIAL LINE TERMINATION
50 ohms

SINGLE ATTENUATOR PAD
50 ohms

VSWR ± 1.2 to 3000 mc.
One watt c.w. power dissipation

Inquiries are invited concerning single pads and turrets having other characteristics

STODDART AIRCRAFT RADIO CO., INC.
6644-C SANTA MONICA BLVD., HOLLYWOOD 38, CALIFORNIA
HOllywood 4-9294

Industrial Engineering Notes

(Continued from page 34A)

the provision that ISM interference be accepted; the 936–950 mc band would be allocated to AM, FM, and TV (sound only) studio-transmitter stations, subject to accepting ISM interference between 936–940 mc; and the band 950–960 mc would be allocated to operational fixed stations. The presently shared "mobile" bands 3500–3700, 6425–6575 and 11700–12200 mc are also proposed to be divided equally for exclusive assignment to the Domestic Public services on the one hand, and the Aviation, Industrial, Land Transportation, Marine and Public Safety Services. . . . **The FCC recently finalized an earlier Notice of Proposed Rule Making by amending Part 2 of its frequency allocation rules to add a footnote to all frequency bands allocated to the broadcasting services in which it does not now appear, thus permitting appropriate experimentation in all of the high frequency broadcasting bands on a non-interference basis.**

RETMA Activities

The RETMA proposal for a joint FCC-industry survey of facilities in use in the mobile radio field was approved by members of the Federal Communications Commission at a recent meeting. Originally suggested by members of the Land Mobile Communications Section, of the Technical Products Division, the proposal has been under study by the Section and FCC staff members for a number of months. A subcommittee to be appointed by Section Chairman Myron Whitney is expected to meet with FCC representatives in the very near future to iron out mechanical details of the survey. The proposed equipment and usage census will go a long way towards setting up a true picture of the mobile radio field. Money for the joint survey was appropriated by the RETMA Board of Directors at Chicago in June, 1953.

(Continued on page 66A)

IRE DIRECTORY QUESTIONNAIRE

Send to: Industry Research Div., Institute of Radio Engineers
1475 Broadway, 22nd Floor, New York 36, N. Y.

Date .....

Firm Name: ..... Phone No. .....

Address: ..... City ..... Zone..... State.....

Name of: Chief Engineer .....

Advertising Manager .....

1-54 Purchasing Agent .....

Method to be used by firms to indicate products manufactured or services rendered. Please read the entire questionnaire. Products are grouped beneath basic headings to facilitate location.

Do not check products that you do not manufacture for sale to the industry.

Components manufactured for incorporation in your own end products (and not intended for sale) should not be listed, as this will only result in annoyance to you, and the firms who make inquiries concerning product availability.

If, after a thorough examination of this form, you do not find a listing for your products, a short description and literature or catalogs that you have should be attached and returned to us.

After checking off the products, please fill in full data on your firm name and address, and mail the entire form to: Frank MacAloon, Industry Research Div., Institute of Radio Engineers, 1475 Broadway, New York 36, N. Y.

Industry Research Division Information Service

1. Audio Frequency Equipment.

01 Amplifiers.

05 ☐ Decade amplifiers
10 ☐ Dynamic noise suppressors
15 ☐ Hearing aid
20 ☐ High fidelity
25 ☐ Inter-coms
30 ☐ Phonograph pre-amplifiers
35 ☐ Power amplifiers—recording
40 ☐ Public address

02 Attenuators.

05 ☐ Decade
10 ☐ Impedance matching
15 ☐ Volume controlling

03 Equalizers.

05 ☐ Dialogue equalizers
10 ☐ Line equalizers
15 ☐ Magnetic recording
20 ☐ Recording diameter equalizers
25 ☐ Sound effect equalizers

04 Filters.

05 ☐ Band pass & band rejection
10 ☐ Loudspeaker dividing networks
15 ☐ Variable band pass

05 Industrial Sound Systems.

05 ☐ Complete systems
10 ☐ Installation & rental contractors

06 Jacks, Jack Fields & Plugs.

05 ☐ Jacks
10 ☐ Jack fields
15 ☐ Patch cords
20 ☐ Plugs

07 Loudspeakers & Headphones.

05 ☐ Commercial grade loudspeakers
10 ☐ Headphones
15 ☐ High-frequency loudspeakers
20 ☐ Low-frequency loudspeakers
25 ☐ Wide-range, 2 & 3 way systems

08 Microphones & Stands.

05 ☐ Carbon
10 ☐ Ceramic
15 ☐ Condenser
20 ☐ Crystal
25 ☐ Dynamic
30 ☐ Velocity
35 ☐ Microphone stands

09 Phonograph, Pick-ups, Record Changers, etc.

05 ☐ Phonograph motors
10 ☐ Pick-ups—crystal
15 ☐ Pick-ups—dynamic
20 ☐ Pick-ups—magnetic
25 ☐ Pick-ups—reluctance
30 ☐ Playback arms
35 ☐ Record changers
40 ☐ Record players
45 ☐ Styli-diamond
50 ☐ Styli-metallic
55 ☐ Styli-sapphire
60 ☐ Tape phonographs & playback equipment

10 Recording Accessories & Supplies.

05 ☐ Blank disks
10 ☐ Cutting heads
15 ☐ Cutting styli
20 ☐ Magnetic recording, playback & erase heads
25 ☐ Magnetic recording tape-acetate base
30 ☐ Magnetic recording tape-paper base
35 ☐ Magnetic recording wire
40 ☐ Stylus resharpening
45 ☐ Suction pumps & equipment
50 ☐ Tape splicing equipment

11 Recording Equipment.

05 ☐ Disk recorders
10 ☐ Embossing types of recorders
15 ☐ Magnetic tape recorders
20 ☐ Magnetic wire recorders

2. Broadcasting: Both Aural & Visual; Receivers, Transmitters & Associated Equipment.

12 Antennas & Accessories.

05 ☐ AM broadcast transmitting
10 ☐ FM broadcast transmitting
15 ☐ TV broadcast transmitting
20 ☐ Supporting towers
25 ☐ Tower lighting equipment
30 ☐ Tower erection & maintenance service

13 Monitor Equipment.

05 ☐ Frequency
10 ☐ Modulation
15 ☐ Remote control monitors
20 ☐ Video monitors

14 Receivers.

05 ☐ AM broadcast
10 ☐ AM shortwave
15 ☐ FM broadcast
20 ☐ TV broadcast — B & W
25 ☐ TV broadcast — Color

15 Services for the Broadcaster.

05 ☐ Frequency measurement
10 ☐ Packaged program services
15 ☐ Transcription libraries

16 Studio Equipment.

05 ☐ Consoles & speech input equip.
10 ☐ TV cameras
15 ☐ TV camera control equipment
20 ☐ TV film projectors
25 ☐ TV studio lighting equipment
30 ☐ Turntables

17 Transmitters.

05 ☐ AM broadcast
10 ☐ FM broadcast
15 ☐ TV broadcast
20 ☐ Relay broadcast & STL equip.

3. Communications: All non-broadcast forms of communication; Amateur, Emergency, Facsimile, Mobile, etc.

18 Aircraft & Airport Equip.

05 ☐ Airborne receivers
10 ☐ Airborne transmitters
15 ☐ Airport receivers
20 ☐ Airport transmitters
25 ☐ GCA equipment, etc.

19 Amateur Equipment.

05 ☐ Antennas & antenna accessories, rotators, etc.

10 ☐ Receiving equipment
15 ☐ Transmitters

20 Carrier Current Equipment.

05 ☐ Complete terminal equip.
10 ☐ Installation & maintenance
15 ☐ Repeat & relay equip.

21 Emergency Communications Equipment.

05 ☐ Despatching equipment
10 ☐ Walkie-talkies

22 Facsimile Equipment.

05 ☐ Receiving equipment
10 ☐ Sensitized paper
15 ☐ Transmitting equipment

23 Marine Equipment.

05 ☐ Contract installers
10 ☐ Pleasure craft radiotelephones
15 ☐ Shipborne communications equip.

24 Mobile Equipment.

05 ☐ Antennas
10 ☐ Central station receivers
15 ☐ Central station transmitters
20 ☐ Citizen radio
25 ☐ Mobile receivers
30 ☐ Mobile transmitters
35 ☐ Selective calling equipment

25 Point-to-Point Equipment.

05 ☐ Antennas & towers
10 ☐ Receivers
15 ☐ Transmitters

4. Component Parts

26 Amplifiers.

05 ☐ DC amplifiers
10 ☐ IF amplifiers
15 ☐ Magnetic amplifiers
20 ☐ Microwave amplifiers
25 ☐ RF amplifiers
30 ☐ Servo-types
35 ☐ TV amplifiers
40 ☐ TV boosters
45 ☐ Wideband amplifiers

27 Antennas & Accessories.

05 ☐ All-wave receiving
10 ☐ Dummy
15 ☐ Feeder line
20 ☐ Indoor types
25 ☐ Insulators
30 ☐ Lightning arrestors
35 ☐ Mounting hardware
40 ☐ Small towers
45 ☐ TV multiple systems

28 Batteries.

05 ☐ Dry batteries
10 ☐ Mercury batteries
15 ☐ Storage batteries — lead-acid
20 ☐ Storage batteries — nickel-alkali
25 ☐ Storage batteries — silver-zinc

29 Blowers & Cooling Fans.

05 ☐ Air wheels & housings
10 ☐ Complete blower assemblies
15 ☐ Fan blades

30 Cabinets, Consoles, & Enclosures.

05 ☐ Metal cabinets
10 ☐ Metal relay rack panels
15 ☐ Moulded plastic cabinets
20 ☐ Wood cabinets

31 Capacitors: Fixed.

05 ☐ Ceramic
10 ☐ Electrolytic
15 ☐ High voltage
20 ☐ Metallized paper
25 ☐ Mica
30 ☐ Oil filled
35 ☐ Paper
40 ☐ Plastic insulated
45 ☐ Power factor
50 ☐ Precision air
55 ☐ Printed circuit
60 ☐ Pulse forming networks
65 ☐ Silvered mica
70 ☐ Tantalum
75 ☐ Vacuum

32 Capacitors: Variable.

05 ☐ Neutralizing
10 ☐ Precision variable
15 ☐ Temperature compensating
20 ☐ Trimmers, air
25 ☐ Trimmers, ceramic
30 ☐ Trimmers, glass
35 ☐ Trimmers, vacuum
40 ☐ Tuning

33 Chassis & Racks.

05 ☐ Aluminum chassis
10 ☐ Chassis slides
15 ☐ Relays racks & cabinets
20 ☐ Steel chassis

34 Coils.

05 ☐ Audio frequency chokes
10 ☐ Deflection yokes
15 ☐ Filter
20 ☐ Flyback
25 ☐ Focusing coils
30 ☐ RF & IF chokes
35 ☐ Toroids
40 ☐ Tuning

35 Connectors.

05 ☐ AN & MIL standard types
10 ☐ Co-axial
15 ☐ Microphone
20 ☐ Multiple circuit connectors
25 ☐ Power
30 ☐ Slip ring and commutating connectors
35 ☐ Sub-miniature

36 Converters.

05 ☐ AC to DC
10 ☐ DC to AC
15 ☐ Inverters
20 ☐ Rotary
25 ☐ Vibrators

37 Cores.

05 ☐ Ceramic
10 ☐ Ferrites
15 ☐ Pre-formed, spirally wound
20 ☐ Powdered metal
25 ☐ Toroids

38 Crystals & Accessories.

05 ☐ Crystal holders
10 ☐ Crystal ovens
15 ☐ Germanium
20 ☐ Oscillating quartz crystals
25 ☐ Piezo electric

39 Delay Lines.

05 ☐ Distributed constant
10 ☐ Distributed perameter
15 ☐ Lumped constant
20 ☐ Sonic

40 Filters.

05 ☐ Mechanical
10 ☐ Microwave
15 ☐ RF noise reduction
20 ☐ TV noise reduction
25 ☐ UHF & VHF

41 Fuses & Fuse Holders.

05 ☐ Fuses
10 ☐ Fuse holders

42 Magnets.

05 ☐ Electro
10 ☐ Permanent

43 Meters (Indicating Instruments).

05 ☐ Ammeters
10 ☐ Elapsed time
15 ☐ Frequency indicating
20 ☐ Galvanometers
25 ☐ Micro-ammeters
30 ☐ Milli-ammeters
35 ☐ Voltmeters
40 ☐ Volume level, (DB & VU)
45 ☐ Wattmeters & watt hour meters

44 Motors & Motor-Generators.

05 ☐ Blower & fan motors
10 ☐ Dynamotors
15 ☐ Frequency changers
20 ☐ Motor-generators
25 ☐ Phonograph
30 ☐ Servo
35 ☐ Synchronous motors
40 ☐ Timing motors

45 Pilot Lights & Assemblies.

05 ☐ Brackets, jewels, mounts, sockets
10 ☐ Incandescent lights
15 ☐ Neon lights

46 Power Supplies.

05 ☐ High voltage, (kilovolt)
10 ☐ Low voltage
15 ☐ Medium voltage
20 ☐ Primary power sources, fuel-engine driven
25 ☐ Voltage regulated power sources

47 Printed Circuits.

05 ☐ Etched or screened prewired chassis
10 ☐ Printed electronic assemblies
15 ☐ Solder dipped prewired chassis

48 Rectifiers, Metallic.

05 ☐ Copper oxide
10 ☐ Germanium
15 ☐ Selenium diodes
20 ☐ Selenium stacks
25 ☐ Silicon

49 Rectifiers, Vacuum Tube.

05 ☐ Grid controlled
10 ☐ Mercury vapor
15 ☐ Receiving
20 ☐ Transmitting

50 Relays.

05 ☐ Co-axial
10 ☐ D'Arsonval
15 ☐ Frequency selective
20 ☐ Hermetically sealed
25 ☐ Keying
30 ☐ Latching
35 ☐ Mercury contact
40 ☐ Polarized
45 ☐ Power control & overload
50 ☐ Stepping
55 ☐ Sub-miniature
60 ☐ Telephone
65 ☐ Time delay

51 Resistors.

05 ☐ Carbon fixed, moulded
10 ☐ Carbon, variable
15 ☐ Deposited carbon
20 ☐ Potentiometers
25 ☐ Printed circuit
30 ☐ Rheostats
35 ☐ Vacuum sealed
40 ☐ Very high megohm
45 ☐ Wirewound, fixed
50 ☐ Wirewound, variable
55 ☐ Wirewound, precision

52 Semi-Conductors.
- 05 ☐ Germanium diodes
- 10 ☐ Transistors, junction
- 15 ☐ Transistors, point contact

53 Sockets.
- 05 ☐ Capacitor, plug-in sockets
- 10 ☐ Receiving tube
- 15 ☐ Transmitting tube
- 20 ☐ UL approved industrial types

54 Switches.
- 05 ☐ Band
- 10 ☐ Coaxial
- 15 ☐ Foot
- 20 ☐ Key
- 25 ☐ Mercury
- 30 ☐ Momentary contact
- 35 ☐ Power
- 40 ☐ Precision snap-acting
- 45 ☐ Rotary
- 50 ☐ Slide
- 55 ☐ Solenoids
- 60 ☐ Spring-return
- 65 ☐ Thermally operated
- 70 ☐ Toggle & push button

55 Telegraph Keys.
- 05 ☐ Manual
- 10 ☐ Mechanical high speed
- 15 ☐ Semi-automatic

56 Thermostats.
- 05 ☐ Bellows
- 10 ☐ Bimetal
- 15 ☐ Hermetically sealed

57 Transformers.
- 05 ☐ Audio
- 10 ☐ Geophysical
- 15 ☐ Hermetically sealed
- 20 ☐ Horizontal output & flyback
- 25 ☐ Isolation
- 30 ☐ Power components
- 35 ☐ Precision matched
- 40 ☐ Pulse
- 45 ☐ Radio Frequency
- 50 ☐ Saturable
- 55 ☐ TV deflection
- 60 ☐ TV impedance matching
- 65 ☐ Variable
- 70 ☐ Voltage regulating

58 Tuners.
- 05 ☐ AM tuners
- 10 ☐ FM tuners
- 15 ☐ Microwave
- 20 ☐ TV tuners
- 25 ☐ UHF & VHF

59 Vacuum Tubes, Receiving.
- 05 ☐ Cathode-ray & TV picture
- 10 ☐ Miniature
- 15 ☐ Standard types, glass & metal
- 20 ☐ Sub-miniature

60 Vacuum Tubes, Special Purpose.
- 05 ☐ Electrometer tubes
- 10 ☐ Geiger-Mueller
- 15 ☐ Phototubes
- 20 ☐ Pirani tubes
- 25 ☐ Thyratrons
- 30 ☐ Voltage regulators & ballasts

61 Vacuum Tubes, Transmitting.
- 05 ☐ ATR & TR types
- 10 ☐ Air-cooled power tubes
- 15 ☐ Iconscopes
- 20 ☐ Image orthicons
- 25 ☐ Klystrons
- 30 ☐ Magnetrons
- 35 ☐ Monoscope tubes
- 40 ☐ Traveling wave
- 45 ☐ Water-cooled power tubes

62 Voltage Regulators.
- 05 ☐ Automatic
- 10 ☐ Manually operated
- 15 ☐ Varistors

5. Applications of Electronics, Industrial, Medical & Military.

63 Computers & Counters.
- 05 ☐ Analog to digital converters
- 10 ☐ Computers, analog
- 15 ☐ Computers, digital
- 20 ☐ Counters, continuous
- 25 ☐ Counters, pre-set

64 Electronic Control Equip.
- 05 ☐ Burglar-alarm & protective systems
- 10 ☐ Combustion & smoke-nuisance controls
- 15 ☐ Electronic controls
- 20 ☐ Production control, counting & sorting equipment
- 25 ☐ Variable speed controls

65 Geophysical Apparatus.
- 05 ☐ Amplifiers
- 10 ☐ Complete exploration equipment
- 15 ☐ Control equipment
- 20 ☐ Microphones
- 25 ☐ Recording galvanometers

66 Induction Heating Equip.
- 05 ☐ Brazing, soldering & welding
- 10 ☐ Di-electric heating
- 15 ☐ Induction heating

67 Medical Equipment.
- 05 ☐ Cardiographs
- 10 ☐ Diathermy
- 15 ☐ Electronic cauterizing
- 20 ☐ Electro-encephalographs
- 25 ☐ Stimulators

68 Nuclear Equipment.
- 05 ☐ Pulse count integrators
- 10 ☐ Radiation detectors
- 15 ☐ Scalers

69 Servo-Mechanisms.
- 05 ☐ Controls
- 10 ☐ Gears, gear-trains, etc.
- 15 ☐ Servo amplifiers
- 20 ☐ Servo motors & generators

70 Telemetering Equipment.
- 05 ☐ Data recorders
- 10 ☐ Data storage mechanisms
- 15 ☐ FM/FM modulators

6. Radar, Microwave, UHF Equipment.

71 Antennas.
- 05 ☐ Feed systems
- 10 ☐ Receiving
- 15 ☐ Reflectors
- 20 ☐ Rotating mechanisms
- 25 ☐ Transmitting

72 Receivers.
- 05 ☐ Microwave relay
- 10 ☐ Position indicators
- 15 ☐ Radar

73 Transmitters.
- 05 ☐ Microwave
- 10 ☐ Microwave relay
- 15 ☐ Radar

74 Waveguides & Accessories.
- 05 ☐ Couplings
- 10 ☐ Flexible waveguides
- 15 ☐ Hybrid junctions
- 20 ☐ Microwave attenuators
- 25 ☐ Rigid waveguides
- 30 ☐ Sampling lines, reflectors, etc.
- 35 ☐ Switches
- 40 ☐ Transmission lines

7. Materials, Fabricating Equipment & Services, Machinery & Tools, Contract Assemblers, Laboratories for Custom Built Equipment, Consulting Engineers.

75 Cable & Wire.
- 05 ☐ Aluminum
- 10 ☐ Co-axial cables
- 15 ☐ Copper, bare
- 20 ☐ Low-loss, high-frequency types
- 25 ☐ Magnet wire
- 30 ☐ Precious & rare metals wire
- 35 ☐ Pre-formed wire harnesses
- 40 ☐ Rubber insulated
- 45 ☐ Shielded
- 50 ☐ Synthetic insulated
- 55 ☐ Wire mesh
- 60 ☐ Woven wire braid

76 Ceramics.
- 05 ☐ Coil forms
- 10 ☐ Custom fabrication
- 15 ☐ Insulators
- 20 ☐ Proprietary forms

77 Consulting Engineers.
- 05 ☐ Acoustical
- 10 ☐ Electrical
- 15 ☐ Mechanical
- 20 ☐ Radio

78 Core Materials.
- 05 ☐ Laminations
- 10 ☐ Metallic powders

79 Fabricators & Services.
- 05 ☐ Contract fabrication & assembly
- 10 ☐ Electro-plating
- 15 ☐ Embedment
- 20 ☐ Hardening & heat-treating
- 25 ☐ Hermetic sealing service
- 30 ☐ Metal stamping & spinning
- 35 ☐ Packaging
- 40 ☐ Plastic fabricators
- 45 ☐ Plastic moulders

80 Gasses & Vapors.
- 05 ☐ Acetylene
- 10 ☐ Argon
- 15 ☐ Hydrogen
- 20 ☐ Krypton
- 25 ☐ Neon
- 30 ☐ Oxygen
- 35 ☐ Xenon

81 Hardware & Findings.
- 05 ☐ Adhesive labels & tapes
- 10 ☐ Bushings & bearings
- 15 ☐ Cans
- 20 ☐ Dials & dial assemblies
- 25 ☐ Flexible shafts
- 30 ☐ Gaskets
- 35 ☐ Grommets
- 40 ☐ Lacing cord & tape
- 45 ☐ Machine screw products
- 50 ☐ Nameplates
- 55 ☐ Paper preforms, for coil foundations
- 60 ☐ Retaining rings
- 65 ☐ Rivets
- 70 ☐ Screws, nuts & washers
- 75 ☐ Shock mounts
- 80 ☐ Springs

85 ☐ Strain reliefs
90 ☐ Terminal boards
95 ☐ Terminals, rivet or stud
100 ☐ Terminals, wire end
105 ☐ Vibration mounts & isolating pads

82 Hermetic Seals.

05 ☐ Ceramic to metal seals
10 ☐ Complete headers
15 ☐ Glass to metal seals

83 Insulating Materials.

05 ☐ Cambric sheeting
10 ☐ Cambric tubing
15 ☐ Ceramic
20 ☐ Cotton tapes
25 ☐ Fiberglass tapes
30 ☐ Fiberglass tubing
35 ☐ Glass bonded mica
40 ☐ Plastic
45 ☐ RF shielding materials
50 ☐ Rubber
55 ☐ Synthetics

84 Laboratories & Custom Builders.

05 ☐ Individual item builders
10 ☐ Pilot run plants
15 ☐ Research & development labs.
20 ☐ Testing & certifying organizations

85 Lacquers, Paints, Compounds & Waxes.

05 ☐ Coil dope
10 ☐ Conducting paint
15 ☐ Finishing & protecting lacquer
20 ☐ Fungus preventatives
25 ☐ Moisture proofing
30 ☐ Sealing & potting compounds
35 ☐ Silicones

86 Machinery & Tools.

05 ☐ Coil winding machines
10 ☐ Dehydrators
15 ☐ General production machinery & tools
20 ☐ Marking & engraving machines
25 ☐ Soldering irons & guns
30 ☐ Vacuum tube machinery
35 ☐ Vacuum pumps
40 ☐ Vibration exciters
45 ☐ Welders & brazers

87 Metals.

05 ☐ Alloys
10 ☐ Copper
15 ☐ Ferrous
20 ☐ Non-ferrous, excluding copper
25 ☐ Powdered
30 ☐ Precious & rare

88 Moulded Products.

05 ☐ Insulators
10 ☐ Knobs & parts
15 ☐ Proprietary mouldings

89 Plastics.

05 ☐ Extrusions
10 ☐ Raw powders
15 ☐ Rods
20 ☐ Sheets
25 ☐ Tubes

90 Solder.

05 ☐ Acid core
10 ☐ Fluxes
15 ☐ Plain
20 ☐ Precious metal
25 ☐ Pre-forms
30 ☐ Rosin core

91 Vacuum Tube Parts.

05 ☐ Anodes
10 ☐ Envelopes, glass
15 ☐ Envelopes, metal
20 ☐ Getters
25 ☐ Grids & grid wire
30 ☐ Guns & gun parts
35 ☐ Phosphors
40 ☐ Pins & prongs
45 ☐ Spacers & insulators
50 ☐ Stampings
55 ☐ Tube caps
60 ☐ Tube clamps
65 ☐ Tube shields
70 ☐ Tubing-cathode

8. Test & Measuring Equipment, Laboratory Standards & Supplies.

92 Audio Frequency Test Equipment.

05 ☐ Beat frequency oscillators
10 ☐ Distortion & noise analyzers
15 ☐ Intermodulation meters
20 ☐ Output power meters
25 ☐ RC oscillators
30 ☐ Spectrum analyzers
35 ☐ Square wave generators
40 ☐ Voltmeters-audio frequency

93 General Laboratory Equipment & Supplies.

05 ☐ Chemicals
10 ☐ General laboratory equipment
15 ☐ RF shielding materials
20 ☐ Shielded enclosures

94 General Test Equipment.

05 ☐ Bridges, capacitance
10 ☐ Bridges, impedance
15 ☐ Bridges, inductance
20 ☐ Bridges, resistance
25 ☐ Decades, capacitance
30 ☐ Decades, resistance
35 ☐ Insulation testers
40 ☐ Multimeters
45 ☐ Ohm & megohm meters
50 ☐ Strain gauges
55 ☐ Stroboscopes
60 ☐ Transistor testers
65 ☐ Vacuum tube testers
70 ☐ Vacuum tube voltmeters
75 ☐ Vibration analyzers

95 Graphic Recorders.

05 ☐ Charts & papers
10 ☐ Ink writing
15 ☐ Inkless types
20 ☐ Fixed speed types
25 ☐ Variable speed types

96 Laboratory Standards of Frequency & Time.

05 ☐ Elapsed time measuring equip.
10 ☐ Primary standards of frequency
15 ☐ Primary standards of time
20 ☐ Secondary standards of frequency
25 ☐ Secondary standards of time
30 ☐ Tuning forks

97 Microwave & Radar Test Equipment.

05 ☐ Bolometers
10 ☐ Frequency meters
15 ☐ Marker generators
20 ☐ Oscillators
25 ☐ Pulse generators
30 ☐ Slotted lines & measuring accessories
35 ☐ SWR meters
40 ☐ Time delay generators
45 ☐ Wavemeters

98 Nuclear.

05 ☐ Dosimeters
10 ☐ Ionization chambers
15 ☐ Survey meters

99 Oscilloscopes—Cathode-Ray.

05 ☐ General purpose
10 ☐ Oscilloscope cameras
15 ☐ Oscillosynchroscopes
20 ☐ Recording oscillographs
25 ☐ Special UHF & pulse analysis equipment.

100 Radio Frequency Test Equipment.

05 ☐ Field strength meters
10 ☐ Impedance & admittance meters
15 ☐ Marker generators
20 ☐ Multi-vibrators
25 ☐ Phase meters
30 ☐ "Q" meters
35 ☐ RF power output meters
40 ☐ Signal generators, AM
45 ☐ Signal generators, FM
50 ☐ Sweep generators

101 Television Test Equipment.

05 ☐ Synchronizing generators
10 ☐ TV calibrators
15 ☐ TV marker generators
20 ☐ TV signal generators
25 ☐ TV sweep generators

9. Education & Publishing.

102 Education.

05 ☐ Technical instruction — home study
10 ☐ Technical instruction — resident

103 Publishing.

05 ☐ Book publishers
10 ☐ Data sheet & manual preparation
15 ☐ Magazines

10. Distributional Functions.

104 Distribution.

05 ☐ Export & Import
10 ☐ Jobbers & Wholesalers
15 ☐ Manufacturers Representatives
20 ☐ Sales & Service
25 ☐ Surplus dealers

LIST ADDITIONAL PRODUCTS BELOW

_____

_____

_____

_____

_____

_____

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 32A)

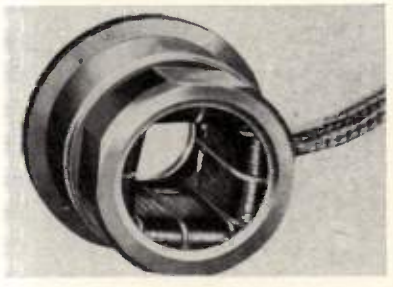
Delay Line Kit

The James Millen Manufacturing Co., 150 Exchange St. Malden, Mass., has developed a delay line kit which provides a means for the development and design engineer to check the effect of various delays in their actual developmental setups without the time loss and expense of producing separate lines for each trial.

The kit is designed to provide a ready means of obtaining various delays from 0.10 μsec through 2 μsec in increments of 0.05 μsec except at the extreme ends of this range. The lines may be used repeatedly without deterioration because they are hermetically sealed—the smaller lines in glass tubes, the 1-μsec line in a metal container.

Magnetic Deflection Yoke

A new magnetic deflection yoke, Type Y15-5, for military and oscilloscope applications has been designed by **Syntronic Instruments, Inc.,** 100 Industrial Rd., Addison, Ill. Type Y15-5 is suited to push-pull applications requiring normal, high or extremely high sensitivity. Type Y15-5 is of the square core, parallel opposed magnetic field type and fits 1½ inch neck diameter cathode-ray tubes. Deflection angle is up to 50°.



Type Y15-5 is available over a wide range of impedances in single ended or push pull combinations. The following typical specifications apply for a single ended yoke coil. The same specifications also apply for the coil between one plate and B (or center tap) of a push-pull yoke: Inductance 140 mh, resistance 145 ohms, sensitivity 80 ma per radius—25° at 10 kv accelerating potential (sensitivities as low as approximately 7 ma available), 99 per cent recovery in 40 microseconds. Catalog available on request.

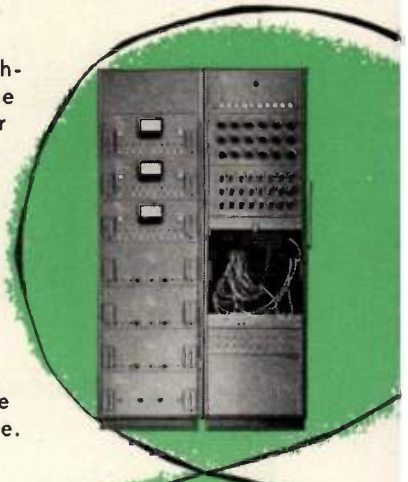
(Continued on page 56A)

Save Engineering Time

ON • Equation Solving • System Simulation • Testing

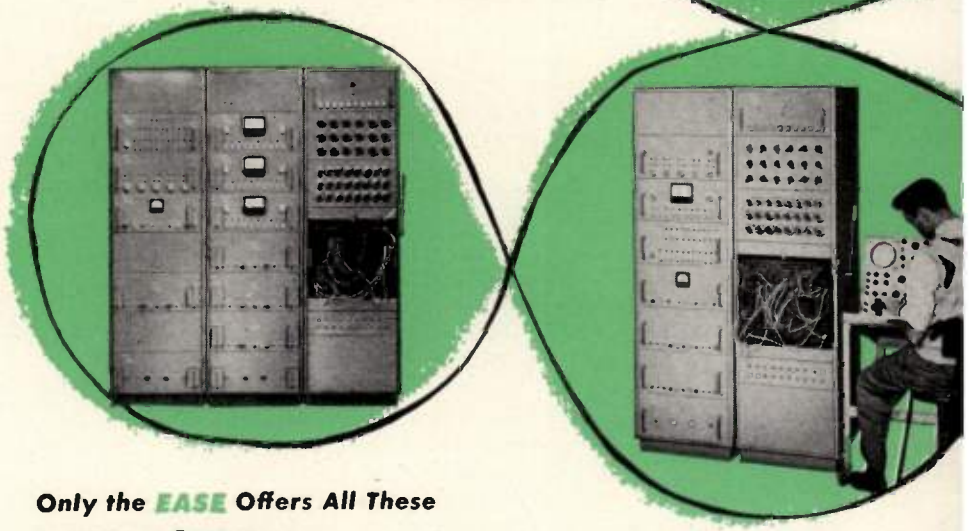
do it with.. **EASE***

***Electronic Analog Simulating Equipment**



As a computer, the EASE solves high-order differential equations in a fraction of the time required for slide-rule and calculator methods. As a system simulator, it can represent the functioning of an entire dynamic system. Effects of changing loads, forces and other variables can be determined rapidly and accurately.

In testing, a physical component may be studied in a complete system through the use of voltage transducers. In this manner, where the EASE simulates the associated equipment, fast yet thorough testing over a wide range of conditions is possible.



Only the EASE Offers All These Important Features:

1. **SIMPLICITY** . . . any engineer or mathematician who can write the equation can solve it on the EASE with only a few hours' training.
2. **VERSATILITY** . . . Additional components may be added at any time to extend its range of application. Its function generator can also serve as a resolver with an accuracy of better than 1%. It can be operated in real time or as a repetitive computer, or the problem can be "held" at any stage.
3. **HIGH ACCURACY** . . . phase shift is less than 1° at 3500 cps. With a 10-channel amplifier stabilizer (optional), drift is less than 100 microvolts referred to grid. Meters on amplifiers permit extremely accurate potentiometer calibration.
4. **COMPACTNESS** . . . unit requires less than 8 sq. ft. of floor space.
5. **EASY INSTALLATION** . . . contains its own regulated power supply; normally merely plugs in to 20 amp, 110 v. 60-cycle a.c. line!
6. **LOW COST** . . . the EASE is the world's first high-quality computer to be mass-produced in practical commercial form. The result is low cost without sacrifice in utility or quality.

For complete data, request Bulletin N2.

Berkeley *division*

BECKMAN INSTRUMENTS INC.

2200 WRIGHT AVE., RICHMOND, CALIF.

M-20

zero to 70,000 cycles per second

...recorded on one tape recorder, the Ampex Model 311

This versatile Ampex recorder is a combination of a wide-range direct recorder for high frequency phenomena and an fm-carrier recorder for highly transient or extremely low frequency phenomena requiring excellent amplitude accuracy. It provides a means for recording nearly every type of data encountered in laboratories and in industry.

The Ampex 311 uses two parallel channels on quarter-inch tape. The direct recording channel handles frequencies from 300 to 70,000 cycles per second. The fm-carrier channel has high transient accuracy in the range from 0 to 5,000 cps.

The Ampex 311 can record diverse types of data signals with a common time base. For instance, the fm channel might record high transient phenomena or low frequency signals while the direct channel records wide-band multiplexed signals or other high frequency information.

Even where serving only one need, the Ampex 311 is effectively a two-channel recorder, since the channel not being used to record data can be used for time signals. Thus, it can serve the same purposes in many cases as either a two-channel direct recorder or a two-channel fm-carrier recorder.

AMPEX Model 311 Combination Recorder
Two parallel channels cover a wide range of data situations:

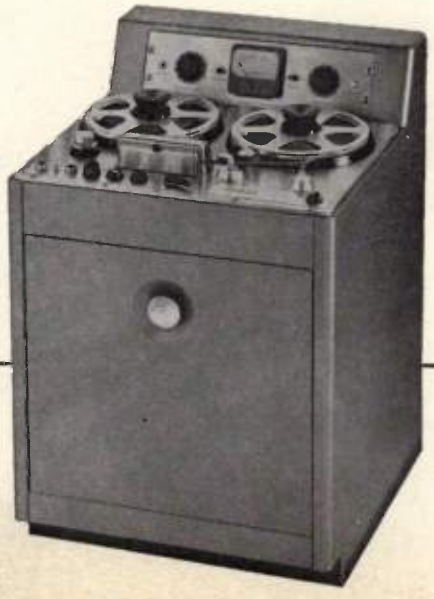
The fm channel

- Frequency response, 0 to 5,000 cps. ± 1 db.
- Transient accuracy independent of minor tape flaws and irregularities.

The direct channel

- Frequency response, 300 to 70,000 cps. ± 3 db.
- Wide band multiplexed or fm-carrier data
- High frequency phenomena

For further information write to Dept. G-1370-RB.



AMPEX
CORPORATION

AMPEX CORPORATION
934 CHARTER STREET • REDWOOD CITY, CALIFORNIA
Branch offices: New York, Chicago, Atlanta, San Francisco and 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

Hysteresis Motors

by COLLINS

New 370A Series Motors Designed Primarily For Operation From Vacuum Tubes

OUTSTANDING FEATURES of the latest addition to the Collins line of Industrial Components:

- high starting torques
- frequency ranges: 0 - 1000 cps, 0 - 500 cps, 60 cps, 400 cps
- absolute synchronous rotation

The superior efficiency and high starting torque of the Collins 370A-Series Hysteresis Motors make them suited for driving timing mechanisms, magnetic storage drums or any other device which must rotate at an absolutely constant speed regardless of fluctuations in load or line voltage. These motors have split windings for operation directly from the plate circuits of two push-pull direct-coupled amplifiers driven 90 degrees out of phase, eliminating bulky, frequency-limiting output transformers.

Typical of this latest Collins Industrial Component is the Type 370A-1 Wide-Band Hysteresis Synchronous Motor, particularly adaptable to operation in automatic frequency control systems with error signals from zero to as high as 1000 cps producing synchronous rotation from motionless to 30,000 rpm.

Other standard models of the 370A-Series include:

Type 370A-2; similar to 370A-1 but with higher torque; 0-500 cps frequency range.

Type 370A-3; single frequency, 60 cps; 1800 rpm.

Type 370A-4; single frequency, 400 cps; 12000 rpm.

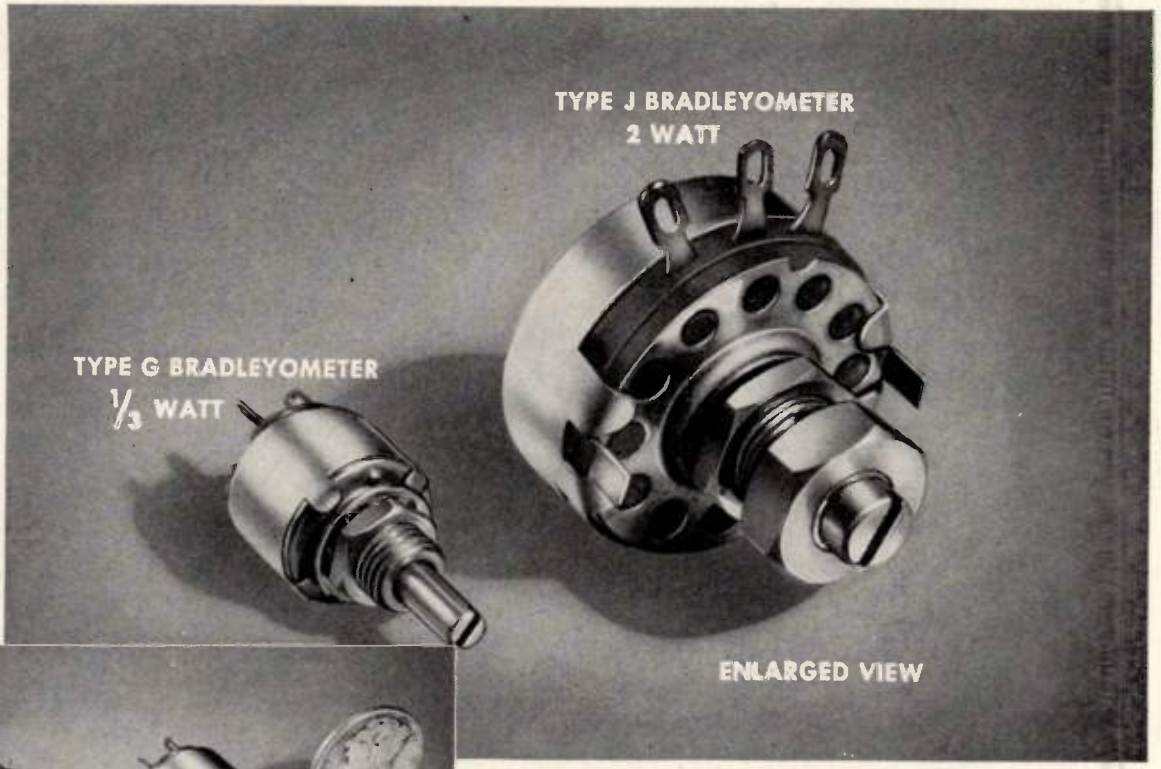
Write today for complete information.

COLLINS RADIO COMPANY
Cedar Rapids, Iowa

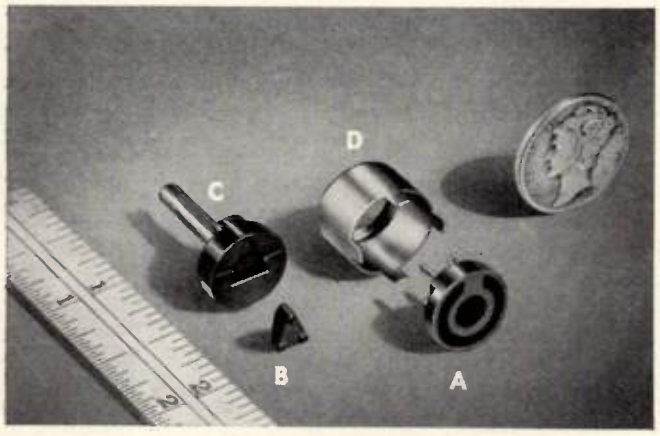
11 W. 42nd Street, NEW YORK 36

1930 Hi-Line Drive, DALLAS 2

2700 W. Olive Avenue, BURBANK



Type G Bradleyometers Made with two types of bushings . . . the standard bushing and straight shaft for knob control, and the split bushing with lock nut and slotted shaft for shaft lock applications. Metal parts are all of corrosion resistant materials.



Type G Bradleyometer Component Parts

A—Resistor element—molded in one piece with resistor and terminals. No rivets or welds—no maintenance problems.
B—Carbon brush—which fits into triangular opening in rotor and makes contact with resistor ring at any point.
C—Rotor and shaft — showing triangular opening or recess for holding the brush; a spring keeps brush firmly against the resistor element.
D—Metal case made of corrosion resistant metal. Diameter—½ inch.

SMALLEST MOLDED ADJUSTABLE RESISTOR

• ONLY ½" DIAMETER •

for Rheostat or Potentiometer Service

The Type J Bradleyometer has long been recognized as the top quality, 2-watt, adjustable resistor for electronic applications. Since it has a solid MOLDED composition resistor . . . and is not a paint or spray type unit . . . its characteristics remain permanent even after long use.

The new, Type G Bradleyometer . . . only ½ inch in diameter and rated at ⅓ watt . . . offers the same advantages of construction and noiseless performance as the bigger, Type J unit. Its solid MOLDED resistor element can be made to satisfy any resistance-rotation curve. The carbon contact brush assures quiet operation, even improving with long service.

Maximum or total resistance values range from 100 ohms to 5 megohms. Maximum continuous power for the entire element in circuit is 0.33 watt. Maximum continuous volts—350 RMS. Maximum current—0.1 ampere.

Samples can be supplied for qualification tests.

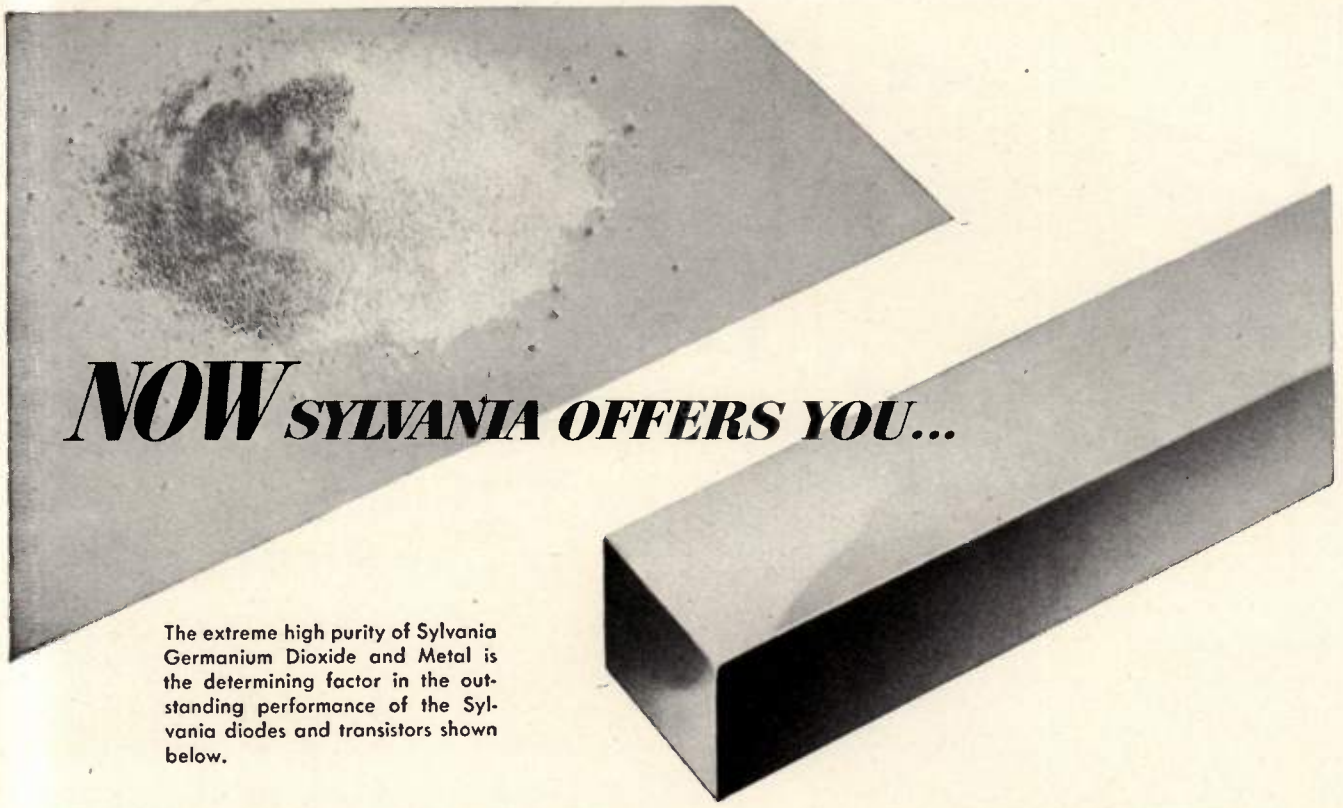
Allen-Bradley Co.
114 W. Greenfield Ave., Milwaukee 4, Wis.

ALLEN-BRADLEY

FIXED & ADJUSTABLE RADIO RESISTORS

QUALITY

Sold exclusively to manufacturers of radio and electronic equipment

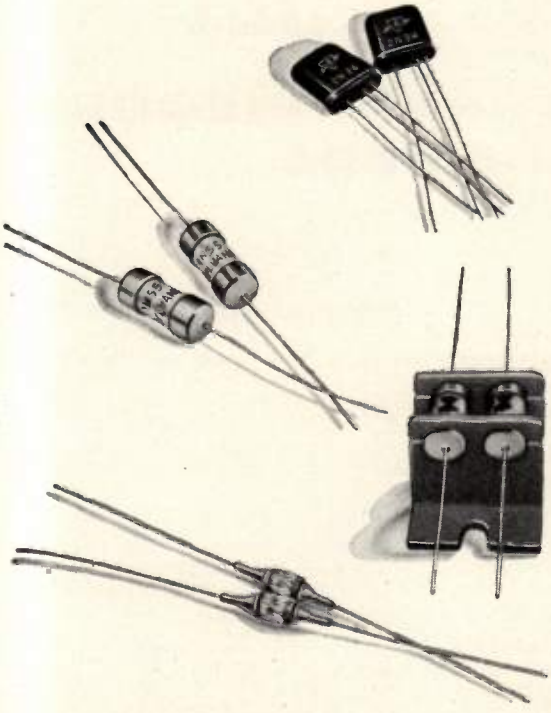


NOW SYLVANIA OFFERS YOU...

The extreme high purity of Sylvania Germanium Dioxide and Metal is the determining factor in the outstanding performance of the Sylvania diodes and transistors shown below.

GERMANIUM DIOXIDE AND METAL

OF HIGHEST CHEMICAL PURITY



Sylvania Germanium Dioxide boasts a purity so high that any critical impurities are not detectable even spectrographically.

Sylvania uses every necessary means of quality control to achieve this unexcelled purity. This close control plus years of metallurgical research and engineering skill insure an absolute minimum content of those impurities in Sylvania Germanium Dioxide which would retard purification when it is reduced to metal and zone purified.

Why extreme purity is essential

Sylvania's famous Germanium Diodes and Transistors are recognized leaders in the electronic industry. Therefore, Sylvania Germanium Dioxide must have the proper grade of purity to maintain unbeatable performance.

So, when you need the very finest, easiest-to-work-with Germanium Dioxide, make sure you get it from Sylvania. Or, if you need Germanium metal, you can obtain it in ingot form with resistivity of 30 ohm/cm or higher in polycrystalline or single crystal material . . . types used in Sylvania's semi-conductors and electronic components, world-famous for finest performance. Write today for further information.

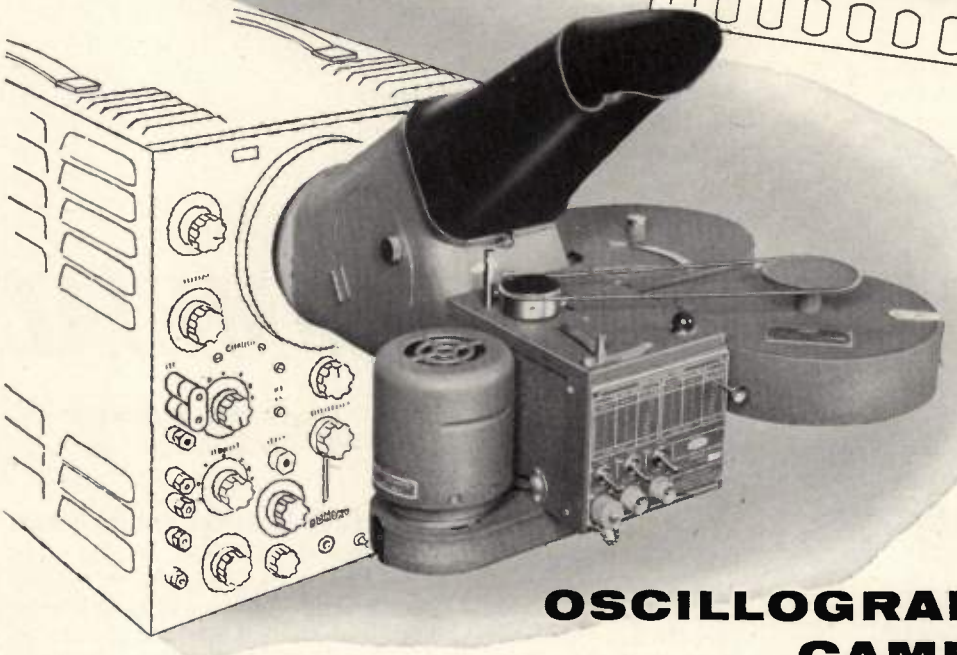
LIGHTING · RADIO · ELECTRONICS · TELEVISION

SYLVANIA

Sylvania Electric Products Inc., Dept. 4T-3102, 1740 Broadway, New York 19, N. Y.

In Canada: Sylvania Electric (Canada) Ltd.
University Tower Bldg., St Catherine Street, Montreal, P Q.

for all continuous-motion or single-frame OSCILLOGRAPH RECORDING



the TYPE 321

OSCILLOGRAPH-RECORD CAMERA

• Continuous motion recording over the entire range of laboratory applications, including the most highly specialized investigations • Film speeds from 1 to 10,800 inches per minute. Film motion either horizontal or vertical • Overall accuracy of film speed guaranteed within 2% at proper line frequency • Friction film drive for simple loading. Permits use of perforated or unperforated film or paper • Single-frame recording of any phenomenon, including high-speed, single transients. Coated f/1.5 lens enables recording of spot-writing rates up to 35 inches/μsec from cathode-ray tube operated at 12,000 volts • Built-in timing lamp • Built-in illuminated data card • Simultaneous viewing and recording • Mounts quickly and easily on any standard 5-inch cathode-ray oscillograph • Delivery FROM STOCK.

(For 60-cycle power) PRICE $875

SPECIFICATIONS

Optical System: Wollensak f/1.5 coated Raptar lens. Image reduction ratio 4.5.

Shutter: Permits "Time" and "Bulb." *Provision for remote operation.*

Writing-rate Capability: 35 in./μsec with Type P11 screen at 12,000 volts.

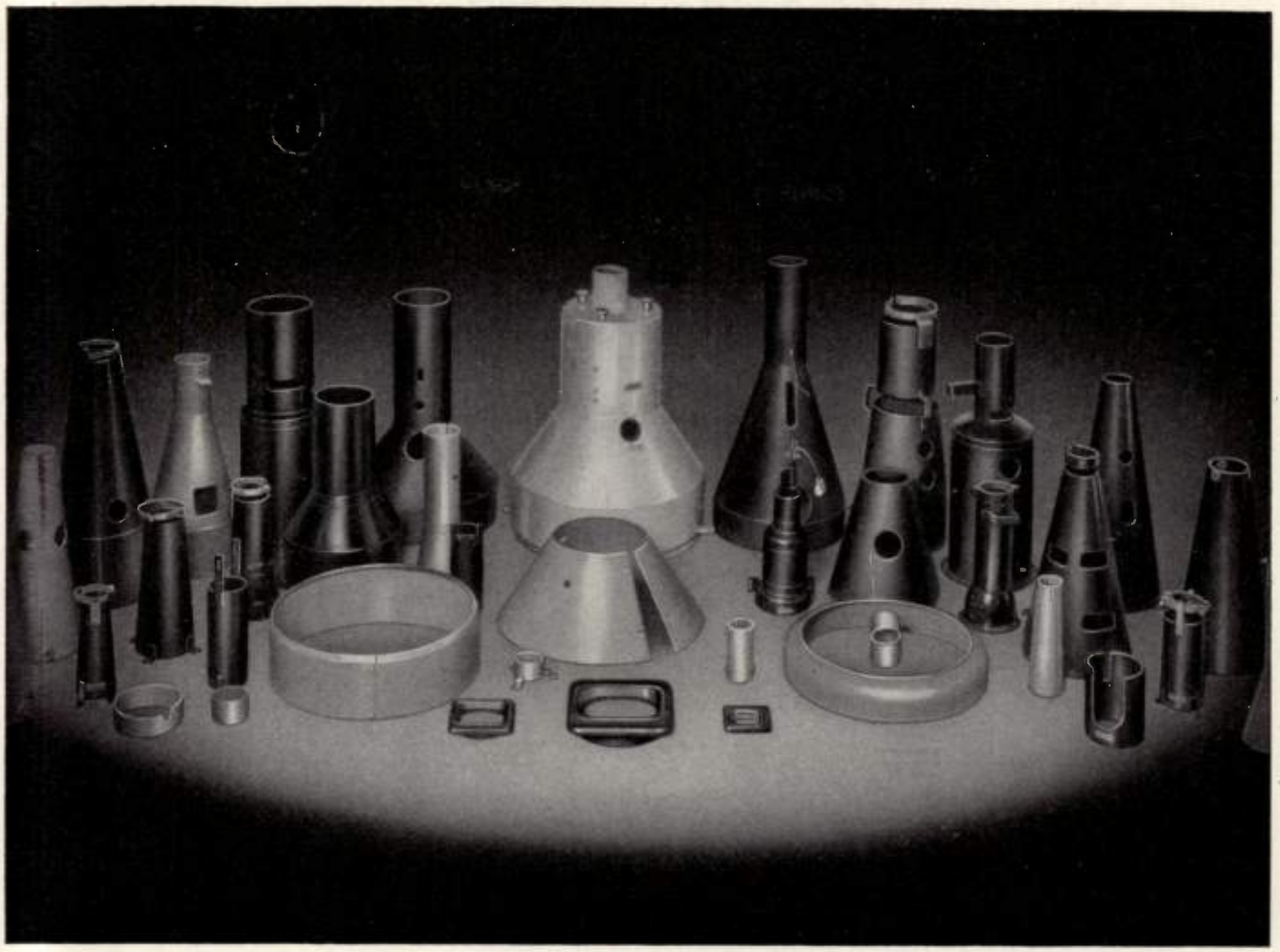
Film Speed: Variable in steps of: 0.8; 1.6; 2.5; 4.9; 7.4; 14.3; 22.2; 44.4; 66.6; 133.3; 200; 400; 600; 1200; 1800; and 3600 in/min with rolls up to 400 feet in length. Also 5400 and 10,800 in/min wth film strips up to 10 feet in length. Accuracy of film speed within 2% at proper Line frequency.

Recording Material: Perforated or unperforated 35 mm film or recording paper in lengths up to 400 ft.

WRITE *for "Techniques of Photo-Recording," a 36-page handbook on cathode-ray photography.*

DU MONT

ALLEN B. DU MONT LABORATORIES, INC., INSTRUMENT DIVISION, 760 BLOOMFIELD AVE., CLIFTON, N J.



Designed for Application

Mu Metal Shields

The James Millen Mfg. Co. Inc. has for many years specialized in the production of magnetic metal cathode ray tube shields for the entire electronics industry, supplying magnetic metal shields to manufacturing companies, laboratories and research organizations. Stock shields are immediately available for all of the more popular sizes and types of cathode ray tubes as well as bezels for 2″, 3″ and 5″ size tubes.

Many production problems, however, make desirable special shields designed in conjunction with the specialized requirement of the basic apparatus. Herewith, are illustrated a number of such custom built shields. Our custom design and fabrication department is at the service of our customers for the development and manufacture of magnetic metal shields of either nicoloi or mumetal for such specialized applications.

JAMES MILLEN MFG. CO., INC.

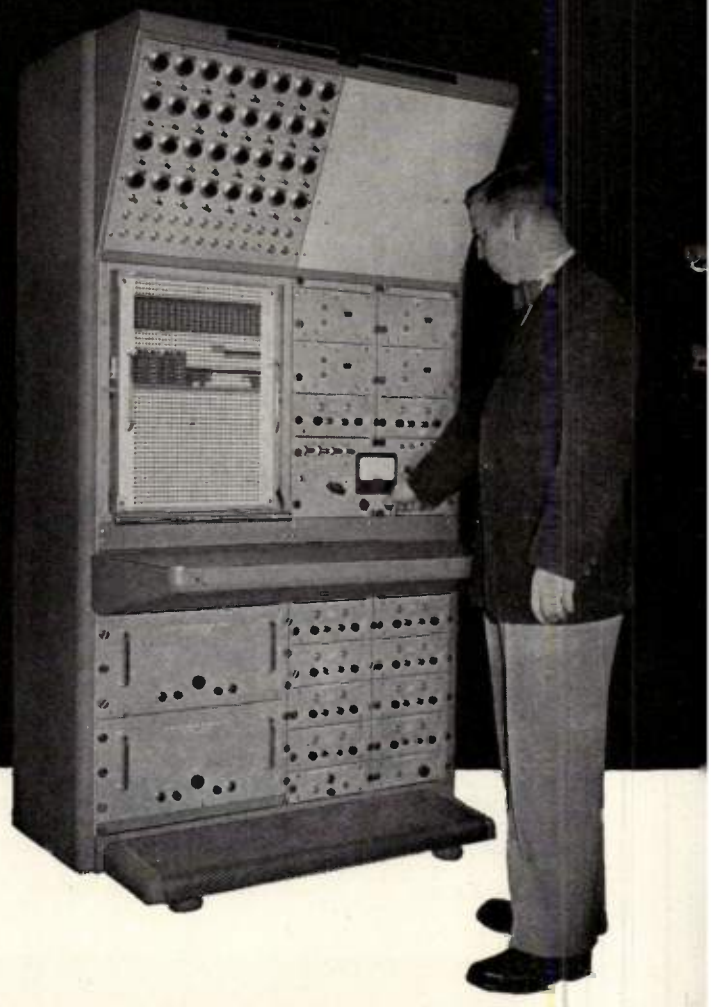
MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.

OUR FIRST 1954 ANNOUNCEMENT

We Proudly Present

A NEW ANALOG COMPUTER
A NEW LOW IN PRICE
A NEW HIGH IN QUALITY
A NEW ADVANCE IN ACCURACY



TYPE 16-31R COMPUTER

Twenty Operational Amplifiers • Four Servo Multipliers • Removable Metal Patch Panel • Controlled Network Environment • Six Diode Limiters • Two Function Relays • Thirty Two Attenuators • Electronically Regulated Power Supplies • Provisions for Future Expansion

With this highly reliable computer, which is efficient and easy to operate, you can obtain accurate answers to your complex engineering problems quickly and conveniently. Requests for detailed information will receive prompt attention.

WRITE DEPT. – R

ELECTRONIC ASSOCIATES Incorporated

ELECTRONIC ASSOCIATES INC

LONG BRANCH • NEW JERSEY

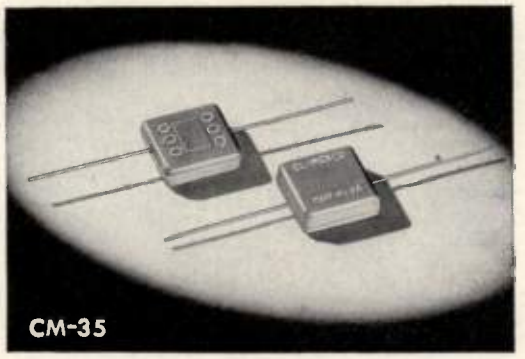
APPLICATIONS FOR ENGINEERING POSITIONS ARE WELCOME

Capacitors shown 2½ times actual size

CM-15 EL-MENCO CAPACITORS are only 9/32"x1/2"x3/16"... but they give DOUBLE VALUE PER DOLLAR

ALL fixed mica El-Menco capacitors are *factory-tested at double their working voltage.* Couldn't that mean that they'll last twice as long as other capacitors which cost no less? They also meet all significant requirements of JAN-C-5. So, you can depend on them to perform perfectly on all military and civilian electronic applications. Our tiny CM-15 silvered mica capacitors come in capacities from 2 to 420 mmf. at 500vDCw — 2 to 500 mmf. at 300vDCw. Our other types run all the way up to 10,000 mmf. It will pay you to compare El-Menco capacitors with all others — performance-wise, price-wise. *The Electro Motive Manufacturing Co., Inc., Willimantic, Conn.*

Electro Motive is now supplying special silvered mica films for the electronic and communication industries in any quantity — just send us your specifications.



WRITE FOR FREE SAMPLES AND CATALOG ON YOUR FIRM'S LETTERHEAD

Jobbers and distributors are requested to write for information to Arco Electronics, Inc., 103 Lafayette St., New York, N. Y. — large stocks on hand — spot shipments for immediate delivery. Sole Agent for Jobbers and Distributors in U. S. and Canada.

El-Menco

MOLDED MICA CAPACITORS MICA TRIMMER

Foreign Electronic Manufacturers Get Information Direct from our Export Dept. at Willimantic, Conn.

THE ELECTRO MOTIVE MFG. CO., INC. **WILLIMANTIC, CONNECTICUT**

precision electronic instruments by DeJUR

PANEL INSTRUMENTS

1. 2½" Ruggedized Meter. Meets MIL-M-10304. Model R210
2. 3½" A.S.A. Meter meets MIL-M-6A. Model S312
3. 1½" Meter. Conserves space. Built to SC-73-3. Model 112
4. 1½" Meter. Metal Case. Single Hole Mounting. Model 120
5. 1½" Ruggedized Meter. Meets MIL-10304. Model R112

PRECISION POTENTIOMETERS

6. Multiple ganging and external phasing 2" Diameter Model C200
7. Single hole mounting Model C200
8. Single Sine-Cosine Model C200
9. Dual Unit Model L402
10. 1-11/16" Diameter Center tapped Model L400
11. 3" Diameter 11 watt JAN-R-19 Const. Model 275
12. Dual Concentric Model 282
13. 3" Diameter 8 watt JAN-R-19 Const. Model 260
14. Non-Linear Card C200
15. Sine-Cosine Card C200

Actual performance records prove that these DeJUR components withstand adverse conditions of vibration, heat and moisture. Each is engineered and manufactured to meet rigid government requirements. In addition to its wide variety of stock instruments, DeJUR offers top-flight laboratory, engineering and manufacturing facilities for production of these precision units adapted to your specifications. Inquiries are invited.

Write for more detailed information on any of the products shown on this page to Dept. 1P-2.

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MANUFACTURERS OF SCIENTIFIC PRECISION EQUIPMENT FOR OVER 30 YEARS

See the DeJur line at Booth 200, Production Road, Radio Engineering Show, March 22-25.

THE THIRD NEW PLANT IN THREE YEARS AGAIN PERMITS

Fast Delivery

ALSIMAG®

CUSTOM MADE TECHNICAL CERAMICS

IN ANY QUANTITY

DIE-PRESSED
MACHINED
EXTRUDED
METAL-CERAMIC COMBINATIONS

MORE THAN 50 YEARS OF KNOW-HOW MEANS ***ASSURED QUALITY***

WIDEST CHOICE OF CERAMIC MATERIALS TO MEET YOUR SPECIFIC REQUIREMENTS

Send blue prints and specifications and let us show you what we can do for you.

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AMERICAN LAVA CORPORATION

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WRH

Stop Troubles Before They Start... With Mallory Vibrators

The best time to do this is while your equipment is still in the design stage. Each element . . . the vibrator, transformer and buffer capacitor . . . must be carefully selected for balanced electrical characteristics if your power supply is to give top performance in service.

You can avoid vibrator power supply trouble by calling on the specialized knowledge and experience of Mallory engineers. Let them translate the power requirements of your equipment into a smooth operating, trouble-free design. You will save time and money and get the kind of performance you want.

Why call on Mallory? There are a lot of reasons . . . good ones. Our experience in this field is backed by an unmatched fund of engineering knowledge that started over 20 years ago when we produced the *first* commercial Vibrator. Our experience includes supplying more Vibrators for original equipment than all other makes combined.

That's not all. If you wish, we are equipped to design and manufacture complete power supply units . . . to your exact requirements . . . to meet your production schedules.

To save engineering time and reduce production costs, write us today. It is the best way to stop troubles before they start.

Expect more . . . Get more from **P.R. MALLORY & CO. Inc. MALLORY**

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

Serving Industry with These Products:

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HUGHES

DIODES

A New Standard of Reliability

Reliability in a germanium diode is determined principally by permanent freedom from the two major causes of diode failure—moisture penetration of the diode envelope, and electrical instability under extreme operating conditions.

HUGHES GERMANIUM DIODES are designed to prevent such failures through two exclusive features:

1. **Fusion Sealing**—The glass-to-metal seal, proved in billions of vacuum tubes, is incorporated to full advantage in diode manufacture by the Hughes-developed process of fusion sealing at high temperature. The result is a rigid *one-piece* glass envelope impervious to moisture.

2. **100% Testing**—Hughes 100% testing procedures *invite* instabilities to occur prior to shipment, assuring rejection of defective diodes. *Each* **HUGHES DIODE** is humidity-cycled, temperature-cycled, JAN shock-tested, and electrically tested under vibration. This testing procedure insures operation of **HUGHES DIODES** under adverse conditions of moisture, temperature, vibration and severe shock.

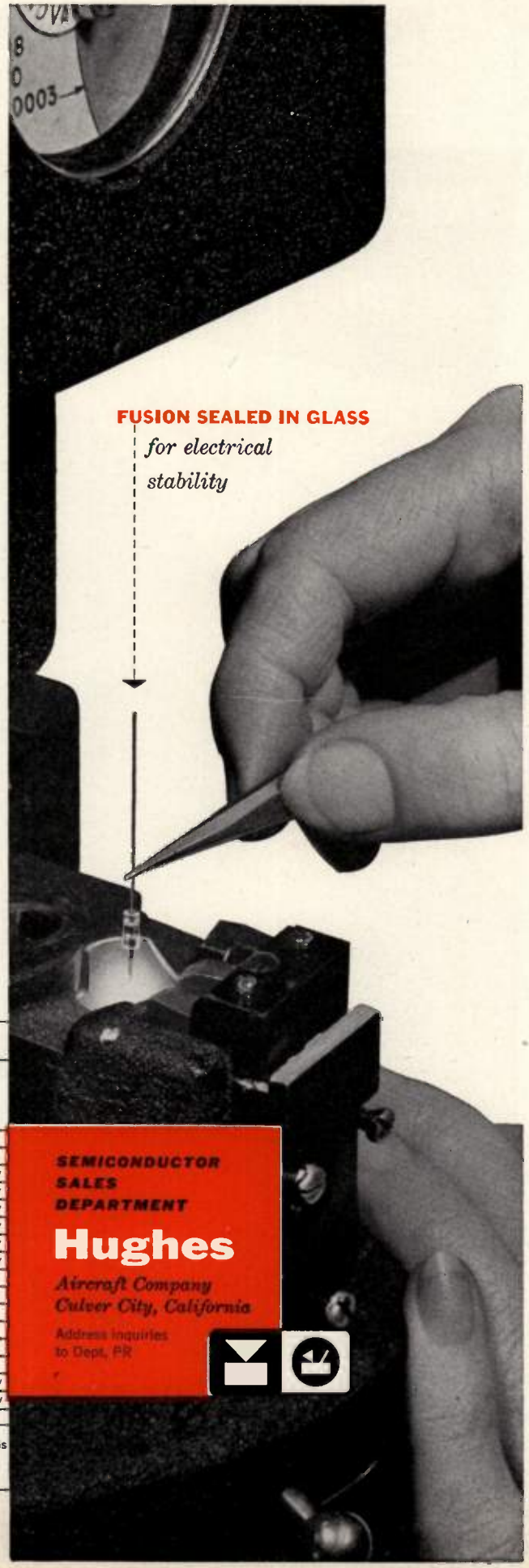
Reliability of **HUGHES DIODES** has been proved in advanced airborne military radar and fire control systems, and for guided missiles.

HUGHES GERMANIUM DIODE ELECTRICAL SPECIFICATIONS AT 25° C.

Description	RETMA Type	Test Peak Inverse Voltage* (volts)	Maximum Inverse Working Voltage (volts)	Minimum Forward Current @ +1 v (ma)	Maximum Inverse Current (ma)	
High Peak	1N55B	190	150	5.0	0.500 @ −150 v	
	1N68A	130	100	3.0	0.625 @ −100 v	
High Back Resistance	1N67A	100	80	4.0	0.005 @ −5 v;	0.050 @ −50 v
	1N99	100	80	10.0	0.005 @ −5 v;	0.050 @ −50 v
	1N100	100	80	20.0	0.005 @ −5 v;	0.050 @ −50 v
High Back Resistance	1N89	100	80	3.5	0.008 @ −5 v;	0.100 @ −50 v
	1N97	100	80	10.0	0.008 @ −5 v;	0.100 @ −50 v
	1N98	100	80	20.0	0.008 @ −5 v;	0.100 @ −50 v
High Back Resistance	1N116	75	60	5.0	0.100 @ −50 v	
	1N117	75	60	10.0	0.100 @ −50 v	
	1N118	75	60	20.0	0.100 @ −50 v	
General Purpose	1N90	75	60	5.0	0.800 @ −50 v	
	1N95	75	60	10.0	0.800 @ −50 v	
	1N96	75	60	20.0	0.800 @ −50 v	
JAN Types	1N126**	75	60	5.0	0.050 @ −10 v;	0.850 @ −50 v
	1N127†	125	100	3.0	0.025 @ −10 v;	0.300 @ −50 v
	1N128‡	50	40	3.0	0.010 @ −10 v	

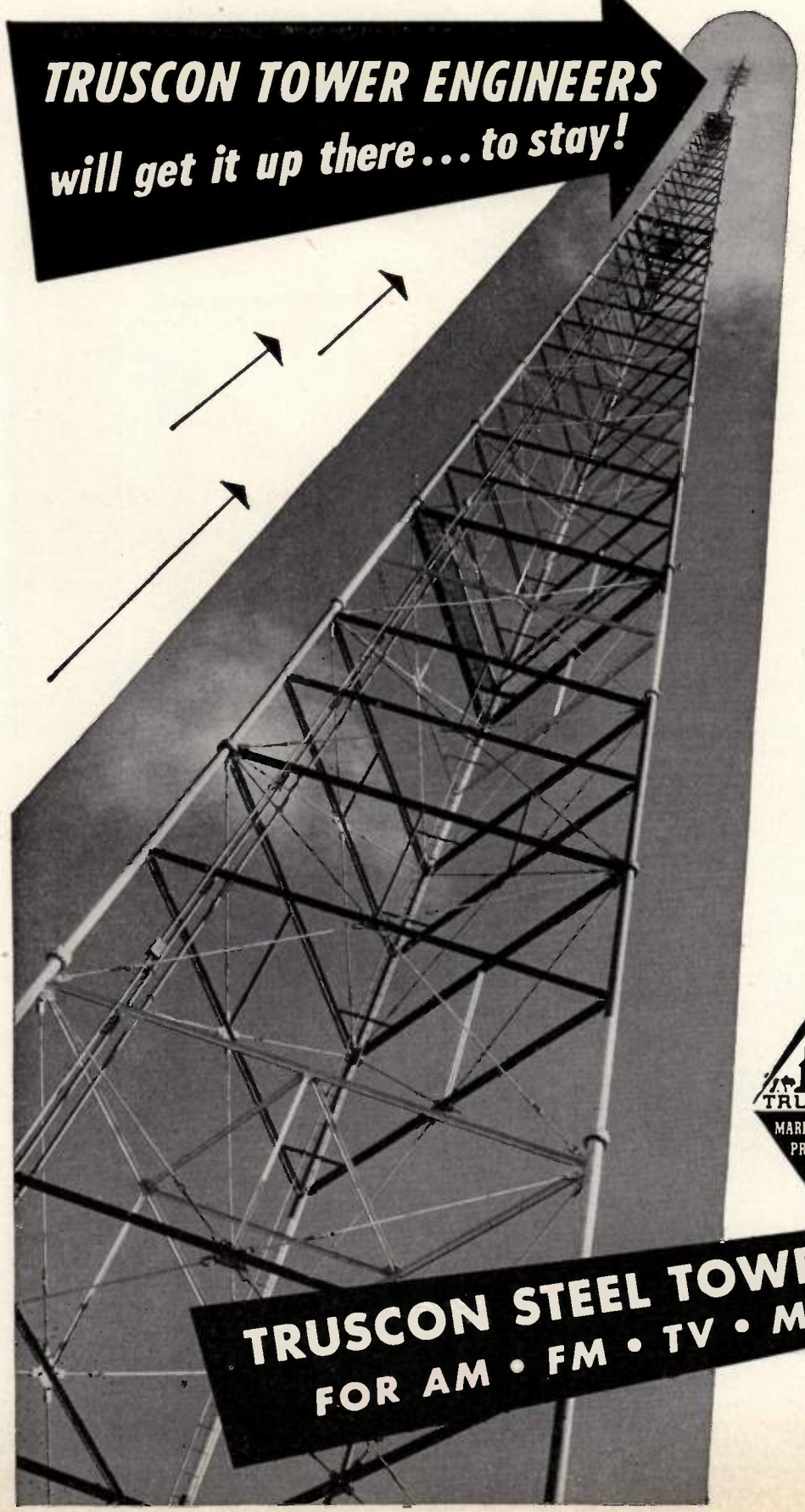
*That voltage at which dynamic resistance is zero under specified conditions. Each Hughes Diode is subjected to a voltage rising linearly at 90 volts per second.
**Formerly 1N69A. †Formerly 1N70A. ‡Formerly 1N81A.

HUGHES DIODES *are also supplied 100% factory-tested to a wide range of customer-specified characteristics, including high-temperature requirements.*



You Name The Height Your Antenna Must Reach...

TRUSCON TOWER ENGINEERS will get it up there... to stay!



● Whether your needs call for a medium-sized 300 foot tower, a lofty 1200-plus foot giant, or any size in between, Truscon engineers have the answer . . . or can get it to you fast. Truscon has designed and built many hundreds of radio towers now serving in all kinds of climate and all types of topography. Truscon builds 'em for you tall or small . . . guyed or self-supporting . . . tapered or uniform in cross-section . . . for AM, FM, TV, and Microwave transmission. You just name the height your antenna must reach; then write, wire or 'phone your nearest Truscon district office or "tower headquarters" in Youngstown to get your tower program going as soon as defense requirements permit.

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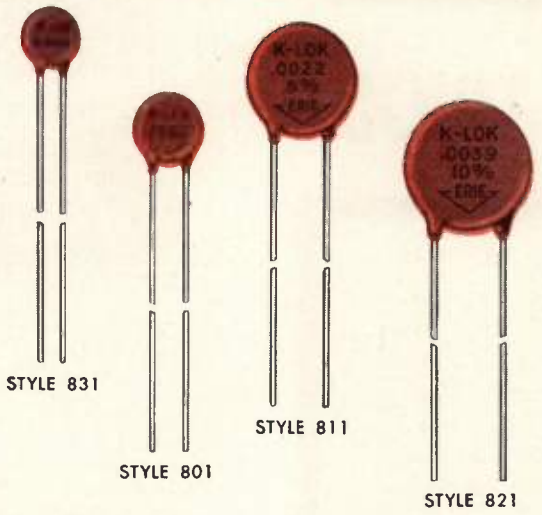
See Truscon's Exhibit of radio towers in booth 327, Computer Avenue, IRE Convention, March 22-25.

The *ultimate* in Flat Temperature Response in high-K

ERIE K-LOK*



STYLE	CAPACITANCE RANGE—MMF
831	220-530
801	530-820
811	820-2400
821	2400-4700
Tolerances: ±5%, ±10%, ±20%.	

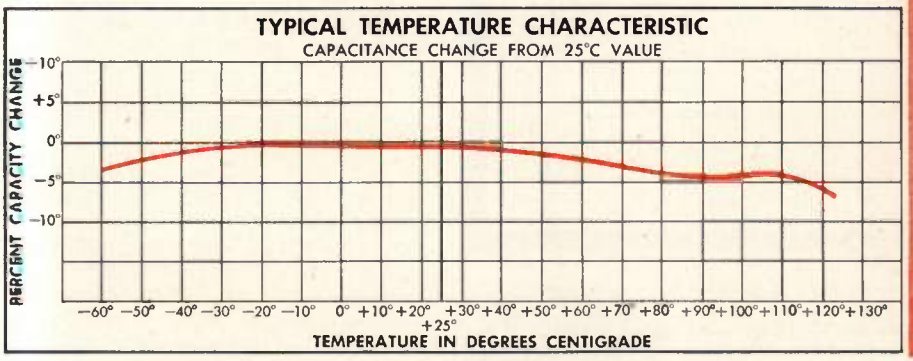
The ERIE K-LOK marks a significant advance in ceramic capacitor design and application. It marks the culmination of several years of development activity with the single objective of producing a ceramic capacitor with truly premium performance characteristics.

NOTE THESE FEATURES:

- Flat temperature characteristic. Greatest capacitance change from 25°C value:
  +5%, −5% from −55°C to +105°C
  +5%, −10% from −55°C to +125°C
- Stability with age: 1% maximum decrease in capacitance per decade of time
- Low voltage coefficient: Capacity change less than 5% from zero to rated voltage
- Negligible piezo-electric effect
- Life rating: Up to 1000 VDCW at 85°C; 500 VDCW at 125°C
- Power factor: 1½% maximum
- Insulation resistance: 10,000 megohms minimum
- Low loss phenolic dip for moisture seal and mechanical ruggedness
- ERIE K-LOK can be used on special order in supplying various high voltage disc capacitors and printed circuits

Samples available on request

*TRADE MARK



ERIE components are stocked by leading electronic distributors everywhere.

ERIE RESISTOR CORP. TRADE MARK

ERIE RESISTOR CORPORATION . . . ELECTRONICS DIVISION

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Ruggedized and aged "RELIABLE" DOUBLE TRIODE

Do you have an aircraft or industrial application that requires *utmost* dependability in increasing or controlling alternating voltages or powers . . . in changing electrical energy from one frequency to another . . . or in generating an alternating voltage?

If so, specify the Red Bank RETMA 6385 "Reliable" Double Triode. For it is specially ruggedized to perform at top efficiency longer, even under operating conditions of severe shock and vibration. And, as further assurance of its extra reliability, each RETMA 6385 is factory-aged with a 45-hour run-in under various overload, vibration and shock conditions, such as it might meet on the job.

Whether you need tubes as amplifiers, mixers, or oscillators, it will pay you to investigate the superior, longer-lasting performance qualities of the Bendix Red Bank RETMA 6385.

RATINGS*

Heater voltage—(AC or DC)**	6.3 volts
Heater current	0.50 amps.
Plate voltage—(max.)	360 volts
Max. peak plate current (per plate)	25 ma.
Max. plate dissipation (per plate)	1.5 watts
Max. peak grid voltage	+ 0 volts −100 volts
Max. heater-cathode voltage	300 volts
Max. grid resistance	1.0 megohm
Warm-up time	45 sec.

(Plate and heater voltage may be applied simultaneously.)

*To obtain greatest life expectancy from tube, avoid designs where the tube is subject to all maximum ratings simultaneously.

**Voltage should not fluctuate more than ±5%.

PHYSICAL CHARACTERISTICS

Base	Miniature button 9-pin
Bulb	T-6½
Max. over-all length	2⅜ in.
Max. seated height	1 15/16 in.
Max. diameter	⅞ in.
Mounting position	Any
Max. bulb temp.	160° C

AVERAGE ELECTRICAL CHARACTERISTICS

Heater voltage, E_f	6.3 volts
Heater current, I_f	0.50 amps.
Plate voltage, E_b	150 volts
Grid voltage, E_c	−2.0 volts
Plate current, I_b	8.0 ma.
Mutual conductance, g_m	5000 μmhos
Amplification factor, μ	35
Cut-off voltage	−10 volts
Direct interelectrode capacitances (no shield)	
Plate-grid (per section)	1.7 μμf
Plate-cathode (per section)	1.1 μμf
Grid-cathode (per section)	2.4 μμf
Plate-plate	0.1 μμf

Bendix Red Bank

Manufacturers of Special-Purpose Electron Tubes, Inverters, Dynamotors and Fractional HP D.C. Motors

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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 41A)

Envelope Delay Curve Tracer

The Model 1603-AR Envelope Delay Curve Tracer, designed by **Telechrome, Inc.**, 88 Merrick Road, Amityville, L. I., N. Y., consists of two units: the Model 1603-AR-1 generator unit, and the Model 1603-AR-2 receiving unit.

The function of the Model 1603-AR Envelope Delay Curve Tracer is to provide a determination of the envelope delay characteristic as well as the amplitude characteristic of any amplifier or system, using conventional sweep techniques. By using two oscilloscopes in conjunction with a video sweep generator, it is possible to plot both the amplitude and envelope phase characteristic simultaneously. Calibrated meters are provided to give the envelope delay on a point by point frequency basis. To insure accuracy in reading, indicating meters of the 270° type are used. The measurements can be made to an accuracy of ±0.01 microsecond in absolute value, and to even greater accuracy in terms of relative values.

The equipment is so designed that the transmitter unit may be located at the beginning of a system and the receiver unit at the end. For looped operation, no no auxiliary equipment is necessary.

Drafting Symbol Stamp

John Griffin Co., 2157 James Ave., St. Paul 5, Minn., is producing symbols and circuit stamps. Drafting time should be reduced markedly because it only takes a few seconds with these stamps to print a symbol that takes minutes to draw.

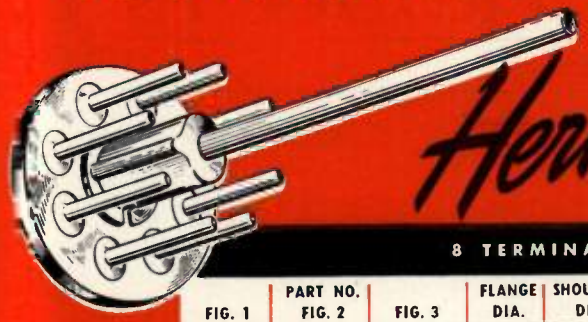
The stamp is mounted on a clear plastic base so that it is possible to see the area in which the draftsman is printing the symbol.

The stamps come in kits of 27, for $27.50 fob St. Paul. A brochure is available of the standard stamps on hand and buyer makes his own selection. Separate stamps are supplied at $1.75 each.

Stamps of the customer's own design, will be supplied in lots of 10 at $3.00 each.

(Continued on page 58A)

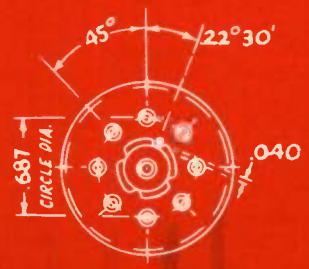
VAC-TITE* COMPRESSION SEALS



Hermetic produces the greatest variety of

8 TERMINAL

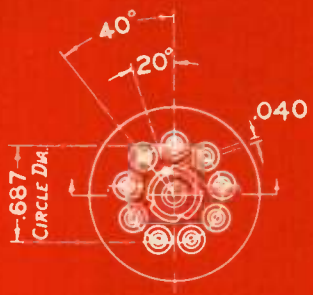
FIG. 1	PART NO. FIG. 2	FIG. 3	FLANGE DIA.	SHOULDER DIA.	A	B	EXHAUST TUBULATION
A7079-1	A7042-1		1.250	1.175	1/8	040	As Shown
A7079-2	A7042-2		1.250	1.175	1/8	040	None
A7045-1	A7137-1		1.250	998	1/8	080	As Shown
A7045-2	A7137-2		1.250	998	1/8	080	None
A7138-1	A7139-1		1.235	1.062	1/8	045	As Shown
A7138-2	A7139-2		1.235	1.062	1/8	045	None
A7108-1	A7140-1		1.125	998	1/8	080	As Shown
A7108-2	A7140-2		1.125	998	1/8	080	None
A7141-1	A7142-1		1.062	998	1/8	080	As Shown
A7141-2	A7142-2		1.062	998	1/8	080	None
		A7041-1	—	1.062	1/8	—	As Shown
		A7041-2	—	1.062	1/8	—	None
		G7143-1	—	998	1/8	—	As Shown
		G7143-2	—	998	1/8	—	None



Bayonet Type

9 TERMINAL

FIG. 1	PART NO. FIG. 2	FIG. 3	FLANGE DIA.	SHOULDER DIA.	A	B	EXHAUST TUBULATION
A7144-1	A7058-1		1.250	1.175	1/8	040	As Shown
A7144-2	A7058-2		1.250	1.175	1/8	040	None
A7038-1	A7145-1		1.250	998	1/8	080	As Shown
A7038-2	A7145-2		1.250	998	1/8	080	None
A7146-1	A7147-1		1.235	1.062	1/8	080	As Shown
A7146-2	A7147-2		1.235	1.062	1/8	080	None
A7148-1	A7149-1		1.125	998	1/8	080	As Shown
A7148-2	A7149-2		1.125	998	1/8	080	None
A7150-1	A7151-1		1.062	998	1/8	080	As Shown
A7150-2	A7151-2		1.062	998	1/8	080	None
		A7047-1	—	1.062	1/8	—	As Shown
		A7047-2	—	1.062	1/8	—	None
		G7152-1	—	998	1/8	—	As Shown
		G7152-2	—	998	1/8	—	None

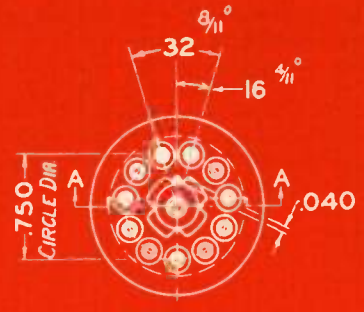


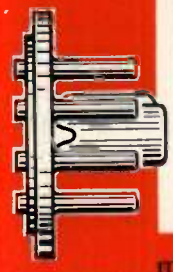
Polarizing Plugs

11 TERMINAL

FIG. 1	PART NO. FIG. 2	FIG. 3	FLANGE DIA.	SHOULDER DIA.	A	B	EXHAUST TUBULATION
A7085-1	A7059-1		1.250	1.175	1/8	040	As Shown
A7085-2	A7059-2		1.250	1.175	1/8	040	None
A7046-1	A7084-1		1.250	998	1/8	080	As Shown
A7046-2	A7084-2		1.250	998	1/8	080	None
A7153-1	A7154-1		1.235	1.062	1/8	045	As Shown
A7153-2	A7154-2		1.235	1.062	1/8	045	None
A7155-1	A7156-1		1.125	998	1/8	080	As Shown
A7155-2	A7156-2		1.125	998	1/8	080	None
A7157-1	A7158-1		1.062	998	1/8	080	As Shown
A7157-2	A7158-2		1.062	998	1/8	080	None
		A7048-1	—	1.062	1/8	—	As Shown
		A7048-2	—	1.062	1/8	—	None

Available also with solid pins





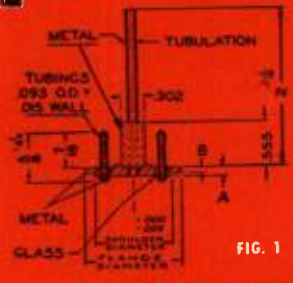


FIG. 1

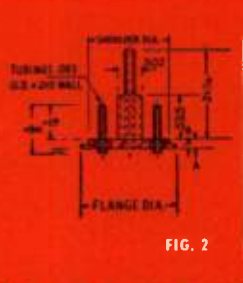


FIG. 2

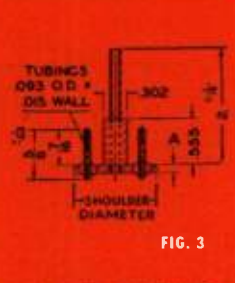


FIG. 3

*VAC-TITE is HERMETIC's new vacuum proof compression-construction glass to metal seal. In addition to special shapes, many standard sizes such as .300 O.D. and .900 O.D. multi-terminal headers and a large variety of individual terminals are available in VAC-TITE Compression Seals.

Write for samples, data, prices.
We welcome the opportunity to work with you on special problems.

HERMETIC SEAL PRODUCTS CO.
29 South Sixth St.
Newark 7, N. J.

FIRST AND FOREMOST IN MINIATURIZATION

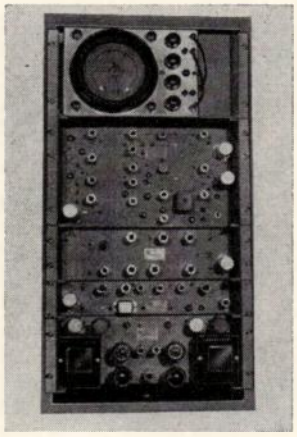
NEWS and NEW PRODUCTS

February 1954

(Continued from page 56A)

Signal Certification and Monitoring Equipment

Telechrome, Inc., 88 Merrick Rd., Amityville, L. I., N. Y., has designed the Model 1601-AR Chromascope (NTSC Signal Certification and Monitoring Equipment) used for accurate checking of the performance of NTSC color signals. The use of this instrument facilitates alignment of encorders, transmitters, and checking of phase errors in color signals. Operating from a composite NTSC signal, the presentation is in the form of an instantaneous vector plot of the colors and their amplitudes. The display is on a 7 inch cathode-ray tube which has a transparent overlay calibrated in degrees and amplitude.



The equipment is self-standardizing and self-checking by means of a built-in quadrature checker with calibration points phase spaced 90° apart. In use, the equipment is continuously self-checked by the time sharing of signals: the color signals are presented for 50 per cent of the time; the self-checking quadrature signals are presented for 25 per cent of the time, with the phase calibrating marker signals during the same interval; 25 per cent of the time no signals are carried; giving a zero reference point. The duty cycles of the various signals may be adjusted at will.

Included in the 1601-AR are: suitable decoding amplifiers, quadrature switching circuits, phase delay circuits, local subcarrier amplifiers, a color subcarrier generator, the 90° marker generator, a reference oscillator, and a regulated power supply.

Supplied as standard equipment is a 7 inch oscilloscope with a calibrated chart on the face of the cathode-ray tube for setting standard values of the major saturated colors: red, green, blue, yellow, cyan, and magenta. The burst amplitude and phase are likewise indicated on this chart.

The equipment is designed for standard relay rack mounting and occupies approximately 40 inches of rack height exclusive of the 7 inch oscilloscope.

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

Community TV Cables

U. S. Wire & Cable Corp., Progress & Monroe Sts., Union, N. J., has developed a line of Co-axial cables which are constructed for use in both double-braid—single jacket, and double-braid—double-jacketed types. These units also fulfill general high frequency needs, as required for joint Army-Navy uses, and correspond to the RG 11/U, and RG 59/U, JAN types.

These new type cables are also available in single braid, air capacitance types, as well as armored types of cables.

For engineering data and prices write to the manufacturer.

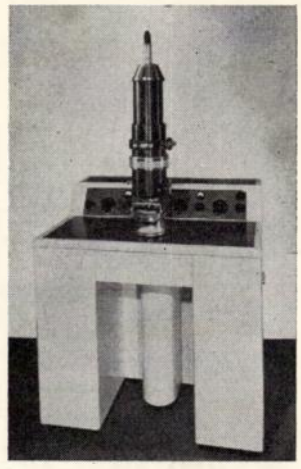
Computer Memory Unit

Computer Control Co., 106 Concord Ave., Belmont 78, Mass., announces the development of a memory unit, Model 3C1-384, composed of a solid acoustic delay line and associated circuitry.

The unit stores 384 bits at a pulse repetition rate of 1 mc. A self-contained heating element and thermal control gives temperature stability to the quartz line. Inherent accuracy of delay control exceeds design requirements. All germanium diodes are grouped in a single plug-in type subunit. All five tubes are 6AN5's operated 80 per cent below manufacturer design center. The over-all dimensions of the plug-in chassis are $4\frac{1}{2} \times 5\frac{1}{2} \times 10$ inches. A self-contained filament transformer is optional. Input voltage requirement into write-erase gate is 10 volts. The reshaped output signal level is 15 volts into a 100 ohm impedance load. Carrier frequency is 20 mc. All circuits are degenerated, with reserve gain in the wide band IF amplifier. There is a gain control for initial adjustment of the IF stage. No tuning is necessary. Temperature coefficient of the quartz delay line is 123 parts per million per degree centigrade. The unit is fitted for airborne use and is insensitive to shock.

Small Electron Microscope

A new Electron Microscope (EM-75) is now available from the **Research & Control Instruments Div., North American Philips Co., Inc.,** 750 S. Fulton Ave., Mt. Vernon, N. Y. Its cost is about half that of the large Philips unit (EM-100), and it delivers approximately half the power.



Suitable for analytical work in industrial processing and research, the new instrument utilizes an electron-optical system that includes capacitor, objective and projector lenses.

Magnification is continuously adjustable between 1500 and 15,000 diameters and are variable between 10 and 75 kv. Resolving power is better than 100 Angstroms under average operating conditions.

The new unit has a magnetic stabilizer for high voltage and lens currents. The electron gun is self-biased, is equipped with an emission meter and the V-shaped filaments are easily replaced.

The electro-magnetic deflection method is used for beam alignment and focusing is facilitated by means of a beam wobbler. Specimen carriers for $\frac{1}{8}$ inch diameter grids and plates are available and a specimen airlock is utilized to obtain quick working cycles.

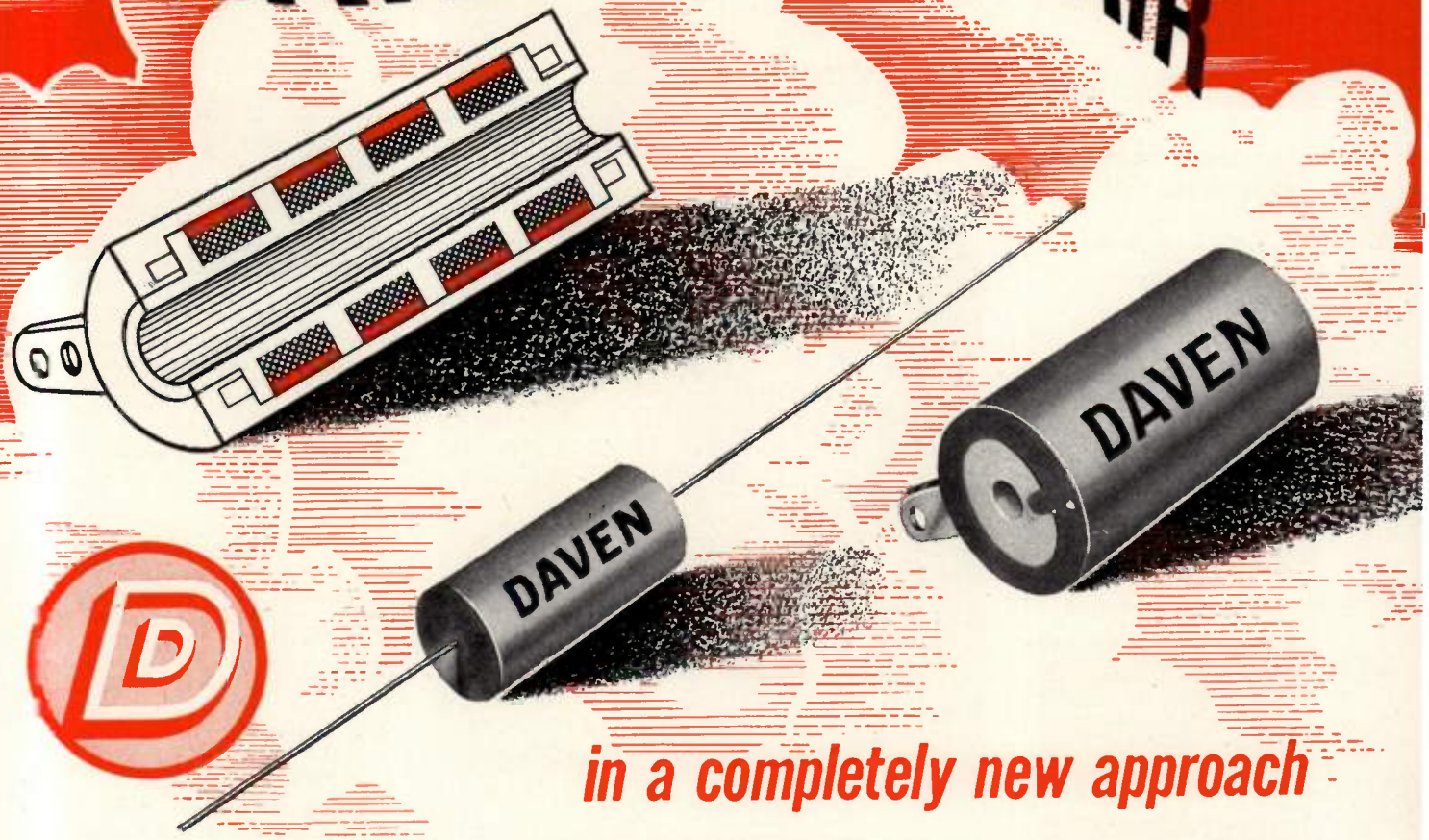
Stereoscopic exposures can be made by tilting the specimen carrier 15 degrees, without breaking the vacuum. Diameter of the fluorescent screen is $4\frac{1}{8}$ inch, diameter of images $3\frac{1}{8}$ inch and a supplemental 2× optical magnifier is available that fits into place on the column.

A camera using 35-mm standard film is employed on the microscope and each loading permits the making of rapid sequential exposures without breaking the vacuum. The camera is completely controlled from the outside of the column and a special device prevents double exposure. An exposure register is built in.

The EM-75 electron microscope is approximately 74 inches high, 37 inches wide, and 25 inches deep. Power supply is 220 volt, 50 or 60 cps and power consumption is about 1 kw.

(Continued on page 60A)

DAVEN TURNS TO AIR



in a completely new approach

to the production of Encapsulated Seald-Ohm Wire Wound Resistors

Only DAVEN turns to air to keep the molding material absolutely separated from the resistance wire in its new line of Super Davohm Encapsulated Seald-Ohm Resistors. The wire is maintained in a slot filled with dry air . . . no external pressures are applied to it. These air pockets, between the wire and the plastic coating, guarantee absolute stability . . . eliminate shorted turns.

Only DAVEN matches the temperature coefficient of expansion of the molding compound with the ceramic bobbin, the resistance wire and the metal terminals. This removes the possibility of cracks or strains on the wire during cycling.

Because of the special construction used, Daven can furnish Encapsulated Wire Wound Resistors with temperature coefficients below ±20 P.PM/°C. when required, and with accuracies to ±.05%.

These exclusive Daven precision, wire wound resistors are completely hermetically sealed . . . yet are no larger than standard lug-type resistors.

In addition, these units are made in accordance with MIL-R-93A specifications, and are substantially more rugged than conventional resistors. They will withstand the JAN-R-93, characteristic A, salt-water immersion test, and, in addition, temperature cycling from −65°C. to +125°C. The strong molding material will resist pressures equivalent to 75,000 ft. altitude, and will not cold flow at temperatures up to 150°C.

Write for latest Resistor Brochure

THE DAVEN CO. 195 Central Ave. Newark 4, N. J.

WORLD'S LARGEST MANUFACTURERS OF ATTENUATORS

Designed for *high-speed pulse* observation

...The Tektronix TYPE 517

EXCELLENT TRANSIENT RESPONSE
Risetime—
0.007 μsec or less.

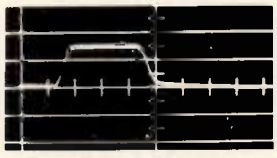
FAST SWEEPS
0.01 μsec/cm to
20 μsec/cm.

HIGH ACCELERATING POTENTIAL
24 KV on
metallized crt.

SENSITIVITY—0.1 v/cm

Type 517 Oscilloscope—$3500
f.o.b. Portland (Beaverton), Oregon

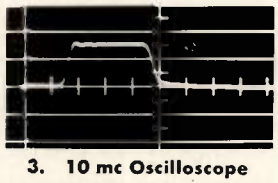
0.3 μsec pulse with cable properly terminated



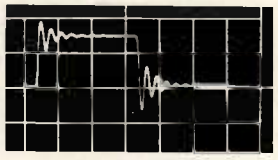
1. 10 mc Oscilloscope

2. Type 517

0.3 μsec pulse with cable improperly terminated



3. 10 mc Oscilloscope



4. Type 517

The waveform photographs at left illustrate the capabilities of the Type 517 in one application dealing with extremely fast-rising pulses. The same 0.3 μsec pulse is shown in all four photographs—first with a three-foot length of 93-ohm coaxial cable properly terminated (photos 1 and 2), and below with improper output termination (photos 3 and 4). Notice in photo 4 how clearly the Type 517 shows the effects of improper termination on the pulse.

If your project is one that requires an oscilloscope with the faithful transient response, fast sweeps, and high accelerating potential of the Type 517, call your Tektronix Field Engineer or Representative for complete specifications—or write to:

Tektronix Inc.
P. O. Box 831B • Portland 7, Oregon • Cable: TEKTRONIX

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 58A)

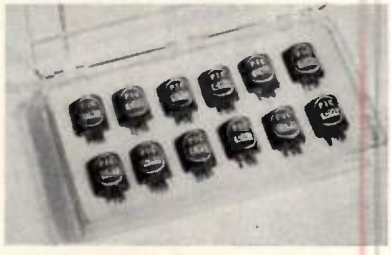
SWR Amplifier

A new instrument is announced by **Browning Laboratories, Inc.**, Winchester, Mass., for the purpose of determining standing wave ratios or the comparison of any two demodulated signals.

The TAA-16B features dual input channels with gain sufficient for full scale meter deflection with less than 2 microvolts input. The amplifier may be used broadband from 500 to 5,000 cps or may be sharply tuned over this range by means of panel controls. For use with bolometers, an internal voltage source is supplied and bolometer current for either input is metered and adjustable from the front panel. A crystal protective circuit is incorporated to minimize possibility of burnout due to accidental application of bolometer voltage. The meter scale is calibrated in swr and db of swr with a precision attenuator permitting db of swr readings up to 50 db. Jacks are provided for remote metering and for dc recorder where a permanent record is desired.

Pulse Transformer Kit

Polyphase Instrument Co., Bryn Mawr, Pa., has a plug-in pulse transformer kit for circuit designers working with pulse circuits.



The circuit designer may wire a 7-pin miniature tube socket into his circuit in place of the pulse transformer and in a few seconds can plug in a wide variety of pulse transformers until the circuit performance required is obtained.

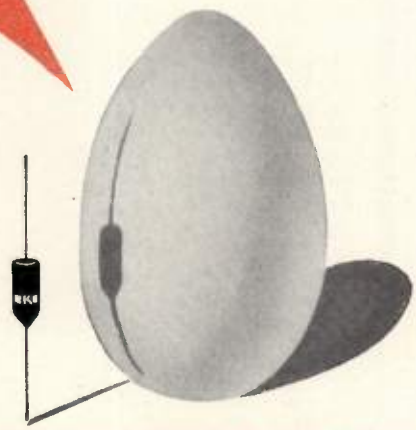
The kit contains a selection of 12 standard plug-in type pulse transformers, each having different impedances or turns ratios. The transformers are ¾ of an inch in diameter and ¾ of an inch high; all units are hermetically sealed and designed to pass MIL-T-27 specifications with pulse widths as low as 0.1 μsec, for impedance matching, phase inversion, or as wide band-pass transformers. They also find use in linear oscillator circuits in the frequency range from 1 kc to 20 mc. Delivery is from stock. The kit price is $97.50.

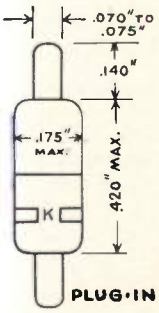
(Continued on page 62A)

NEW RAYTHEON hermetically sealed GERMANIUM DIODES

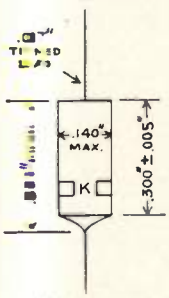
a new sealed package worth crowing about

- Completely sealed against atmospheric conditions
- 100% testing of all diodes
- *Every* diode temperature-humidity cycled for 32 hours; also held for 12 hours at 105°C, 4 hours at −25°C
- Samples of every production lot put through JAN-193 humidity test
- Oscilloscope tests for hysteresis and the various types of instability
- *Smaller size* for solder-in design
- NEW plug-in design
- Extra rugged and shock resistant
- Coated with protective insulating lacquer





PLUG-IN



SOLDER-IN

TYPE	TYPICAL APPLICATION	MAX. DC INVERSE VOLTAGE	MAX. PEAK ANODE CURR. ma.	MAX. AVG DC ANODE CURR. ma.	MIN. FORWARD CURRENT AT +1V ma.	MAXIMUM INVERSE CURRENT AT −5V ma.	AT −10V ma	AT −50V ma.	AT −100V ma.	MIN. INVERSE VOLTAGE	AVG INVERSE CURRENT −50V
1N67 1N67-P	50V DC Restorer	80	100	35	4.0	0.005		0.05		100	0.1
CK705 CK705-P	Gen. Purpose Diode	60	150	50	5.0		0.05	0.8		70	0.43
CK705A CK705A-P	Gen. Purpose Diode	60	150	50	5.0		0.01	0.8		70	0.43
CK707 CK707-P	50V DC Restorer	80	100	35	3.5	0.010		0.10		100	0.1
CK708 CK708-P	100V DC Restorer	100	100	35	3.0				0.625	120	0.15
CK713 CK713-P	Computer Diode	65	150	50	30 at +2V			0.25 at −40V	(DC characteristics at 50°C)		

P Indicates plug-in type DC Characteristics measured at 25°C

These types are available in production quantities through Newton, Chicago and Los Angeles sales offices. They are also stocked by over 500 Raytheon Special Tube Distributors.

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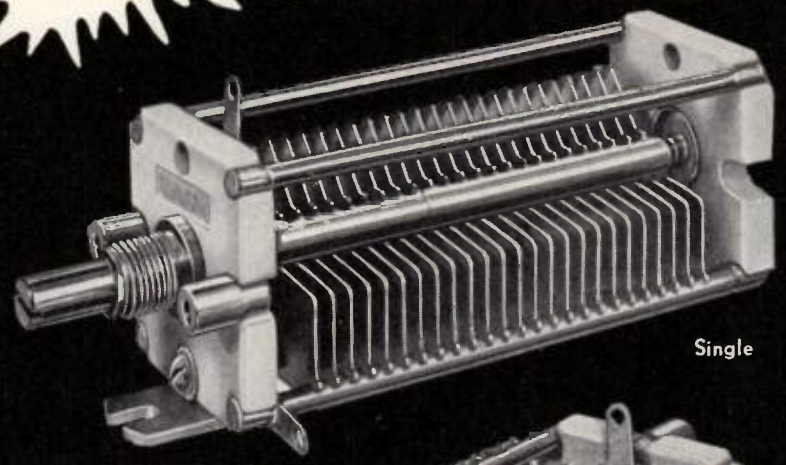
Want a rugged high frequency capacitor able to withstand extreme vibration?

CERAMIC SOLDERED*

SPECIFY JOHNSON TYPE "L" CAPACITORS



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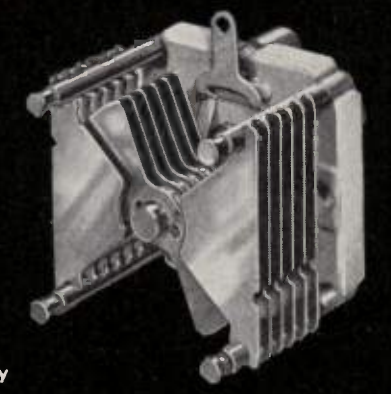
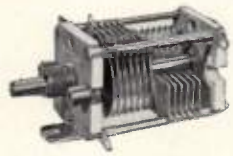


Butterfly

CERAMIC SOLDERED*

An important forward step in variable capacitor design. Bonding points on steatite end frames are subjected to a special metalizing process; then stator rods and rotor bearing are soldered (under carefully controlled heat) directly to the steatite. Resulting bond is permanent and vibration proof.



Dual

Differential

Type L variables also available with .020, .060, and .080 spacing for special applications in production quantities. Special plating, mounting brackets, shaft lengths, terminals also furnished to specifications.

Full soldered construction and rugged design make JOHNSON type L high frequency capacitors the perfect choice for any electronic application requiring extreme stability and rigidity. Rotor bearing, stator rods, and structural tie rods are soldered direct to 3/16" steatite end frames. Tests prove resulting ceramic to metal bond is stronger than the rugged end frames themselves . . . won't loosen or break loose in roughest vibration . . . rotor and stator alignment is permanent, capacity wobble is eliminated.

OTHER DESIGN FEATURES

Extra thick (.020) brass plates and large diameter (⅛") stator rods for extremely low inertia mass . . . special split sleeve tension bearing and silver plated beryllium copper wiring contact for constant torque and smooth silent capacity variation even at highest frequencies. Standard spacing .030 rated at 1,500 volts peak DC breakdown; over 500 volts peak at 50,000 feet altitude. Panel space required 1⅜" square. Mounting posts tapped 6-32 on 1-3/32" centers. Shaft is ¼" with screwdriver slot. Standard plating corrosion-resistant bright alloy.

Write today for quotations or additional data

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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 60A)

UHF Wide Band Power Oscillator

A uhf wide band power oscillator useful for precision test and power measurement has been developed by **The W. L. Maxson Corp.**, 460 W. 34 St., New York 1, N. Y. Designed for use in antenna radiation, antenna field strength, wave filter, noise, and interference measurements, the instrument is adaptable as a general purpose, low-power portable transmitter and in production testing and measurement of uhf equipment.

It consists of a grid-separation, dual cavity, coaxial line oscillator, a 2C39A disc-seal triode, and an integral rectifier power supply. Plate and grid circuit line lengths are simultaneously set by a single tuning control, with provision made for individual tuning of the grid-cathode line for maximum power output. A panel-mounted meter indicates optimum tuning of the oscillator. For load matching the output coupling from the cavity to the output line is adjustable.

The frequency range is 200 mc to 2,500 mc with one band change-over by minor adjustment of feedback assemblies and parts. Calibration accuracy of the greater of ±1 per cent or ±5 mc is claimed.

The power output varies with the frequency: 50 w at 200 to 400 mc, 25 w at 400 to 1,000 mc, 10 w at 1,000 to 2,500 mc. Using type LN connector and RG 8/U or RG/9U cable it has a 50 ohms output impedance.

The power consumption of the uhf oscillator is 200 w at 115 v, 60 cps. External plate modulation of either audio or video frequencies is provided by a type N, coaxial input jack with matching impedances. It will operate with amplifier output impedances of 4,000 ohms or less with a minimum modulation power requirement of approximately 5 w. Cabinet is 24 inches long, 12 inches high and 16 inches deep, and weighs approximately sixty pounds.

Filter Pumps Bulletin

A new eight-page bulletin issued by **Sethco Mfg. Co.**, 70–78 Willoughby St., Brooklyn 1, N. Y., describes filter pumps and equipment. The booklet goes into the design and application of corrosion resistant filter pumps, and discusses solutions that can be filtered and the result of its operation. Photographs show the filter pump models and motor types. Price is also included.

(Continued on page 119A)

PROCESSING MACHINES

LET SAFEGUARDS GO ON WITH THIS LONG LIFE

GUARDIAN®

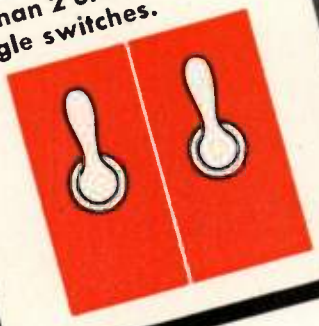
8 P. D. T. Switch

Design FOR SAFETY

Hit the button or throw a lever on the Guardian 8 P. D. T. Switch and this overriding master unit instantly switches or disconnects 8 primary circuits. Constant contact pressure (4 ounces per point) is maintained throughout the stroke until positive *snap-action* occurs in either direction. No dead center. Contacts handle 5 amps, 28 V. D.C. inductive—10 Amps, 115 V. A.C. non-inductive—60 to 400 cycles. Unit mounts on 1½" x 1½" panel area to conserve space and decrease overall costs. Sand and dust type enclosure easily removed for inspection. Designers of processing machines seeking to improve multiple control of any machine or a battery of machines can increase the efficiency and safety factors with the Guardian 8 Pole Double Throw Switch. Built to exceed military specification ANS-63 applicable to this type switch to resist shock and vibration and to operate at −65° F. to +160° F. Easy to wire—easy to mount—compact—low cost. Available with auxiliary "push-to-test" S. P. D. T. circuit and dial light.

Design FOR EFFICIENCY

Save costs as well as space. This 8 pole double throw switch requires no more panel space than 2 ordinary toggle switches.

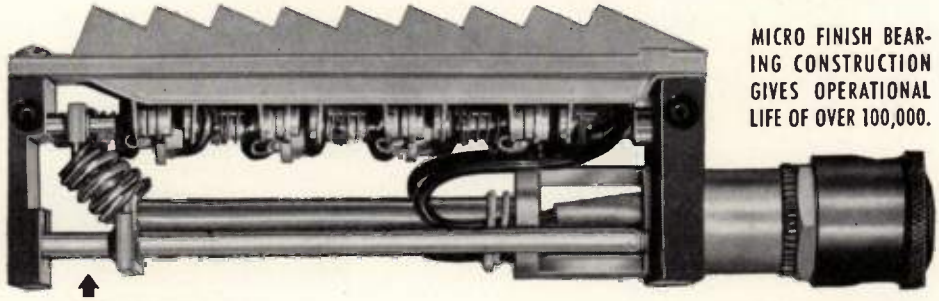


Weight: approx. 8 oz.

8 Pole Double Throw

Designed for Easy Multiple Wiring to Each Terminal.

Stop carriers—release machine loads—stop machines—reverse processing—resume operations. For efficient control of multiple machine operations use the Guardian 8 P.D.T. Switch. Shuts off boiler feeds—dumps fuel—starts extinguishers.



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New RCA Tubes for Economical TV Receiver Designs

RCA-6AN8
General-Purpose Triode-Pentode

RCA-6AN8 is a general-purpose multi-unit tube combining a medium-mu triode and a sharp-cutoff pentode in one envelope. It is designed for versatility in many TV receiver applications. A special basing arrangement, together with internal shielding, virtually eliminates coupling between units of the tube. All of these features contribute to economical design of TV receivers.

RCA-6AS8
Video IF Amplifier and Detector

RCA-6AS8 features two units—a high-perveance diode and a sharp-cutoff pentode—in a single envelope. Designed for use in the last picture if amplifier stage and video detector stage, the 6AS8 offers a pentode transconductance of 6200 μmhos and a basing arrangement which provide effective isolation of the diode and pentode units. These features facilitate economy in TV receiver design.

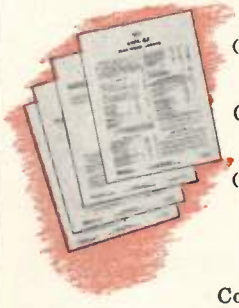
RCA-6BY6
Pentagrid Amplifier

RCA-6BY6 is a pentagrid amplifier designed for gated amplifier service in TV sync-separator circuits. An important operating feature of 6BY6 design is its favorable ratio of plate current to grids No. 2 and No. 4 current—which permits achieving the desired output signal with relatively low power input to grids No. 2 and No. 4.

RCA-6AU4-GT
Damper Diode

RCA-6AU4-GT is half-wave rectifier of the glass-octal type. It is especially useful as a damper diode in TV receiver circuits. This tube can withstand a maximum peak inverse plate voltage of 4500 volts (absolute), and can supply a maximum peak plate current of 175 ma. The RCA-6AU4-GT can handle negative peak pulses up to 4500 volts between heater and cathode (with a dc component as high as 900 volts).

For technical data on these important TV tubes—or for help on specific design problems—write RCA, Commercial Engineering, Section B50Q, Harrison, N. J. Or call your nearest RCA Field Office:



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Complete technical bulletins on RCA-6AN8, 6AS8, 6BY6, and 6AU4-GT are yours for the asking—from RCA, Commercial Engineering, Section B35Q, Harrison, N. J.

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For Peak Performance

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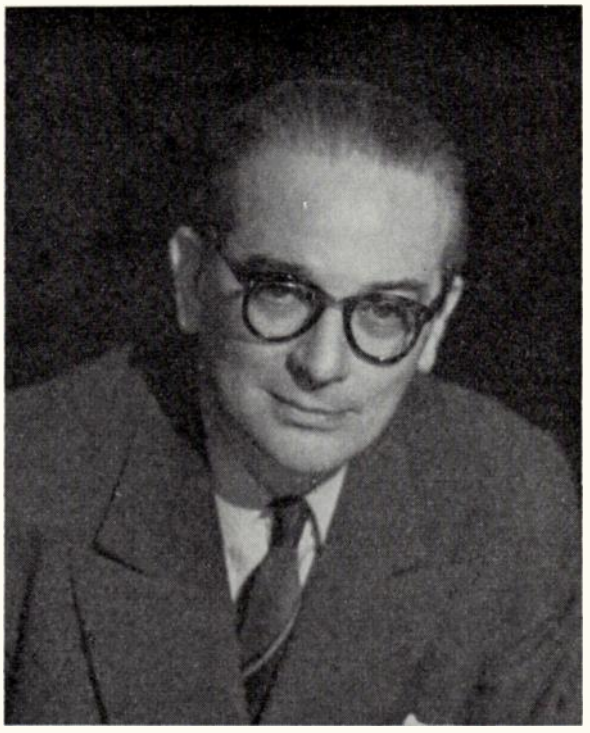

Maurice J. H. Ponte

VICE-PRESIDENT, 1954

Maurice J. H. Ponte was born on April 5, 1902 in Voiron, France. He was graduated from the Ecole Normale Supérieure of Paris, in the Sciences Division. He then took a Sciences Doctorship, with a thesis on wave mechanics. He also holds a University Professorship in Physical Sciences.

His early research works were made under the direction of H. Abraham and E. Bloch, and were devoted to thermionic emission. After a period of research work on X-ray emission from gaseous substances and alkaline compounds with the laboratories of Maurice de Broglie, he received in 1926 a Rockefeller Fellowship for working in the British laboratory of Sir William Bragg, where he continued his studies on X-ray diffraction. In 1928 he returned to France, and began work on method and experimental equipment devised to check the wave mechanics law of Louis de Broglie. In 1929 Dr. Ponte joined a French industrial organization, and was later put in charge of the Research Laboratory which became the Compagnie Generale de Telegraphie Sans Fil Research Centre. Later Dr. Ponte was placed in charge of the Technical Management of this organization, and in 1951 was appointed General Manager. In 1952 he took charge of the general management of the affiliated company, the Société Français Radioélectrique, in which position he developed several high-power transmitting and microwave tubes.

Dr. Ponte has delivered several lectures both in France and abroad, and has published many important papers in various scientific journals. He has filed several patents, of which the main ones are relative to microwave detection unit and radio-navigational devices.

Dr. Ponte became a Senior Member of the Institute in 1948, and in 1954 was awarded the Fellow Grade "for his contributions to high-powered electron tubes, and his unswerving leadership in electronic research in France." He is Vice-President of the French National Committee, U.R.S.I., and a member of the Scientific Council, French Atomic Energy Commissioner Office. He is also a member of various other scientific groups in France, Great Britain, and the United States.

An Agenda for Communication Engineers

ARTHUR A. COLLINS

In our search for new technical worlds to conquer, we are apt to overlook the relatively long-established and well-explored areas of our engineering domain. And yet, it is in just such areas that some of our most pressing problems and promising opportunities may lie.

How this situation applies to the radio communication field and what problems are in urgent need of solution are skillfully described in the following guest editorial written by a Fellow of the IRE who is President of the Collins Radio Co., Cedar Rapids, Iowa.—*The Managing Editor*

It has been a long time since communications has occupied the spotlight in the radio engineering field. In more recent years interest has been focussed instead on such new and dramatic developments as radar, television, transistors, and automatic control and computers. Meanwhile, we have given less attention to communications. Perhaps the feeling has been that, as the oldest branch of the art, communications has reached such a high state of development that there no longer remain any problems to attract the attention of the now vast electronics industry, other than a gradual refinement of details and elaboration of existing systems.

In reality, we are now faced with several challenging problems that are in urgent need of solution, an urgency born of the fact that our modern society is going to require an expansion of communication facilities at a rate which is even now far outstripping previous forecasts.

Already we are experiencing the expanding use of communication channels for utilities, pipelines, conservation services, transportation operations, and the like. With the present rapid decentralization of industry and the resulting far-flung operations of all sorts, there will soon be additional demands for large quantities of direct communication channels so that inventory and production control center, and eventually production machines and processes themselves, can be interconnected, no matter how far separated.

Accompanying the expansion problem is the need for improved system performance. World War II experience with military equipment revealed case after case of communication failure at crucial times. In the main, these failures were due to inherent difficulties which are still with us, namely, the vagaries of propagation, the extreme weakness of signals, and the necessary complexity of communication systems.

With the need for expanding and improving our means of radio communication what then are the most promising lines of attack? High on the agenda would be the following:

First, there is the urgent matter of increasing the available number of channels. We are making very inefficient use of the portions of the spectrum known to be suitable for long range circuits. We have in our laboratories reliable means of controlling frequencies to one part in 10^8; yet the usual spacing of communication channels is today based on operating stabilities of no better than one part in 10^4, indicating a tremendous waste of precious spectrum space. Also, partially as a result of our loose frequency stability practices, we still use unnecessarily wide modulation and keying bandwidths. Straightforward engineering solutions are capable of correcting these deficiencies and of effecting a several-fold increase in the traffic capacity of the most crowded portions of the spectrum.

Second, it is important for us to acquire a better and more widespread understanding of the principles of detecting signals in the presence of noise in order that we can improve the reliability and extend the periods of usefulness of available frequency channels.

Third, we can look for new portions of the spectrum which can be made to serve our purposes. Here is a field for the iconoclast, for history is replete with instances where theoretical limitations imposed by existing concepts of wave propagation have been shattered by the experiments of radio amateurs and professionals alike.

Finally, as soon as we learn how to make our radio signals more positive and reliable so that we do not have to depend so heavily upon the human ear and brain to decipher them, we can plan our radio communication systems to use many land-line practices such as selective calling, automatic signalling, automatic exchange switching, and automatic traffic loading of trunk channels, together with extensive interconnection between systems. Such measures will extend greatly the convenience, flexibility, and capacity of radio communication.

Here, then, indeed is a challenging agenda worthy of the immediate attention of the radio engineering fraternity.

A *P-N-P* Triode Alloy-Junction Transistor for Radio-Frequency Amplification*

C. W. MUELLER†, SENIOR MEMBER, IRE, AND J. I. PANKOVE†, ASSOCIATE, IRE

Summary—**The performance of alloy-junction transistors falls off as the frequency increases. Heretofore, the chief cause of this has been the resistance-capacitance low-pass filter effect in the base-emitter input. The effect is produced by the germanium resistance (between the external base connection and the active junction region) and the emitter-to-base diffusion capacitance. The latter may be extraordinarily high because of the relatively slow diffusion of charge carriers into the base region which must be charged up and discharged with minority carriers. In the new transistor design of the present paper, resistance and capacitance have been reduced by using a thick wafer of low-resistance germanium and placing the active junctions on a very thin section produced by drilling a well into the wafer. The junctions are about 0.0005 inch apart. Other capacitances are reduced by using small diameter junctions. To prevent limitation due to transit-time dispersion, nearly planar junctions are obtained by using germanium-indium alloy discs from which the junctions are formed.**

The new unit will give 39-db gain (neutralized) at 455 kc, 12 db (unneutralized) at 10 mc and has an oscillation limit of 75 mc. A noise factor at 1 mc as low as 4 db is achieved.

Introduction

IN NORMAL broadcast receiver design, an amplifier giving good gain in the frequency region of ½ to 3 mc is desirable. In order to fulfill this need, keep power consumption to a minimum, and retain the simplicity and versatility of the triode alloy-junction transistor, a development was undertaken to extend the frequency range of this type of transistor.

To follow the course of the present development, it is desirable to discuss first the important factors which limit the frequency response of the triode transistor such as the TA-153.[1] The important factors stem from a combination of two effects; one inherent to transistor operation and the other extrinsic to transistor action but arising from the constructional details. In the operation of this type of transistor it is well known that minority charge carriers, after injection by the emitter, flow to the collector by a process of diffusion through the base region and receive little or no aid from electric fields. Such diffusion currents flow under the influence of a density gradient (i.e. different carrier charge densities in different regions of the base layer). Thus, when a signal is applied to the transistor, the number of carriers in the base region must vary in accordance with the signal, so that the diffusion current flow to the collector will reproduce the signal in the output circuit. This change in the charge of the base region induced by the applied signal acts as an emitter-to-base capacitance. Furthermore, because this charge flow in and out of the base region is a diffusion flow, an extraordinarily high capacitance results. (Of the order of 0.01 mf in the TA-153 transistor.[1])

This capacitance is much greater than that associated with the transition region of the emitter junctions. The "diffusion" capacitance is given approximately by

$$C_{b'e} = \frac{q}{kT}\frac{W^2 I_E}{2D}$$

where $C_{b'e}$ is the input capacitance element of the common emitter representation of Fig. 1, W is the effective thickness of the base region, I_E is the dc emitter current, D is the diffusion constant for minority carriers in the base region, q is the electronic charge, k is Boltzmann's constant and T is the absolute temperature.

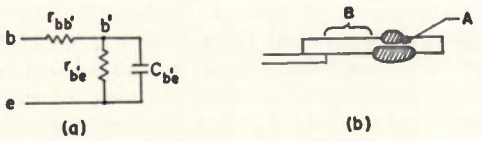

Fig. 1—(a) Equivalent input circuit of the transistor.
(b) Transistor cross section showing the physical location of the parts of $r_{bb'}$ on a TA-153 audio transistor.

When I_E is expressed in milliamperes, and W is in mils, this formula gives the capacitance in farads as

$$C_{b'e} = \begin{cases} 2.8 \times 10^{-9}\, W^2 I_E & \text{(for } p\text{-}n\text{-}p \text{ units)} \\ 1.3 \times 10^{-9}\, W^2 I_E & \text{(for } n\text{-}p\text{-}n \text{ units).} \end{cases}$$

Since this capacitance is proportional to the square of the thickness of the base region, it is important to minimize this dimension in transistors designed for high frequency applications. Note that this capacitance is proportional to the dc current and does not involve the junction area. It is evident that a measurement of $C_{b'e}$ will provide a measure of the effective thickness of the base region.

Now, in the construction of the alloyed-junction transistor, the external base connection to the germanium wafer is made at some distance from the active junction region. The resistance of the germanium between the base connection and the active junction constitutes a series base lead resistance ($r_{bb'}$ of Fig. 1) through which the input signal must pass, with consequent attenuation before it can be applied to the intrinsic transistor. Furthermore, the charging current for the emitter-to-base capacitance discussed above must also flow through this resistance. This combination of series resistance and emitter-to-base capacitance constitutes a low-pass filter which is one of the most impor-

* Decimal classification: R282.12. Original manuscript received by the Institute, August 12, 1953. This paper was discussed at the IRE-AIEE Transistor Research Conference at State College, Pennsylvania, July, 1953.

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

[1] R. R. Law, C. W. Mueller, J. I. Pankove, and L. D. Armstrong, "A developmental germanium *p-n-p* junction transistor," PROC. I.R.E., vol. 40, pp. 1352–1357; 1952.

tant limiting factors of transistors such as those previously described.

The presence of this base lead resistance in addition to diminishing the input signal available to the actual transistor junctions is also a feedback element and its presence in the transistor leads to a reduction in the overall output resistance and an increase in the overall output capacitance of the transistor.

In addition to the input circuit effects discussed above, there are also further consequences resulting from the fact that charge carriers flow from emitter to collector by a diffusion process. Diffusion flow is essentially a random process with the individual carrier making many collisions and changing direction of motion many times during its travel across the base region. Directivity to this random motion is supplied by the concentration gradient discussed above. Thus, the transit time for the individual carriers must vary widely, giving rise to a dispersive effect in the transfer function (i.e. transconductance). This transit time dispersive effect in the transfer of carriers to the output circuit also limits the frequency response of transistors but, for transistors such as the TA-153, its effect is not as detrimental as the input circuit effects discussed above.

The above transit-time dispersion effects in the transfer function have been discussed as resulting only from diffusion phenomena. There may also be an additional effect of similar nature (i.e. transit time dispersion) due to different path lengths caused by non-planar geometries and end effects. This difference in path length may be minimized by small emitter-to-collector spacings, planar junctions and accurate centering.

The picture given above of the consequences of diffusion flow in the base region has tended to separate the effects on the input circuit from those concerned with the transfer generator. This leads quite naturally to the π equivalent-circuit representation used in this paper (Fig. 4). It may be well to point out that, in other representations, such as an equivalent T circuit, the transfer generator (alpha) involves both the transfer transit-time dispersion effects and base-charge variation effects. This is as it should be, for alpha is the current gain amplification factor, and must account for changes of current between the input and output. Indeed, the collector-to-emitter current gain cut-off frequency is related to $C_{b'e}$ by

$$f_{c\alpha} = \frac{q}{2\pi kT} I_E / C_{b'e}$$

and a measurement of $C_{b'e}$ at low frequencies may be substituted for a measurement of $f_{c\alpha}$ at considerably higher frequencies.

Another frequency-limiting factor, which may become significant as transistors are designed to minimize the above effects, is the presence of other capacitances, such as that of collector-to-base and collector-to-emitter. The collector-to-base capacitance may be reduced by reducing the area of the collector junction. There is also a contribution to this capacitance due to transit-time variation effects caused by a variation of the effective base thickness with the instantaneous collector voltage.

The above consideration will be applied in this report to the development of a *p-n-p* medium-frequency transistor. Similar considerations would also apply to *n-p-n* transistors.

Factors Governing Design

As will now be understood from the preceding discussion, in the design of high-frequency transistors it is important that, in addition to satisfying the requirements for good transistor action at audio frequencies, particular attention must be given to the following parameters.

1. Thickness of base layer.
2. Series base lead resistance.
3. Capacitance.

As pointed out in the general discussion the value of $C_{b'e}$ (or $f_{c\alpha}$) is degraded in proportion to the square of the distance between the emitter and collector junctions. This distance should be as small as can conveniently be obtained and a reasonable value of spacing to aim for is 0.0005 inch.

In the alloy-type transistor, an emitter-to-collector spacing of 0.0005 inch can be obtained in two ways: (a) by using a thick base wafer and a large impurity alloy penetration; or (b) by using a thin germanium base wafer with small impurity alloy penetration. With method (a), however, the junctions are hemispherical rather than parallel planes and consequently the possible hole path lengths may vary considerably. In method (b) the junctions are more nearly flat and parallel and this method was therefore chosen. With thin base wafers it is necessary that the impurity alloy depth of penetration be small. Accurate control is then necessary to prevent emitter-to-collector shorts.

The factors governing the penetration of indium by solution into the germanium base material were discussed.[1] A shallow penetration as desired here can be secured by: (a) a short firing time, i.e., shorter than that required for an equilibrium solution of the alloying action; (b) using a small volume of indium; (c) using a disc of indium-germanium alloy. The method described in this paper is a combination of all three but depends principally on the latter two.

In order to secure plane parallel junctions, heating at a high temperature is desirable to secure rapid wetting over the entire region of the junction. Various combinations of alloy disc composition and temperature are possible and several have been tried. However, as the atomic per cent of germanium in the indium-germanium alloy is raised above 10 per cent, it is more difficult to secure a homogeneous alloy, and the mechanical properties of the alloy become poorer. The alloy which has been found most advantageous at the present time contains 5 atomic per cent germanium in indium and this is the alloy used in all experiments described in this report.

As discussed previously it is important that $r_{bb'}$, the resistance of the germanium connecting the active junctions and the base tab be as small as possible. This resistance is shown in the input equivalent circuit of Fig. 1(a). The values of this equivalent circuit can be measured by means of specially designed bridges.[2]

The series resistance, $r_{bb'}$, consists of two regions, as shown in Fig. 1(b): region "A" directly between the junctions and region "B" between the junctions and the base connection. Obviously, the resistance of both these regions is reduced by using as low a resistivity base material as possible. In the "B" region, no limitation on resistivity exists. However, on the "A" region, two good junctions must be formed.

Shockley[3] has shown that in order to have a good hole emitter, the conductivity of the emitter section must be much higher than that of the base region. Experimentally, it is found that, as the resistivity of the base region is decreased, good junctions are more difficult to make. Here, then, an engineering compromise must be made between low base resistance and good junctions, as evidenced by high alpha and high back-resistance values. The resistance of the "A" region, which is directly between the junctions, is also reduced by the carriers injected by the emitter, i.e., conductivity modulation. Considering the present state of the art and the necessary engineering compromises, a resistivity of 0.6 to to 0.8 ohm-cm was chosen after some exploratory tests.

Let us examine region "B" to see how this resistance can be reduced. In order to obtain an idea of the magnitude or resistance involved in the region "B" for a thin wafer base, consider the simplified geometry shown in Fig. 2. The resistance in ohms between concentric cylinders is given by:

$$R = \frac{\rho}{2\pi L} \ln \frac{r_2}{r_1} \text{ ohms}$$

ρ = resistivity of germanium in ohm-cm
r_1, r_2 = radius of the inner and outer cylindrical surfaces.
L = wafer thickness in cm.

If r_1 is kept constant at 0.005 inch (the size of the emitter junction) and r_2 varied from 0.010 inch the curves of Fig. 2 are obtained.

If such a flat-wafer construction were used for the high-frequency transistor, and $\rho = 3$ ohm-cm., we see from Fig. 2 that a series resistance of 200 ohms would be introduced to a ring contact 0.040 inch from the center of the emitter. The curves also show that the contribution of the first few thousandths of an inch around the junctions is important. From these curves we can see the disadvantages of high resistivity germanium and of the flat-wafer construction.

[2] L. J. Giacolletto, "Equipments for measuring junction transistor admittance parameters for a wide range of frequencies," *RCA Review*, vol. 14, pp. 269–296; 1953.

[3] W. Shockley, "The theory of p-n junctions in semi-conductors and p-n junction transistors," *Bell Sys. Tech. Jour.*, vol. 28, pp. 435–489; 1949.

Several methods of reducing the resistance of the region "B" are possible. As a guide, methods other than those finally used are discussed first. One way to reduce the resistance of portion "B" is to solder a ring, a few thousandths of an inch larger than the junctions, to the germanium that surrounds the junctions. Great care must then be taken to keep the indium and solder from shorting and to obtain proper etching after assembly. Another possible procedure is to form junctions, protect the junctions with a suitable lacquer, and then plate a metal on the exposed base wafer. It is necessary to make a plated contact which is ohmic, i.e., does not inject an appreciable number of holes. Such ohmic contact can be made by sand blasting the base and then plating. A word of caution on these methods is in order. As has been shown in another report,[4] surface recombination is a critical factor in obtaining low-frequency transistor action. An ohmic contact or plated region of considerable area too near the emitter junction could lead to such a high recombination rate as to make the transistor unsatisfactory. A compromise is necessary which leaves sufficient low-recombination surface around the junctions, and yet not so much as to leave the base resistance high.

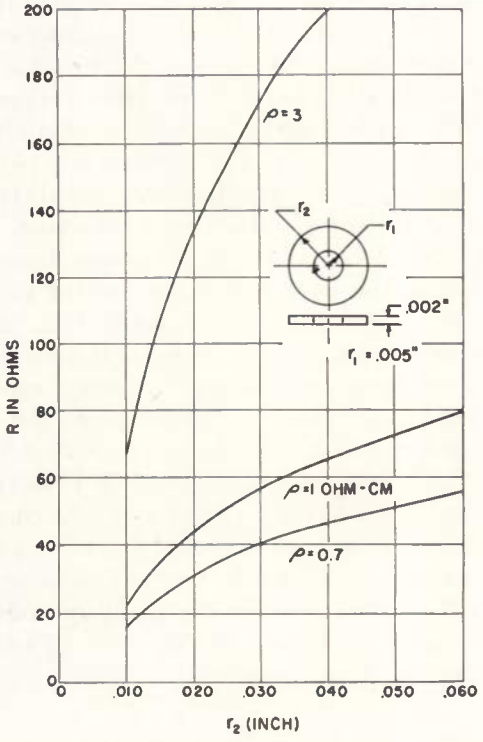

Fig. 2—Resistance between concentric cylinders of germanium.

After exploration of the above methods, a design was chosen in which only germanium surrounds the junctions. The structure is shown in Fig. 3 which is drawn approximately to scale. Here we see that, as soon as we leave the immediate vicinity of the junctions, the thickness of the wafer increases and thus decreases the series base lead resistance; since only ger-

[4] A. R. Moore and J. I. Pankove, "The variation of current gain with junction shape and surface recombination in alloy transistors," To be published in the PROC. I.R.E.

manium surrounds the junctions, low surface recombination can be attained at the same time.

The collector-to-base capacitance should be no larger than necessary. The junctions of the alloy-type transistor are usually of the abrupt or Shottky type. Formu-

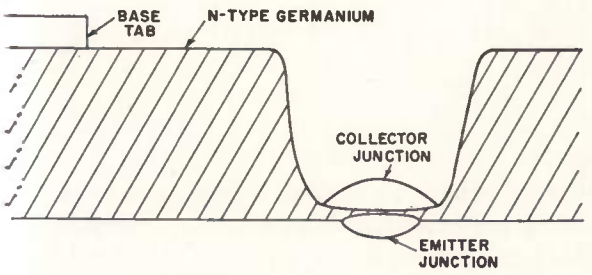

Fig. 3—Cross-section view of junction geometry.

las for this case are developed by Schockley[3] and, when expressed in convenient units, lead to the following equation for the capacitance of an abrupt junction.

$$C = 0.071 d^2 \frac{1}{\sqrt{\rho_b V_c}} \mu\mu\text{fd},$$

where:

d = diameter of the junction in mils
V_c = collector voltage
ρ_b = resistivity of the base material in ohm-cm.

Thus the diameter of the collector has the greatest influence and should be made small. This should, preferably, be done without decreasing α_{ce}, collector-to-emitter current amplification factor. In order to keep α_{ce} high, the area of the emitter must be kept to two-thirds or less than that of the collector.[4]

As the area of the emitter is reduced, the ratio of emitter circumference to area increases and surface recombination becomes more important. Consequently etching must be carefully done. Also, as the areas of the emitter and collector become small, alignment of emitter and collector is more difficult. As an engineering compromise, the emitter was made 0.010 inch in diameter and the collector 0.015 inch in diameter.

Electrical Measurements

The electrical engineer likes to reduce an electron device to an equivalent circuit composed of elements with which he is familiar so that he can apply standard circuit theory. Many types of equivalent circuit are possible and all engineers do not prefer the same types. The simplest circuit, with the fewest parameters that are all independent of frequency, is best but not always obtainable. Most of the published literature on transistors is in terms of the T-equivalent circuit or loop-derived parameters. Many engineers that are familiar with vacuum-tube terminology prefer the π-equivalent or nodal-derived parameters. One set of parameters can of course be derived from the other set if the set is complete.[5]

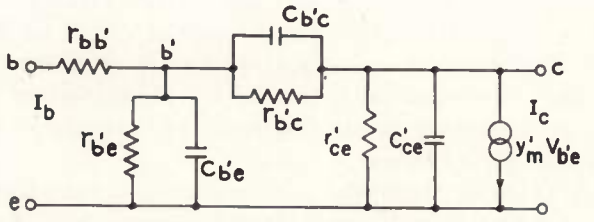

Fig. 4—Single-generator base-input π-equivalent circuit.

In Fig. 4 a single-generator π-equivalent circuit is shown for the common emitter connection. The circuit parameters have been tested by L. J. Giacoletto and found to be independent of frequency as long as the frequency is not too high. Although not yet completely investigated, the circuit is believed sufficiently close as long as $(\omega W^2/4D_p) < 0.2$ where ω = angular frequency, W = spacing between junctions and D_p = diffusion constant for holes. For the transistors here described the circuit is believed to be a reasonable approximation up to a frequency of the order of 3 mc. The elements shown in this circuit were measured in bridges previously described.[2] The values measured on 6 transistors are given in Table I. The value of W, the spacing between

[5] L. J. Giacoletto, "Terminology and equations for linear active four-terminal networks including transistors," *RCA Review*, vol. 14, pp. 28–46; 1953.

TABLE I

π-Equivalent Circuit Parameters Measured at $E_c = -6$ V, $I_e = -1$ ma

Transistor	A	B	C	D	E	F (Note 3)	Approx. TA-153 range
$r_{bb'}$ ohms	50	115	60	70	77	100	200– 500
$r_{b'e}$ ohms	1,250	1,500	1,400	600	430	500	600– 2,000
$C_{b'e}$ $\mu\mu$f	300	700	900	600	800	2,100	10,000–25,000
$r_{b'c}$ megohms	1.2	0.10	0.14	0.86	0.13	1.9	2
$C_{b'c}$ $\mu\mu$f	10.4	7.7	16.7	9.5	17	11	35
W mils (Note 1)	0.33	0.50	0.56	0.46	0.52	0.86	1–3
$f_{c\alpha}$ mc (Note 2)	20.5	8.8	6.8	10.2	7.7	2.9	0.4

Note 1—Calculated from $C_{b'e}$ as previously described.

Note 2—Calculated from $f_{c\alpha} = \frac{q}{2\pi kT} \frac{I_E}{C_{b'e}}$.

Note 3—This transistor is included to show the deviation to be expected from a higher-than-normal value of W.

TABLE II

T-EQUIVALENT CIRCUIT PARAMETERS $E_C = -6$ V; $I_E = 1$ ma

Transistor	A	B	C	D	E	F	Approx. TA-153 range
r_e ohms	4	18	20.6	14.6	19.4	16	5–20
r_b ohms	1,000	300	140	270	130	250	400–2,000
r_c megohms	0.4	0.06	0.16	0.40	0.16	0.45	0.5–3.0
α_{ce}	0.977	0.982	0.976	0.955	0.952	0.963	0.95–0.99
α_{cb}	41	51	40	21	19	25	20–100
I_{c0} μa	20	10	10	14	18	8	2–20

junctions, is calculated from $C_{b'e}$ by applying diffusion theory as previously described. The values of the circuit constants in the output were not determined because their effect is negligible in comparison with the other parameters and their measurement is difficult.

For comparative purposes, a range of values for TA-153 audio *p-n-p* transistors, described in reference 1 is included. The value of W for the TA-153 transistor was not as accurately controlled since the transistor was intended for audio use.

In Table II above the low-frequency parameters of the common-base T-equivalent circuit of Fig. 5 are shown, again in comparison with those of the TA-153 The parameters r_e and α_{ce} are slightly poorer in the high.

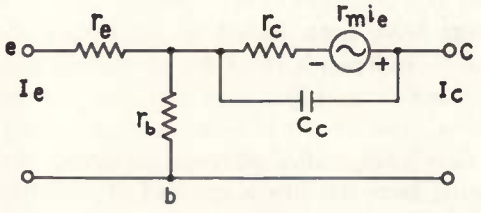

Fig. 5—Low-frequency emitter-input *T*-equivalent circuit.

frequency transistor because they were compromised to get better high-frequency performance. These low-frequency parameters, of course, give no clues to high-frequency performance but are included in the data for those who wish to make comparisons.

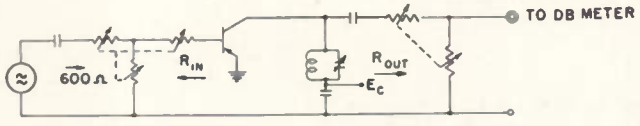

Fig. 6—Simplified diagram of circuit used for gain measurement versus frequency.

An important characteristic of a high-frequency transistor is, of course, the single-frequency power gain as a function of frequency. Fig. 6 shows a simplified schematic circuit of the test set used to measure power gain. Input matching is made only with the resistive component while the output is conjugate matched by adjusting the capacitance and output impedance. The feedback is not neutralized. In Fig. 7 is plotted the gain of a transistor with common emitter and common base

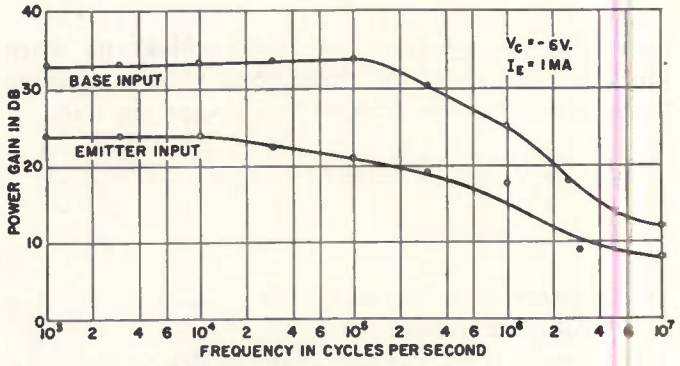

Fig. 7—Single-frequency power gain measured in the circuit of Fig. 6.

connections. In Table III the gain of the six transistors is shown for the various conditions described in the footnotes of the table.

TABLE III

GAIN AND NOISE FACTOR VALUES

Transistor	A	B	C	D	E	F	Approx. TA-153 range
4-kc gain db (Note 1)	35	37	39	34	34	38	35–48
455-kc gain db (Note 2)	35	32	39	35	34	34	<0–10
1-mc gain db (Note 3)	24	23	25	24	23	22	<0–7
10-mc gain db (Note 3)	13	10	11	9	8	5	—
1-kc noise factor db (Note 4)	14	23	33	24	16	20	15–25
1-mc noise factor db (Note 5)	4	5	8	7	5	6	—

All values of gain measured at $E_C = -6$ V, $I_E = 1$ ma

Note 1—$R_{in} = 500$ ohms, $R_{out} = 30{,}000$ ohms.
Note 2—Input and output conjugate matched and feedback neutralized.
Note 3—Input resistive matched and output conjugate matched in circuit of Fig. 6.
Note 4—$E_C = -1$ V, $I_E = 1$ ma, $R_{in} = 560$ ohms.
Note 5—$E_C = -1$ V, $I_E = 1$ ma, $R_{in} = 100$ ohms.

In Table III the noise factor is also given. There is a considerable variation of 1-kc noise factor among various units at the present time, but the one-megacycle noise factor does not vary as much. In Fig. 8 the noise factor is plotted as a function of frequency up to 1 mc.

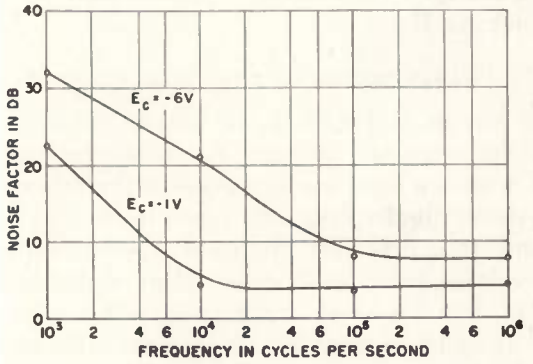

Fig. 8—Noise factor versus frequency; common emitter connection.

Experiments with a simple oscillator circuit show that the high-frequency transistors described will oscillate at frequencies up to 75 mc.

Conclusions

A *p-n-p* triode transistor was achieved, capable of as much as 39-db gain at 455 kc when used in a neutralized, conjugate-matched circuit. Without neutralizing, 12 db at 10 mc can be obtained. The noise factor at 1 mc is 4 to 8 db.

A spacing between junctions of 0.0005 inch and a low base-lead resistance is obtained by using a low-resistivity base section with the collector junction located at the bottom of a well in the germanium. An indium-germanium alloy is used to obtain more planar junctions.

With the continued improvement in the techniques of making and controlling junctions further improvements by the use of lower resistivity germanium and a reduction in all dimensions are possible.

Acknowledgment

The gain and noise measurements were made on equipment designed and built by P. Werenfels. The neutralized gain measurements at 455 kc were made by D. D. Holmes and oscillation tests by G. B. Herzog. Analysis of diffusion reactances by H. Johnson was most valuable. Mrs. E. Moonan made many of the measurements.

An Amplitude Stabilized Transistor Oscillator*

E. R. KRETZMER†, SENIOR MEMBER, IRE

Summary—**This paper describes an audio-frequency oscillator using two junction transistors in push-pull Class C operation. The circuit features an over-all efficiency as high as 75 per cent, low distortion, signal-to-noise ratio better than 100 db, and a high degree of amplitude stability. The amplitude can be held to the order of one per cent over wide ranges of supply voltage, output loading, and temperature. The stabilization is achieved by comparing the output amplitude with a reference voltage, which is obtained by means of the so-called break-up effect in semiconductor *p-n* junctions. An analysis of the circuit is appended.**

Basic Oscillator Circuit

THE BASIC FORM of the circuit to be described is analogous to a well-known vacuum-tube circuit—the plate-tuned oscillator with magnetic coupling from plate to grid. A push-pull transistor version appears in Fig. 1. The transistors are used in the grounded-emitter connection, which makes the base analogous to the vacuum-tube grid; the major difference is that the base impedance is finite in the control range, whereas the grid impedance is virtually infinite over most of the control range.

Like its vacuum-tube counterpart, this circuit will oscillate with the maximum peak swing permitted by the B-supply. The extraordinarily linear characteristics[1] of junction transistors permit the peak swing to be within one- or two-tenths volt of the supply voltage. The output loading and the transformer turns ratio determine the type of operation (Class A, B, or C) and

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

[1] R. L. Wallace, Jr. and W. J. Pietenpol, "Some circuit properties and applications of n-p-n transistors," PROC. I.R.E., vol. 39, pp. 753–767; July, 1951. *Bell Syst. Tele. Jour.*, vol. 30, pp. 530–563; July, 1951.

the angle of conduction, as shown in Appendix I. Unlike the vacuum-tube grid current, the transistor base current has the same waveform as the load current. The direct current flowing into the base bears a definite relationship to the direct collector current (depending largely on the parameter α).[2] The required base current must be supplied through the resistor G and any additional shunt paths capable of bleeding current into the base circuit. One such additional shunt path, though perhaps not obvious, is the back resistance of the collector-to-base junction of the transistor. The associated current, usually referred to as I_{co}, is in effect bled into the base circuit, as if by way of an external current source, and therefore actually augments the base current. Far from being an asset, however, I_{co} is generally a severe liability, since it is sharply dependent on temperature.

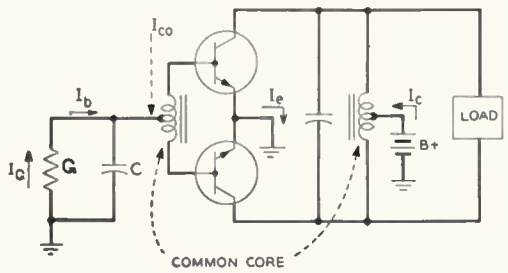

Fig. 1—Basic oscillator circuit.

Method of Stabilization

Amplitude stabilization can be introduced into the circuit of Fig. 1 by comparing the output voltage swing to a fixed reference voltage and feeding back the difference voltage so as to control the base current. A simple embodiment of this principle appears in Fig. 2. In this

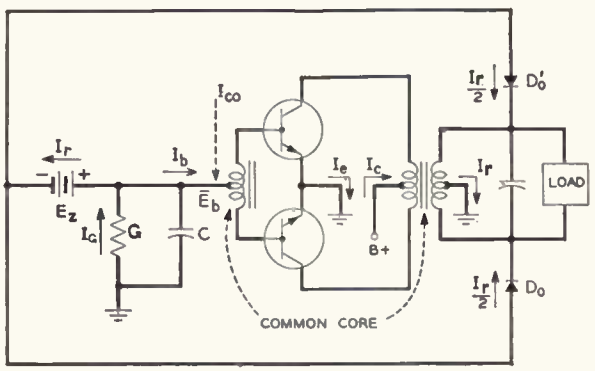

Fig. 2—Oscillator with amplitude stabilization.

circuit, whenever either end of the output winding attempts to swing negatively below the voltage $\bar{E}_b - E_r$ either diode D_0 or D_0' conducts a pulse of current. Provided that I_b is sufficiently large to sustain an output swing larger than $\bar{E}_b - E_r$, these current pulses flow periodically at twice the frequency of oscillation, giving rise to the average current I_r indicated in Fig. 2. This current detracts from the base current I_b, thereby reducing the amplitude of oscillation until it barely exceeds $\bar{E}_b - E_r$. The magnitude of I_r is not critical and can be adjusted to any desired value by adjusting G (and hence I_G). The current bled externally into the base winding, I_G, must always equal the total current flowing out, $I_b + I_r$. More will be said about these currents and their connection with the amplitude regulation in a later section, and a complete analysis is given in Appendix II.

[2] A list of symbols and their meanings precedes Appendix I.

Refinements of the Basic Circuit

The circuit of Fig. 2 is, of course, no more stable than the reference voltage E_r. Since the current I_r flows in such a direction as to charge the reference battery, its terminal voltage, E_r, tends to rise as a function of time. It is desirable, therefore, to obtain the reference voltage from some source more stable and more reliable than a battery. Such a source has recently become available in the form of the semi-conductor (silicon) junction diode. Such a diode generally has extremely high incremental resistance (thousands of megohms) in the back direction up to a certain voltage depending on the thickness of the junction and the conductivity of the semi-conductor. At this voltage, commonly called the break-down voltage, the incremental resistance changes abruptly to a much lower value (a few hundred ohms or less). Consequently, as in the case of a gas-tube glow discharge, the voltage cannot be raised further without passing relatively large currents. The voltage characteristic of a typical silicon junction

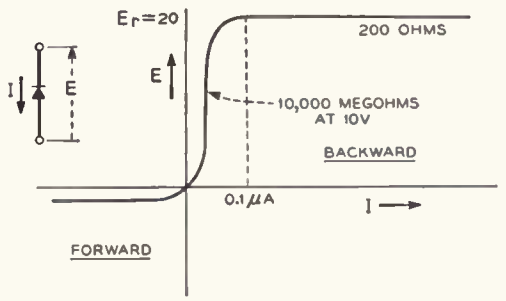

Fig. 3—Typical silicon junction diode characteristic showing break-down effect.

diode is shown in Fig. 3. As soon as any appreciable reverse current is forced through the diode, its voltage remains at the break-down voltage of 20 volts. If the current is raised from, say, 10 microamperes to 100 microamperes, the voltage increases by about 20 millivolts or 0.1 per cent.

Examination of Fig. 2 shows that the function of the reference battery can be performed equally well by a junction diode, hereafter called the regulator diode. Since a current I_r is required anyway in order to establish amplitude stabilization, this current can serve also the purpose of maintaining break-down voltage. The "cathode" and "anode" take the place of the plus and minus battery terminals, respectively.

The regulator-diode-stabilized oscillator is shown in Fig. 4 with several added refinements, indicated by A, B,

C, and D. "A" is an output level adjustment. The associated small winding on the main transformer furnishes rectified voltage, of which an adjustable portion, E_A, appears in series with the fixed reference voltage. Similarly, "B" inserts another variable voltage, E_T, which increases with decreasing temperature as a result of the increasing resistance of the thermistor T. The purpose of this temperature-variable voltage, E_T, is to compensate for the temperature dependence of the regulator voltage, E_r, which rises approximately 0.1 per cent per degree (Centigrade) temperature rise.[3] The thermistor and associated resistor are chosen so that E_r+E_T is nearly constant as a function of temperature. The action of E_T will be referred to as series compensation, since E_T is inserted in series with E_r.

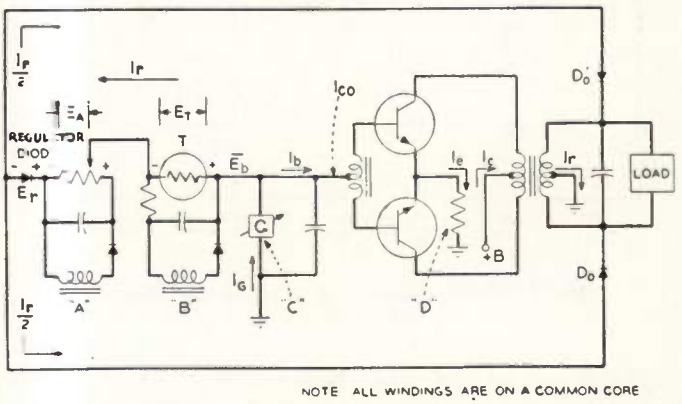

Fig. 4—Amplitude-stabilized oscillator using regulator diode with amplitude adjustment and series temperature compensation.

The box marked G (refinement "C") takes the place of the resistor G (in Fig. 2). It is a somewhat complicated variable conductance, being dependent on manual adjustment, temperature, and possibly the base voltage $\bar{E}_b$ or the oscillator output loading.[4] The temperature dependence is intended primarily to compensate for the variation of I_{co} with temperature. This action is termed shunt compensation, since G is effectively in shunt with the regulator-diode path.

Refinement "D" is simply a degenerative emitter resistor which helps distribute the load evenly among the two transistors and tends to stabilize the operation of the oscillator.

Typical Numerical Values

Before inquiring further into the operation of the circuit, let us consider some typical numerical values. Suppose that the transformer output ratio is unity and that the load (including transformer losses and all other ac losses) requires 20 milliwatts at 10 volts rms per each half of the transformer primary winding. The rms fundamental collector current must then be 1 milliampere in each side of the collector winding. The corresponding average direct collector current per side would be 1.4 ma or more for Class A operation, or possibly as low as 0.7 ma for extreme class C operation (conduction angle of, say, 60 degrees or less). For a conduction angle in the order of 90 degrees, the total average collector current, I_c, is approximately 1.5 ma; the peak current is about 5 ma. As shown in Appendix I, such a conduction angle, with the numbers assumed above and with an emitter resistor of 50 ohms, yields an average base voltage, $\bar{E}_b$, of approximately -1 volt. The base winding, with one-tenth as many turns as the collector winding develops 1 volt rms or 1.4 volts peak per side; the upper 0.4 volt[5] lies above the zero axis, thus forming the conduction cycle.

Knowing $\bar{E}_b=-1$ volt and $\hat{E}_c=14$ volts peak (per side), we may conclude that the required regulator voltage is $E_r=13$ volts. Furthermore, to insure proper regulation, the collector supply voltage must exceed the 14-volt collector swing by at least a few tenths of a volt. The peak current during the 90-degree conduction period is 5 ma, which calls for a few tenths of a volt collector-to-base voltage in the average junction transistor. Consequently a 15-volt battery is adequate. Any higher voltage is equally adequate (so long as the maximum voltage between collector and base does not exceed the collector-to-base junction's break-down voltage), but the excess over 15 volts in conjunction with the 1.5-ma collector drain gives rise to a substantial dc power loss.

Now consider the currents bled into and out of the base winding. This requires some assumptions as to transistor parameters. Suppose that α is constant at 0.95 over the conduction cycle while I_{co} averages a total of 20 microamperes for both transistors, at room temperature. Disregarding I_{co} for the moment, we may say that, for every milliampere of emitter current, we have 0.95 ma of collector current which in turn requires 50 μa flowing into the base. The effect of I_{co} is twofold: first, it constitutes a useless component of the collector current, a relatively negligible effect; second, and much more important, it constitutes a current flowing into the base, so that, instead of 50 μa, only 30 μa need be supplied externally. The current flowing into the base externally is termed I_b (see Fig. 4). In addition to I_b, the regulating current I_r must be supplied externally. Both I_b and I_r are supplied through the variable conductance G, so that we have the relation

$$I_G = I_b + I_r. \tag{1}$$

Furthermore, I_b+I_{co} is a fraction $1-\alpha$ of the emitter current, I_e, or $(1-\alpha)/\alpha$ of the useful collector current, I_c-I_{co}. We therefore have the following relation between the various currents: $I_G+I_{co}=[(1-\alpha)/\alpha](I_c-I_{co})+I_r$, which reduces to

[3] Smaller temperature coefficients have been achieved more recently.

[4] This so-called loading compensation is a refinement discussed in detail in a later section.

[5] This means that 0.4 volt is required to produce the 5-ma peak collector current. For typical values of α, r_b, and r_e only one third of the 0.4 volt is required from base to emitter; and the remaining two thirds of this appears across the 50-ohm stabilizing emitter resistor, which makes the conduction angle less dependent on transistor parameters.

$$I_G + \frac{I_{co}}{\alpha} = \frac{1-\alpha}{\alpha} I_e + I_r. \qquad (2)$$

The two middle terms of (2) have been specified; I_G can be adjusted (by adjusting G) so as to yield the desired value of I_r. Theoretically, the regulating current, I_r, need be no more than a fraction of one microampere to insure proper Zener action. From a practical viewpoint, however, I_r must be large enough so that variations in the other currents of (2) will not carry I_r to zero. The expected variations depend on the variations with time, temperature, and loading of I_{co}, $(1-\alpha)/\alpha$, I_e, and I_G. Suppose that there are no known significant aging effects, but that the temperature range is 20 degrees C, and that the loading might increase from 20 milliwatts to 25 milliwatts.

Consider, first, the effect of temperature alone. A twenty-degree rise, starting from room temperature, may easily cause an increase in I_{co} of 100 microamperes for the two transistors. (This depends, of course, on the particular transistor and may be much smaller in many cases, or larger in a few cases.) In addition to the increase in I_{co}, there may also be an increase in α perhaps from 0.94 to 0.96, so that $[(1-\alpha)/\alpha I_e]$ would decrease from 98 microamperes to 63 microamperes. Equation (2) shows that these changes both work in the same direction, so that if I_G remained constant, I_r would increase by 140 microamperes. Such an increase may well be intolerable, both from the power dissipation and from the regulation viewpoint. Compensation can be achieved by making G (see Fig. 4) change with rising temperature so that I_G changes like $[(1-\alpha)/\alpha]I_e - (I_{co}/\alpha)$, thus keeping I_r constant. In practice, this shunt compensation involves various technical problems (see following section V), but the variation in I_r can usually be held to ± 15 microamperes or less.

As for the assumed loading increase of 5 milliwatts, this would increase the collector current by 25 per cent (of 1.50 ma). For $\alpha = 0.95$, the component $[(1-\alpha)/\alpha]I_e$ would increase from about 80 microamperes to 100 microamperes. Unless I_G is somehow changed to compensate, the 25 per cent loading increase will cause a 20 microampere decrease in I_r.

The preceding considerations of temperature and loading effects dictate that, for the orders of magnitude assumed, I_r should average at least 40 microamperes in order to maintain regulation at all times. The magnitude of I_r can be likened to the pressure of a spring, holding an "output" member in firm contact with a "reference" member; little pressure is needed, but we must be sure that the pressure *never* reaches zero as a result of temperature variations or other external influences.

Practical Circuit

If we incorporate the numerical values of the preceding section in the circuit of Fig. 4 and add shunt temperature compensation, we obtain the circuit shown in Fig. 5. A few additional changes have been made in order to increase the usefulness of the oscillator.

All windings shown are on a common transformer core, their turns ratios being as indicated by the adjacent numbers. The base winding performs triple duty, providing, aside from the ac feedback voltage, a positive dc voltage through diode D_1 for shunt compensation, and a negative dc voltage through D_2 for series compensation.

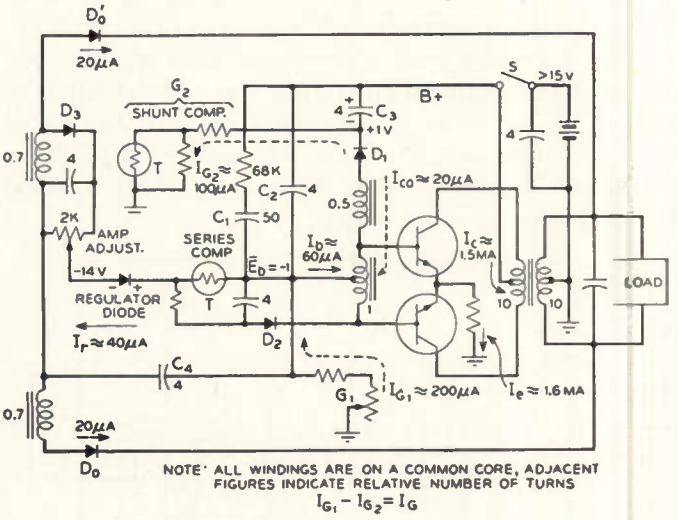

Fig. 5—Practical version of amplitude-stabilized transistor oscillator.

The principal difference between Fig. 4 and Fig. 5 lies in the implementation of G. In Fig. 4, G is indicated simply as a variable conductance which supplies the current I_G from ground to the base winding. Equation (2) of the preceding section has shown that the required magnitude of I_G depends on transistor parameters α and I_{co}, as well as the desired magnitude of the regulating current I_r. With the typical values assumed ($\alpha = 0.95$, $I_e = 1.50$ ma, $I_{co} = 20$ μa, $I_r = 40$ μa), the required value of I_G turns out to be 100 microamperes. In Fig. 5, instead of bleeding the required value of I_G into the base winding, we bleed in a much larger value, $I_{G1} = 200$ microamperes. The excess of 100 microamperes, designated I_{G2}, is pumped back to ground the required voltage is obtained by rectification through diode D_1 of ac obtained from the base winding plus a small auxiliary winding. Thus we have $I_G = I_{G1} - I_{G2}$, and I_G can be, say, decreased either by decreasing I_{G1} or by increasing I_{G2}.

There are a number of reasons for performing this seemingly complicated maneuver. First, proper temperature compensation for I_{co} (and possibly α) requires that I_G be decreased with rising temperature. Commercial thermistors, however, have conductances which *increase* with temperature and can therefore be incorporated only in G_2. Second, for numerical values different from those assumed, notably for higher I_{co}, higher α, or lower I_e, the required value of I_G may be very close to zero, or possibly even negative at the higher temperatures. Nevertheless, a change of more than 100 microamperes may still be required for shunt temperature compensation. For these reasons it is ex-

pedient to implement I_G as the difference between two larger currents, I_{G1} and I_{G2}.

Two other differences between the practical circuit of Fig. 5 and the illustrative circuit of Fig. 4 are in the series compensation and the amplitude adjustment, respectively. The series compensation, while unchanged in principle, now obtains its rectified voltage from one-half of the base winding, thus obviating the need for an added transformer winding. On the other hand an additional winding is used to advantage for the amplitude adjustment. In Fig. 4, an adjustable amount of rectified voltage is simply inserted in series with the regulator diode. In Fig. 5 we insert, in addition, the ac voltage from which the rectified voltage is obtained (by means of the two balanced windings marked 0.7). The object of this is to maximize the adjustment range while minimizing the associated power dissipation. Thus, by making the two balanced windings and the diode D_3 reversible and by adjusting the potentiometer, we can secure a continuous voltage change equal to four times the peak ac voltage of either of the two windings. If that peak voltage is 1 volt, the total adjustment range is 4 volts (making the output range 14 ± 2 volts peak). The power dissipated in the $2K$ potentiometer is $\frac{1}{4}$ milliwatt, while the maximum resistance inserted in the regulating path is $2K/4$ or 500 ohms. A 20-microampere change in I_r would therefore, at worst, cause a 10-millivolt change causing less than 0.1 per cent change in output amplitude. The reversal of the two windings and of diode D_3 can, of course, be instrumented simply by a four-position switching arrangement.

A further point in connection with the oscillator circuit of Fig. 5 involves capacitors C_1, C_2, and C_3. C_1 and the series-connected $68K$ resistor supply a surge of current into the base winding when the battery switch, S is first closed. This is necessary in order to start the oscillation, since zero bias would mean "cutoff" (unlike the case of the vacuum tube, where a negative bias is usually required to produce the cutoff condition). For the same reason, C_2 (which serves to prevent sudden changes in average base voltage) and the rectifier filter capacitor C_3 return to $B+$ instead of to ground.

All capacitors in the circuit, except the tuning capacitor and the starting unit C_1, are simple bypass or filter units. Their minimum required values therefore depend on the lowest operating frequency. The value of 4 μf is suitable for operation in the upper audio range; tantalum units are ideal for this application because their size is commensurate with that of the transistors and because their leakage is extremely small.

Capacitor C_4 bypasses the regulator diode as well as the amplitude adjustment and compensating voltages in series with it, thus making I_r flow in the form of a smooth dc instead of short pulses and consequently minimizing the effective resistance of the regulating path. It should be noted, however, that there are, at this early stage of the art, so-called soft-break-down diodes in which the current must exceed perhaps 100 microamperes or more in order to establish the required low incremental resistance. In such a case it would be wiser to leave the regulator diode unbypassed, so that the relatively high current peaks of I_r (1 ma peak for 40 μa average) drive the diode into true break-down saturation.

It must be noted here that all quantitative considerations in this and the preceding section merely concern the static operating conditions. The amplitude regulation, of course, constitutes a feedback system, subject to gain and stability considerations. The feedback loop is analyzed in detail in Appendix II.

Achievement of Zero Output Impedance

One of the manifestations of good amplitude stabilization of an oscillator (or any power source) is low output impedance. In other words, a substantial change in loading causes very little change in voltage.

As is shown in Appendix II, the output impedance of the transistor oscillator depends on a number of circuit parameters, such as transistor α, base and emitter resistances, and the resistance of the regulator diode and other elements in series with it. With the typical values previously assumed, the output impedance from either side to ground is computed to be roughly 200 ohms, which is well born out by direct measurements. Thus, connecting a $20K$ resistance from one side to ground (representing a 5 milliwatt or 25 per cent loading increase) reduces output voltage about one per cent.

For many applications this sort of performance is entirely adequate. In some cases, however, it may be desirable to vary the loading over wide ranges without affecting the output voltage. In other words, zero output impedance may be desired. This is achievable in the present circuit by relatively simple means.

An increase in output loading causes an increase in collector current which in turn calls for an increase in the base current, I_b. Unless this required increase is supplied by a corresponding increase in I_G (see Fig. 4 and (1) and (2)), the regulating current I_r will decrease, thus tending to lower the output voltage.

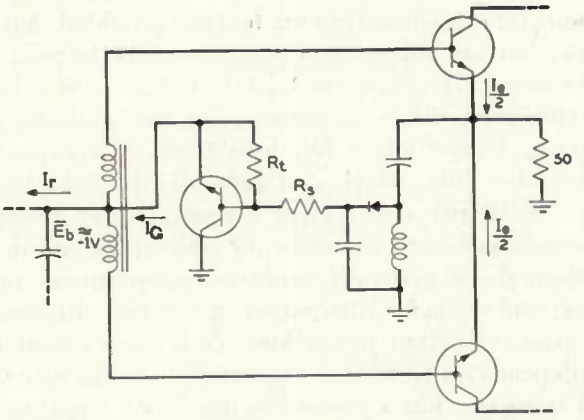

Fig. 6—Base circuit of oscillator with added transistor for loading compensation (to achieve zero output impedance).

Fig. 6 illustrates a method of adjusting I_G automatically with loading, so that I_r remains substantially constant regardless of loading. The current I_G is now sup-

plied through an additional junction transistor, which is controlled by the emitter current of the oscillator. Thus, if the load is increased, both the ac and dc components of I_e will increase. The ac component of the emitter voltage is stepped up to a higher voltage and impedance level before being rectified and applied through R_s to the base of the auxiliary transistor. Consequently, the current flowing into this base varies directly with I_e and hence the oscillator loading. The transistor amplifies this current and then feeds it into the oscillator base winding as I_G.

The resistors R_s and R_t can be chosen so as to give the compensation required to keep I_r constant. If the value of R_t is raised, I_G increases and over-compensation may be achieved, so that the output voltage actually rises with increased output loading. In addition to providing load compensation, the auxiliary transistor has much higher incremental resistance than a resistor G supplying the same direct current, I_G. This means higher regulating loop gain, from Appendix II. As in the case of Fig. 5, it may be desirable to instrument I_G as the difference I_{G1} minus I_{G2}. In this case, "G_1" may be the auxiliary transistor, while "G_2" may be another transistor in combination with a thermistor.

There are, of course, a great many methods of instrumenting I_G along with all its compensations ranging from the use of ordinary thermistors to that of n-p-n or p-n-p junction transistors, junction diodes, or combinations of some or all these devices. The degree of sophistication depends largely on the extent of the stability requirement. Some of the results obtainable are discussed in the next section.

Performance

The transistor oscillator described in this article exhibits unusual performance principally in two respects: efficiency and amplitude stability. The high efficiency is primarily a result of characteristics of junction transistors, while stability is that of the regulating system.

Over-all efficiencies as high as 75 per cent have been obtained. Such efficiencies are feasible provided that the supply-battery voltage just barely exceeds the peak collector swing. Doubling the battery voltage would halve the efficiency, the extra power going into collector dissipation. Power losses fall into three categories: the transformer loss, which can generally be held to the order of 10 per cent of the output; power losses in various bleeder and compensator resistances and in the regulator diode, generally not exceeding one or two milliwatts; and collector dissipation in the two transistors, the most important power loss. To a large extent this loss depends on transistor characteristics. For example, each transistor has a power loss due to I_{co}, equal to the product of collector-to-base voltage and I_{co} integrated over the cycle. At the higher temperatures, if I_{co} per unit exceeds 40 or 50 microamperes at 40 volts collector voltage, this may account for a loss of one or two milliwatts. During the conduction period an additional inevitable loss is incurred, namely the average product of the useful collector current and the collector-to-emitter voltage. In the limiting case of infinitesimal conduction angle and complete collector saturation (zero minimum collector voltage) this power loss would approach zero; in the practical case, the regulating circuit holds the minimum collector voltage to one, two, or several volts, depending on the battery voltage. The resulting power loss is of the order of 20 to 50 per cent of the ac power generated. Thus under typical conditions, with the battery voltage exceeding the regulated collector swing by 10 or 20 per cent, the over-all efficiency may be in the order of 50 per cent.

The output voltage stability is, of course, the other remarkable feature of this oscillator. Battery voltage variations of, say, 50 per cent change the output by only fractions of one per cent, provided that the battery voltage exceeds the regulating voltage and that the transistors maintain their typically high collector resistance over the entire voltage range.

Similar constancy of amplitude is maintained when various typical transistors are plugged into the circuit. With only a single transistor, the oscillator operates uniformly at an amplitude two or three per cent lower than with push-pull operation. This difference in amplitude results from a readjustment in average base voltage which provides the required increase in peak base voltage (required since the single transistor must now deliver current pulses twice as large as before).

Amplitude variations caused by temperature changes may be held to one or two per cent over a range of 40 degrees C. provided appropriate compensation is used. Without this, the amplitude might rise as much as 0.2 per cent per degree of temperature rise.

Amplitude stability with variable loading depends on whether or not the refinement of load compensation (requiring an additional transistor) is used. With such compensation, it is possible to double the output loading, say from 15 to 30 mw without changing the amplitude more than a small fraction of one per cent. Without loading compensation, an increase, say from 20 to 25 mw, might cause a drop in output of approximately one per cent.

The good loading stability, in combination with the high efficiency, provides us with an adjustable source of regulated power. The oscillator, feeding a junction-diode rectifier, for example, may serve as a highly efficient regulated dc power supply. Thus, with a 10-volt battery, direct voltages of up to 200 volts at 20 to 200 microamperes have been obtained at an over-all efficiency of 60 per cent and with the same stability performance described in the preceding paragraphs.

Another important type of stability is the so-called short-time stability—or freedom from noise. The principal noise contributors are the transistors. Diode noise is negligible if silicon junctions are used for amplitude regulation (D_0 and D_0') and for high-voltage rectification. The transistor noise power varies approximately inversely as the frequency, so that it is concentrated at low frequencies. The resulting low-frequency amplitude

modulation of the oscillator output is conveniently measured by rectifying the output and applying it to a recording meter. For example, the 200-volt dc output mentioned above might have a superimposed noise voltage of roughly 2 millivolts peak to peak. The ratio of rectified voltage to peak-to-peak noise voltage would then be 10^5 or, loosely, 100 db. In practice, this ratio has ranged from 70 db (for the noisiest transistors) to 120 db (for the quietest units). Because of the inverse-frequency variation of the noise, all octaves, within limits, contribute similar amounts of noise. For the signal-to-noise ratios quoted above, the effective bandwidth in which the noise was measured covered ten octaves. The 1-kc noise figures of the transistors used range from roughly 20 db to 60 db.

Performance characteristics other than efficiency and amplitude stability are much like those of any oscillator. The frequency stability depends, of course, mainly on the tuning elements; to the extent that it depends also on the loading effects and operating conditions, the frequency stability is somewhat improved by the amplitude stabilization. Second harmonic distortion, with an output Q of about 15, is 60 to 70 db below the fundamental with push-pull operation, or 30 to 40 db with one transistor removed. Third harmonic distortion is in the order of -40 db with single-sided or push-pull operation.

To sum up, we have an oscillator circuit capable of supplying roughly 10 to 100 mw of power at efficiencies better than fifty per cent and amplitude stability in the order of one per cent. Aside from serving as a compact battery-powered source of ac, the oscillator, in conjunction with junction diodes, can provide stable dc outputs over a wide range of voltages.

Acknowledgment

The idea of using the break-down effect to obtain a reference voltage in this application is due to Dr. B. M. Oliver. His helpful suggestions, as well as those of Mr. R. E. Graham, are hereby gratefully acknowledged.

Definitions of Symbols

Note: *N-p-n* transistors are assumed throughout. For *p-n-p* transistors, reverse all voltage polarities and current directions. Unless otherwise noted, all currents are total currents for both transistors.

α: Emitter-to-collector current multiplication ratio. *Note:* this is assumed to be independent of operating conditions and therefore applies to incremental or direct currents (not including I_{co}). For example, $\alpha=(I_c-I_{co})/I_e$, provided that the collector is not saturated.

I_{co}: Leakage current of collector-base junction. This current is superimposed on the useful (controllable) collector current.

I_c: Total direct collector current $[=\alpha I_e+I_{co}]$.

I_b: External base current (dc flowing into base).

I_e: Emitter current $[=I_b/(1-\alpha)]$.

I_G: Current supplied to base circuit (dc).

I_r: Regulating current abstracted from base circuit (dc).

E_b: Instantaneous base voltage.

$\bar{E}_b$: Average base voltage.

$E_{b\ peak}$: Peak positive base voltage.

E_r: Regulator voltage.

E_A: Voltage added to E_r to adjust amplitude.

E_T: Voltage added in series with E_r to compensate for temperature effects

r_b: Incremental base resistance.

$\bar{r}_b$: DC base resistance.

r_e: Incremental emitter resistance.

$\bar{r}_e$: DC emitter resistance. *Note:* any external emitter resistance may be included in r_e and $\bar{r}_e$.

P_{ac}: Total ac power.

$\tilde{E}_c$: Collector voltage amplitude.

$\tilde{E}_b$: Base voltage amplitude.

$\tilde{I}_c$: Collector current amplitude.

$i_c(x)$: Instantaneous collector current.

n: Turns ratio from collector winding to amplitude control winding; e.g., $n=1$.

N: Turns ratio from collector winding to base winding; e.g., $N=10$.

e_0: Incremental change in output amplitude, collector to ground.

e_1: Assumed incremental change in internally generated voltage, collector to ground.

e_1': Value of e_1 caused by a voltage change $e_{b\ peak}$.

e_s: Change in regulator voltage, E_s.

e_b: Change in average base voltage, $\bar{E}_b$.

μ, β: Factors whose product equals the loop transmission, called loop gain for brevity.

$e_{b\ peak}$: Change in peak base voltage, $E_{b\ peak}$.

k: Ratio of change e_1' to change $e_{b\ peak}$ (which causes change e_1').

R_N: Resistance of regulating path.

R_1: Incremental base-circuit input resistance, open loop.

R_0: Output resistance (collector to ground), open loop.

R_{out}: Output resistance (collector to ground), closed loop.

Appendix I

The purpose of this appendix is to show the dependence of average and instantaneous base voltages as well as conduction angle on feedback turns ratio, transistor parameters, and loading of the oscillator. Consider the numerical example of the text, in which we have the following values:

$P_{ac}=20$ mw (total ac power)

$\tilde{E}_c=14$ volts (collector amplitude)

$\tilde{E}_b=1.4$ volts (base amplitude, for 10:1 feedback turns ratio)

$\tilde{I}_c=1.4$ ma (fundamental collector ac amplitude).

The corresponding waveforms are shown in Fig. 7. The four quantities given above, in conjunction with α,

$\bar{r}_b$, and $\bar{r}_e$ (including the external emitter resistor), determine the required peak positive base voltage, which in turn determines the average base voltage and the conduction angle. The peak positive base voltage, 0.4 v, is such that the resulting conduction angle, 90 degrees, and peak collector current, 5 ma, yield a fundamental component of collector current equal to 1.4 ma peak.

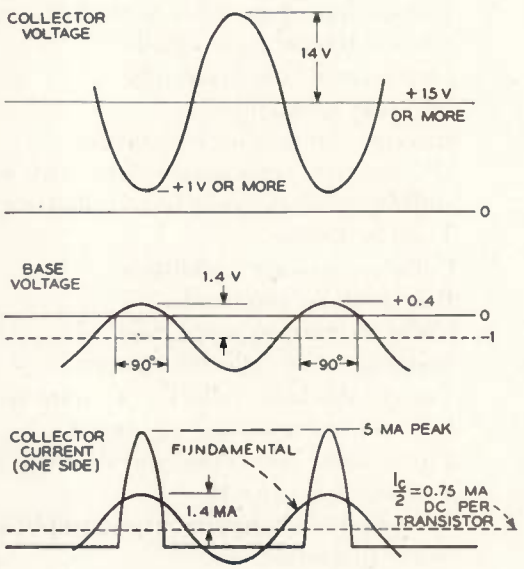

Fig. 7—Oscillator waveforms.

This can be verified as follows: For an instantaneous base voltage E_b, we have (ignoring I_{co})

$$I_b = E_b \Big/ \left(\bar{r}_b + \frac{\bar{r}_e}{1-\alpha}\right) \tag{3}$$

and

$$I_c = \frac{\alpha E_b}{1-\alpha} \quad \frac{1-\alpha}{\bar{r}_b(1-\alpha)+\bar{r}_e} \approx \frac{E_b}{\bar{r}_e + (1-\alpha)\bar{r}_b}. \tag{4}$$

Typical values are:

$$\alpha = 0.95,$$
$$\bar{r}_b = 400\,\omega;$$

at the peak current of 5 ma the internal dc emitter resistance is in the order of 10 ω, to which we must add the external 50 ω resistor, so that we have

$$\bar{r}_e \approx 60\,\omega,$$
$$\bar{r}_e + (1-\alpha)\bar{r}_b \approx 80\,\omega.$$

Then (4) yields

$$I_c \approx E_b/80 \tag{5}$$

and for $E_b \approx 0.4$ volt,

$$I_c \approx 5\text{ ma},$$

as shown in Fig. 7.

It remains only to check the fundamental ac component of the collector current. If E_b reaches a peak value of 0.4, its average value must be -1 volt, and if cutoff occurs at approximately zero volts, then the conduction angle is approximately $2\sin^{-1} 1.0/1.4 = 90$ degrees. The collector current (ignoring I_{co}) is therefore expressible as

$$I_c = 15[\cos x - 0.7], \quad -45° < x < +45°. \tag{6}$$

I_1 (fundamental)

$$= \frac{2}{\pi}\int_0^{45°} 15[\cos x - 0.7]\cos x\,dx$$

$$= \frac{30}{\pi}\left[\left(\frac{x}{2} + \frac{\sin 2x}{4}\right) - 0.7\sin x\right]_0^{\pi/4} = 1.4\text{ ma}$$

$\frac{1}{2}I_c(dc)$

$$= \frac{1}{\pi}\int_0^{45°} 15[\cos x - 0.7]dx$$

$$= \frac{15}{\pi}[\sin x - 0.7x]_0^{\pi/4} = 0.75\text{ ma per transistor.} \tag{7}$$

The results obtained from (5) and (6) above prove that all the values shown in Fig. 7 are consistent with the assumed parameter values and operating conditions. This is also borne out experimentally.

Appendix II

In this appendix the oscillator with its amplitude stabilization is analyzed as a feedback control system. The object is to determine the regulator loop gain, output impedance and stability margin of the oscillator.

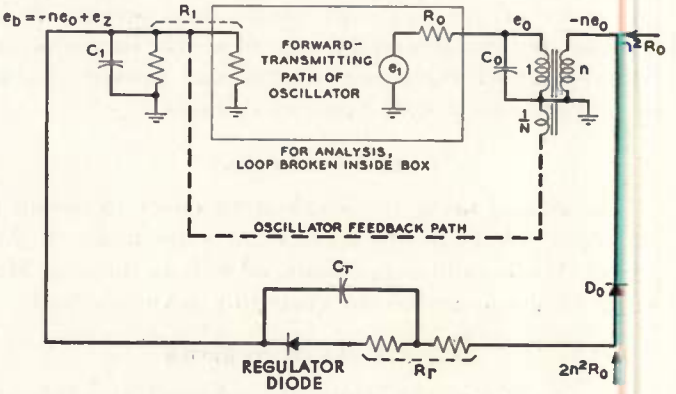

Fig. 8—Schematic representation of oscillator and amplitude control loop.

Fig. 8 represents the oscillator and its control loop. All voltages shown represent increments of the steady-state direct or peak alternating voltages at the respective points. For example, e_1 is an increment in the internally generated peak voltage from one side to ground. R_0 is the incremental output impedance as viewed across either half of the collector winding, R_1 is the incremental input resistance seen in the base circuit (center tap of the base winding to ground).

Since a feedback-stabilized oscillator contains more than a single loop, we must be careful to state which loop is to be analyzed here. It is the composite loop which is opened by breaking the forward-transmitting

path of the oscillator. The loop-gain considerations apply only to incremental envelope changes about the usual operating conditions.

A. *Formulation of Approximate Loop Gain* ($R_r=0$)

The loop gain can be found by introducing an internal voltage change e_1, which produces a certain output voltage change e_0, equal to μe_1. This in turn produces a change in base voltage, $-ne_0$, which would, in turn, cause a change e_1', equal to βe_0 or $\mu\beta e_1$ under open loop conditions. The product $\mu\beta$ is the loop transmission but will be referred to as loop gain for the sake of brevity. The assumption of $R_r=0$ simplifies the analysis a good deal. The effect of finite R_r is analyzed in section E.

In response to a change e_1 we have, neglecting the oscillator feedback loading, a change

$$e_0 = \frac{R_1}{R_1 + 2n^2R_0}\, e_1 = \mu e_1. \tag{8}$$

The factor n^2 which multiplies R_0 results from the transformer ratio n, which has been introduced for the sake of generality; the factor 2 accounts for the impedance doubling which generally attends peak rectification of a sinusoid.[6] A change e_1 produces a change $-ne_0$ in the average base voltage. But the resulting internal voltage change, e_1', depends on the change in peak base voltage, as follows:

$$e_1' = ke_{b\ \text{peak}} \tag{9}$$

where k is a gain factor associated with the circuit. By definition, we have

$$e_{b\ \text{peak}} = e_b + \frac{1}{N}\, e_0[N = \text{coll./base turns ratio}]$$

$$= -ne_0 + \frac{1}{N}\, e_0 = -\left(n - \frac{1}{N}\right)e_0. \tag{10}$$

Substituting (10) into (9), we obtain

$$\frac{e_1'}{e_0} = -k\left(n - \frac{1}{N}\right) = \beta. \tag{11}$$

The loop gain is equal to the product of μ, given by (8), and β, given by (11). This product is

$$\mu\beta = \text{Loop Gain} = \frac{-R_1}{R_1 + 2n^2R_0}\, k\left(n - \frac{1}{N}\right). \tag{12}$$

The incremental output resistance (across either half of the collector winding) with the feedback loop closed is given by R_0 in parallel with $R_1/2n^2$ divided by $1-\mu\beta$. If we assume $1-\mu\beta \approx -\mu\beta$, we have

$$R_{\text{out}} \approx \frac{R_0R_1}{R_1 + 2n^2R_0}\, \frac{1}{-\mu\beta} = \frac{R_0}{k\left(n - \frac{1}{N}\right)}. \tag{13}$$

This factor of two obtains exactly only under certain idealized conditions. In the present case it can be shown to represent a fairly good approximation.

B. *Evaluation of R_0 and R_1*

In order to evaluate $\mu\beta$ and R_{out}, we must first determine R_0, k, and R_1. R_0 is simply the square of the rms collector voltage (10^2) divided by the total ac power (20×10^{-3} watts), i.e., $R_0=5K$. R_1 is the average incremental input resistance of the two transistors, in parallel with any external conductance, such as G (see Fig. 4). The input resistance of each transistor, during conduction, is $r_b+[r_e/(1-\alpha)]$. Typical values are: $r_b=400\,\omega$; r_e, with the collector current ranging from 0 to 5 ma will average $10\,\omega$, to which we must add the external $50\,\omega$, so that $r_e=60\,\omega$, and $\alpha=0.95$. Therefore $r_b+[r_e/(1-\alpha)]=1{,}600\,\omega$. This resistance is presented for just half the total time by one or the other transistor during its 90-degree conduction. The remainder of the time, the resistance is practically infinite. Capacitor C_1, however, spreads the $1{,}600\,\omega$ over the entire 360-degree period, so that its effective value is $R_1\approx3{,}200\,\omega$ which does not include the shunting external current supplier "G" (see Fig. 4). If G is arranged as a current source (using a resistor supplied from a fairly high voltage source such as the B-battery, or using a transistor, as shown in Fig. 6, then its incremental value is high enough to have negligible shunting effect. In this case we have $R_1\approx3K$. If, on the other hand, G is implemented as in Fig. 5, its incremental value may be about $3K$ under typical conditions. In that case we have $R_1\approx1.5K$. Therefore, R_1 may range from $1.5K$ to $3K$, under typical conditions. For higher-α transistors, of course, (e.g. $\alpha=0.99$) the upper limit of R_1 would be more than $10K$.

Knowing approximate values of R_0 and R_1, and letting $n=1$, as in the text, we can evaluate the potentiometer fraction μ, given by (8). For $R_0=5K$ and $R_1=2.5K$, we obtain $\mu=2.5/12.5=0.20$.

C. *Evaluation of k*

The gain factor k has been defined (9) as $e_1'/e_{b\ \text{peak}}$. This is equal to $dE_c/dE_{b\,\text{peak}}$ if the loop is opened, say, to the right of R_0 (see Fig. 8). We may begin by evaluating dE_b/dI_e, as follows:

$$\frac{dE_b}{dI_e} = \frac{1-\alpha}{\alpha}\, \frac{dE_b}{dI_b},$$

$$= \frac{1-\alpha}{\alpha}\left(r_b + \frac{r_e}{1-\alpha}\right) \approx (1-\alpha)r_b + r_e. \tag{14}$$

At the peak current of 5 ma, we have

$$r_e = 5 + 50 = 55\,\omega$$

$$(1-\alpha)r_b = (0.05)(400) = 20\,\omega,$$

so that

$$\frac{dE_{b\ \text{peak}}}{aI_{e\ \text{peak}}} \approx 75\,\omega.$$

We must remember that I_e as used above is the peak collector current (5 ma). The quantity

$$\frac{d\tilde{E}_c}{dI_{c\ \text{peak}}}$$

is therefore not R_0, but simply the voltage swing (14 volts) divided by the peak current (5 ma), i.e., $2.8K$.[7]

$$k = \frac{d\tilde{E}_c}{dE_{b\ \text{peak}}} = \frac{d\tilde{E}_c}{dI_{c\ \text{peak}}} \frac{dI_{c\ \text{peak}}}{dE_{b\ \text{peak}}} = \frac{2800}{75} = 37. \quad (15)$$

Experimentally, k has been measured for a variety of transistors, and found to center in the lower thirties. It should be noted that removal of the 50-ohm emitter resistor which adds to r_e would roughly triple the value of k (and reduce R_1 by 30 to 60 per cent).

D. *Approximate Values of Loop Gain and Closed-Loop Output Impedance*

From section B, we have

$$\mu = \frac{R_1}{R_1 + 2R_0} = 0.2,$$

and from section C we have $k = 37$. Therefore we have

$$-\beta = k\left(n - \frac{1}{N}\right) = (0.9)(37) = 33,$$

for $n = 1$, $N = 10$. The loop gain (12) is therefore

$$\mu\beta \approx (0.2)(33) = 6.6.$$

The closed-loop output impedance (13) is

$$R_{\text{out}} \approx \frac{R_0}{\left(n - \frac{1}{N}\right)(k)} = \frac{5{,}000}{33} = 150\ \omega.$$

Some facts of general interest can be deduced from the preceding results. In the derivation of loop gain (Section A), the factors μ and β were chosen in such a way that, under closed-loop conditions, we have

$$\frac{e_0}{e_1} = \frac{\mu}{1 - \mu\beta} \approx \frac{1}{\beta} = \frac{1}{k\left(n - \frac{1}{N}\right)}. \quad (16)$$

This shows, for example, that a change (such as a noise disturbance) e_1 produces a smaller change e_0 at the output if we increase k and also the turns ratio n. This also has the effect of reducing R_{out}, as shown by (13). These conclusions hold only so long as $|\mu\beta|$ is at least somewhat above unity. But (8) shows that, beyond a certain point, raising n (while holding R_0 constant) drops μ faster than it raises β, so that $\mu\beta$ decreases. In brief, there are certain optimum relations between the operating parameters, k, n, R_1, R_0, etc., which lead to minimum noise and output impedance.

[7] An assumption implicit in this derivation is that there is no first order change in conduction angle. This can be shown to be a valid approximation: changes in conduction angle are an order of magnitude (percentagewise) below imposed voltage or current changes.

E. *Effect of Regulator Resistance*

The results obtained above are approximate in that they are based on a zero-impedance source of reference voltage, i.e., $R_r = 0$ and hence $e_r = 0$. We shall now repeat the derivation of Section A without neglecting R_r.

In response to a change e_1, we have

$$e_0 = \frac{R_1 + R_r}{R_1 + R_r + 2n^2R_0} e_1, \quad (17)$$

and

$$e_b = \frac{-nR_1}{R_1 + R_r + 2n^2R_0} e_1. \quad (18)$$

Also, $e_{b\ \text{peak}} = e_b + (1/N)e_0$, (N = coll./base turns ratio)

$$= -\frac{nR_1 - \frac{1}{N}(R_1 + R_r)}{R_1 + R_r + 2n^2R_0} e_1 \quad (19)$$

and

$$e_1' = ke_{b\ \text{peak}}, \text{ as before (9).}$$

Combining (9) and (17), we obtain the following expression for loop gain (e_1'/e_1):

$$\text{Loop Gain} = \frac{-R_1}{R_1 + R_r + 2n^2R_0} k\left[n - \frac{1}{N}\left(1 + \frac{R_s}{R_1}\right)\right]. \quad (20)$$

The incremental output resistance is approximately equal to R_0 in parallel with $(R_1 + R_s)/n^2$ divided by the magnitude of the loop gain.

$$R_{\text{out}} \approx \frac{R_0(R_1 + R_r)}{R_1 + R_r + 2n^2R_0} \frac{1}{-\mu\beta} = \frac{R_0\left[1 + \frac{R_r}{R_1}\right]}{k\left[n - \frac{1}{N}\left(1 + \frac{R_r}{R_1}\right)\right]}. \quad (21)$$

By comparing (20) and (21) to (12) and (13), respectively, we can readily see the manner in which R_r lowers the loop gain and raises the output impedance.

Assuming the same numerical values as before, i.e., $R_0 = 5K$, $R_1 = 2.5K$, $n = 1$, $N = 10$, $k = 37$, and in addition assuming $R_r = 1K$[8], we obtain from (20):

$$\text{Loop Gain} = \frac{2.5}{13.5} 37[0.86] = 5.9.$$

From (21),

$$R_{\text{out}} \approx \frac{5K[1.4]}{37[0.86]} \approx 220\ \omega.$$

[8] R_s includes not only the incremental resistance of the regulator diode and the compensating and adjustment resistances in series with it (see Fig. 3), but also the effective incremental resistance of the diodes (diode D_0 in Fig. 8) which rectify the oscillator voltage and compare it to the regulator voltage. With modern junction diodes this effective resistance can be as low as 500 ohms at average currents of 40 to 100 microamperes.

Experimentally measured values of R_{out} center around 250 ohms. The presence of R_r has dropped the loop gain by about ten per cent but has raised R_{out} by nearly fifty per cent.

F. *Frequency Characteristic of Loop Gain*

All of the preceding results in this appendix pertain to the steady-state or zero-frequency behavior of the amplitude control system. If we desire to ensure the stability of this system we must inquire into the loop gain over a finite frequency range.

As we depart from the operating frequency of the oscillator by an amount ω rad/sec the rectified control signal (after diode D_0, Fig. 8) departs from zero frequency by ω rad/sec. Under this condition the loop gain is modified by the finite bandwidth $[(1/R_0C_0)$ rad/sec$]$ of the output circuit, by the smoothing capacitor C_1 shunting the input resistance R_1, and to a smaller extent by the capacitor C_r shunting the regulator resistance R_r. By applying a frequency shift equal to the operating frequency we can simulate the tuned circuit by $2n^2R_0$ in parallel with C_0/n^2. In order to determine the relative loop gain we may now determine e_b/e_1 for the equivalent network shown in Fig. 9, where typical values have been indicated. The value of n is taken as unity.

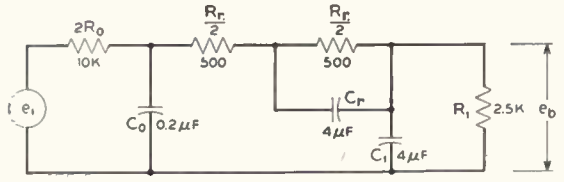

Fig. 9—Network for computing frequency dependence of loop gain.

The regulator resistance has been divided into two halves. One half is associated with the rectifier diodes[8] and is therefore not capacitor by-passed. The transmission of the above network can be approximated as follows: At (radian) frequencies below $2/R_rC_0$ or 10,000 rad/sec, C_0 robs the other branches of relatively very little current and can therefore be neglected. The effects of R and C_r may be taken into account by adding R_r to $2R_0$ at frequencies below $2/R_rC_r = 500$ rad/sec. Above 500 rad/sec only $R_r/2$ remains between C_0 and C_1.

From $\omega = 0$ to $\omega = 500$, we have, for the loop gain (relative to zero-frequency loop gain)

$$\frac{\text{Loop Gain } (\omega)}{\text{Loop Gain } (0)} = \frac{1}{1 + j\dfrac{\omega}{\omega_0}}, \tag{22}$$

where $\omega_0 = 1/RC_1$,

and

$$R = R_1 \text{ in par. with } 2R_0 + R_r$$
$$\approx 2K,$$

and $\omega_0 \approx 10^6/(2000)(4) = 125$ rad/sec.

From 0 to just below 500 rad/sec the relative loop gain is therefore

$$\frac{1}{1 + j\dfrac{\omega}{125}},$$

which means a flat response from 0 to near 125 rad/sec, followed by a 6-db per octave drop above 125 rad/sec. Around 500 rad/sec, as R_r is replaced effectively by $R_r/2$, the 6-db per octave slope briefly diminishes, but the resulting "gain step" amounts to less than one-half db. Near $\omega = 2/R_rC_0$ or 10,000 rad/sec an additional slope of 6 db per octave begins, making the total slope 12 db per octave. Assuming the zero-frequency loop gain 5.9 (15.4 db), as found in Section E, we can sketch the loop gain frequency characteristic roughly as in Fig. 10.

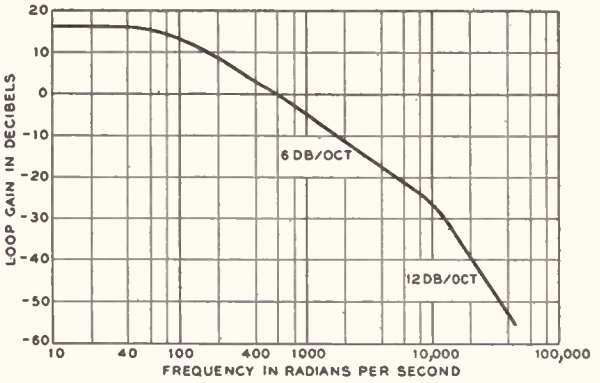

Fig. 10—Incremental envelope loop gain vs. radian frequency. The gain plotted is that of the composite loop which is opened by breaking the forward-transmitting path of the oscillator. (See Fig. 8.)

The fact that the gain drop-off increases from 6 to 12 db/octave at -25 db guarantees that both gain and phase margin are more than sufficient to ensure stability of the control system.

Control of Vertical Radiation Patterns of TV Transmitting Antennas*

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Summary—**This paper deals with the application of the principles presently employed in the design of directional antennas for the Standard Broadcast band to the control of radiation from TV transmitting antennas in order that optimum use is made of the energy radiated from such antennas.**

Until very recently TV antennas were designed primarily to concentrate the radiated energy in the direction of the horizontal plane and the greater this concentration the more desirable the antenna. During the past year the engineering profession has realized the importance of determining the nature of the area to be served by a given antenna system and then designing the antenna to cover this area most effectively with the power available. The need for such design is shown by several examples of problems already solved by this method and the general method of attack on the problem is developed in some detail.

PART I

WITH THE ADVENT of nation-wide commercial television operation during the mid-forties and its immediate acceptance by the public, it became the problem of the prospective television broadcaster to get his station on the air as rapidly as possible and with the equipment then available. The industry was prepared for this and produced the necessary quantity of television transmitters, antennas, and studio equipment to meet this demand. By the time of the so-called "freeze," equipment had been standardized and in particular, the antennas in use were limited to those having from three vertical bays up to a total of six vertical bays. The power output of the transmitters was limited but there was likewise a limit on the effective radiated power and an even greater limit was imposed when the height of the antenna exceeded 500 feet. Because of these limiting factors, the service provided by the available equipment and antenna system was generally satisfactory and in the cases where television coverage appeared inadequate it was generally ascribed to the transmission characteristics of the very high frequencies.

Although practically all of the 100-odd stations operating at the time of the "freeze" had antenna heights of 500 feet or less, there were a few exceptions, notably those on Mount Wilson, near Los Angeles, and in particular, KECA-TV, the Los Angeles operation of the American Broadcasting Company. Operation from this point gave what was probably the first definite indication of the need for control of the radiation pattern of the transmitting antenna. The studies undertaken during the "freeze" on the need for correcting the Mount Wilson situation laid the groundwork for the solution of the problems presented at the termination of the "freeze" when increases in allowed power and increases in antenna height emphasized the necessity for controlled radiation.

The first television antennas were designed to augment the effective power of the relatively low-powered transmitters then available by concentrating the energy along the horizontal plane. This of course was only accomplished by reducing the energy radiated in other directions but with the limited gains then contemplated and the limited antenna height of 500 feet or less, this introduced no problem.

Fig. 1—Vertical plane field distribution typical 3-bay superturnstile.

Fig. 1 shows a portion of the vertical directivity pattern of a typical three-bay superturnstile antenna. It will be noted that from a maximum along the horizontal plane the field tapers off gradually until it reaches the value of ten per cent of the maximum field at some 15° below the horizon. The field does not drop off to a null until an angle of 17° below the horizontal plane is reached. If we refer now to Fig. 2 we can interpret the effect of this drop-off in field in terms of the distance from the transmitter at which serious effects may be noted. It will be seen that for an antenna height of 500 feet the field does not decrease to the ten per cent value until one is within 0.3 mile of the transmitter. For most transmitter locations this would not create a noticeable

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area of low signal. However, if we increase the antenna height to 1,000 feet the drop-out occurs at 0.6 mile and for 2,000 feet it is 1.2 miles from the transmitter. At these distances the area of low signal is more likely to lie in populated districts.

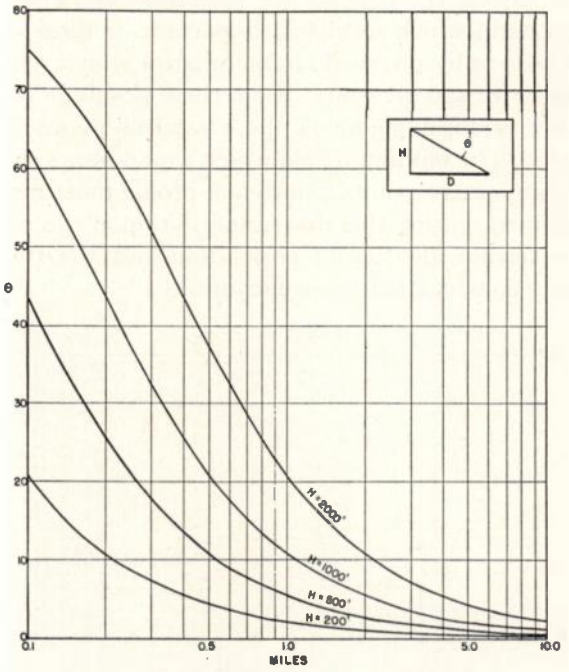

Fig. 2—Depression angle vs. distance for various antenna heights.

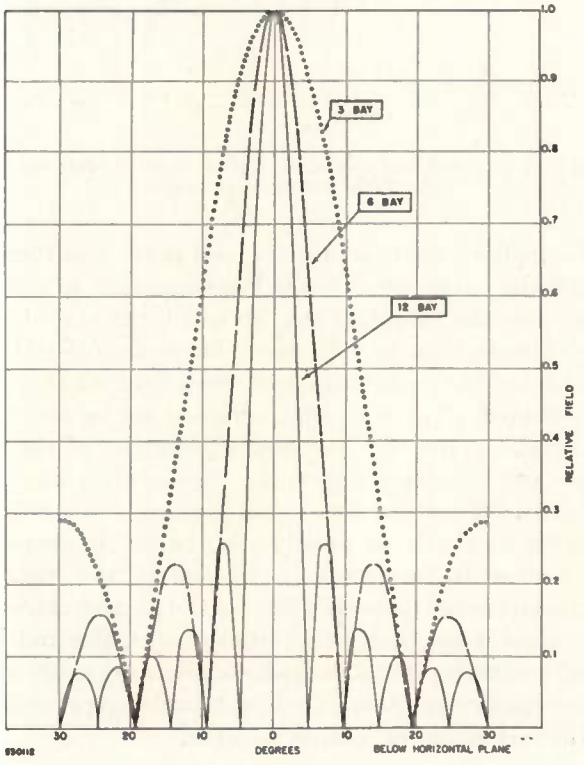

Fig. 3—Vertical plane field distribution typical 3-6- and 12-bay superturnstiles.

If now we wish to obtain more effective radiated power by employing an antenna of greater gain, we might use a 6-bay or even a 12-bay antenna. This would provide quite high values of effective radiated power with relatively low investment so far as transmitter costs are concerned. However, if we refer to Fig. 3 which shows the vertical pattern of the 6-bay antenna and of the 12-bay antenna superimposed on that of the 3-bay antenna, it will be seen that the rapid decrease in field as we go below the horizontal plane might well introduce serious problems, particularly at antenna heights of 1,000 feet or more. The 12-bay antenna, for example, has three complete nulls, occurring at roughly 5°, 9°, and 14°, before reaching the first null of the 3-bay.

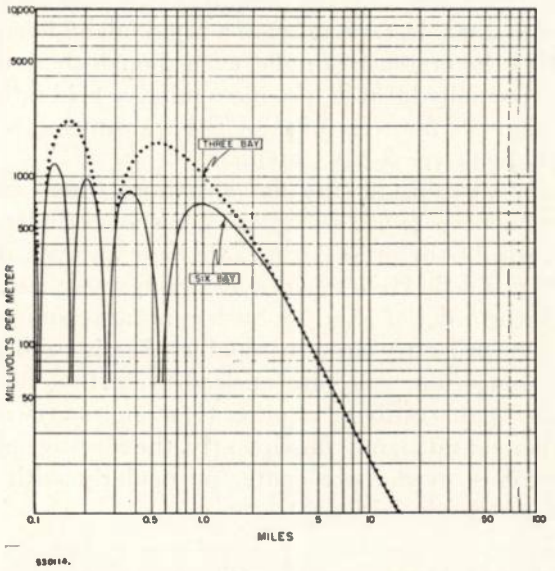

Fig. 4—Calculated field intensities typical 3- and 6-bay antennas for 500 feet effective height.

Fig. 4 is a plot of the calculated field intensity against distance for a 3-bay antenna and for a 6-bay antenna with an effective radiated power of 100 kw in the frequency range 54–88 mc, and an antenna height of 500 feet. The effect of the nulls in the pattern is very apparent; however, they are still very close to the transmitter and in many cases would be unnoticed. If we refer to a situation more typical of present planning we should consider Fig. 5, which shows the calculated field intensity versus distance with 100 kw effective radiated power (i.e., along the horizontal plane), and with an antenna height of 1,000 feet for the 3-bay antenna and for the 12-bay antenna. It should be noted particularly that the major limitation of the 12-bay antenna is not so much the existence of the additional nulls but rather is directly related to the rapid decrease in radiated field at very small angles below the horizontal plane. Since complete nulls are rarely present in practice due to slight inaccuracies in tuning and adjusting the antenna, it has frequently been maintained that there will be no adverse effect on coverage if the 12-bay antenna were employed rather than, for example, the 3-bay. Inspection of Fig. 5 reveals the fallacy of this argument. Even if we were to assume the null to be filled in, to a value equal to the peaks of the lobes, the field from the 12-bay antenna remains substantially constant from a point five miles away from the transmitter to within a half mile of the

transmitter, whereas that from the 3-bay antenna continues to rise (under the same assumptions as to null fill-in) in, to the transmitter site. From a competitive viewpoint this is an important factor which cannot be overlooked. If a ratio of signal intensities between two otherwise comparable transmissions is of the order of 5 to 1, viewer preference will invariably select the strong signal. When, in addition, the field drops out in the presence of the nulls to the extent normally found in practice, this ratio may exceed 20 to 1.

Fig. 6 shows a similar comparison to that of Fig. 5 with the exception that the antenna height has been increased to 2,000 feet. Here again a rapid fall off in signal as well as the presence of nulls, even though those nulls may be partially filled in, produces relative field intensities within the first ten miles of the transmitter which strongly favor the 3-bay antenna.

If now we consider the uhf situation, where available transmitter power is relatively low and admittedly inadequate at the present time, we have an even more difficult problem. Here we are considering antennas with gains of the order of 15 to 25. Such gains are required at present in order to meet reasonable effective radiated power values, but they are misleading in that no appreciable field is radiated at other than angles very close to the horizontal plane; consequently, the coverage problem becomes even more acute, particularly with antennas of substantial height.

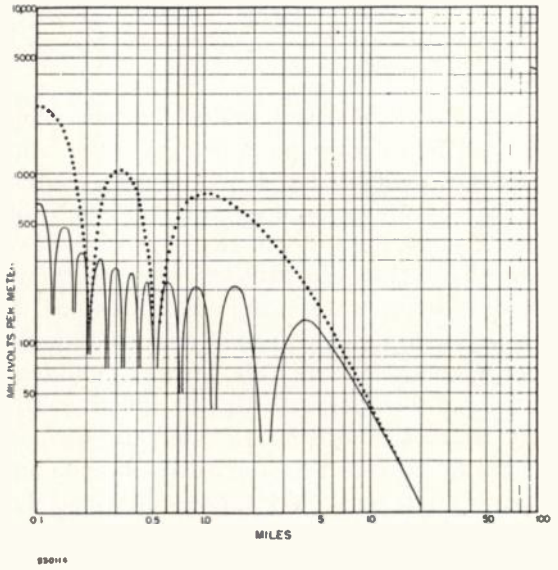

Fig. 5—Calculated field intensities typical 3- and 12-bay antennas for 1,000 feet effective height.

Fortunately, we have at hand a means for correcting this situation; that is, the application of directional antenna design such as is commonly used in standard broadcasting. By application of the same formulas and principles presently used to compute directional antennas where the directivity is in the horizontal plane, we may adjust the directivity in the vertical plane until we arrive at an optimum distribution of such radiation. It is, of course, impossible with a physically realizable structure, to abstract all of the energy ordinarily radiated above the horizon and direct it below; however, it is possible to change the shape of the directive pattern materially and approximate this condition.

The procedure to be followed in order to arrive at a suitable antenna configuration is similar to that employed in determining upon a suitable pattern for standard broadcasting. Profiles are prepared in the various directions in which service is important or those directions possessing physical characteristics which indicate necessity for special study. From these profiles a plot of distance versus angle below the horizontal plane can be prepared. The minimum field which would be acceptable at the important points along each profile must next be determined, basing this determination upon the nature of the terrain, density of population and the type of building construction to be encountered.

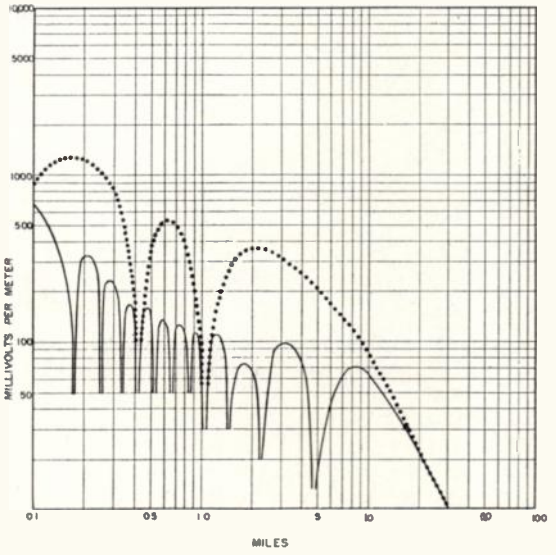

Fig. 6—Calculated field intensities typical 3- and 12-bay antennas for 2,000 feet effective height.

Particular attention must be paid to the point beyond which the coverage becomes less important in order to determine the extent, if any, to which the main lobe of the antenna may be depressed. These data permit calculation of the desired unattenuated field which should be radiated at each of the pertinent angles below the horizontal plane. By suitable adjustment of the field ratios and phasing between the various elements of the antenna, the vertical directional pattern is then made to conform as nearly as possible to the values previously determined to be necessary. Obviously, the more elements employed, the more flexible the pattern; however, in general it has been found from weight, size and complexity considerations that six elements are satisfactory under many conditions, and that twelve represents for vhf a reasonable maximum number.

Part II

Let us consider, for example, a 6-bay antenna as modified for use by KECA-TV, of the American Broadcasting Company, on Mount Wilson. The standard 6-bay antenna divides the power equally above and below the horizontal plane and has its first null 9.5° below.

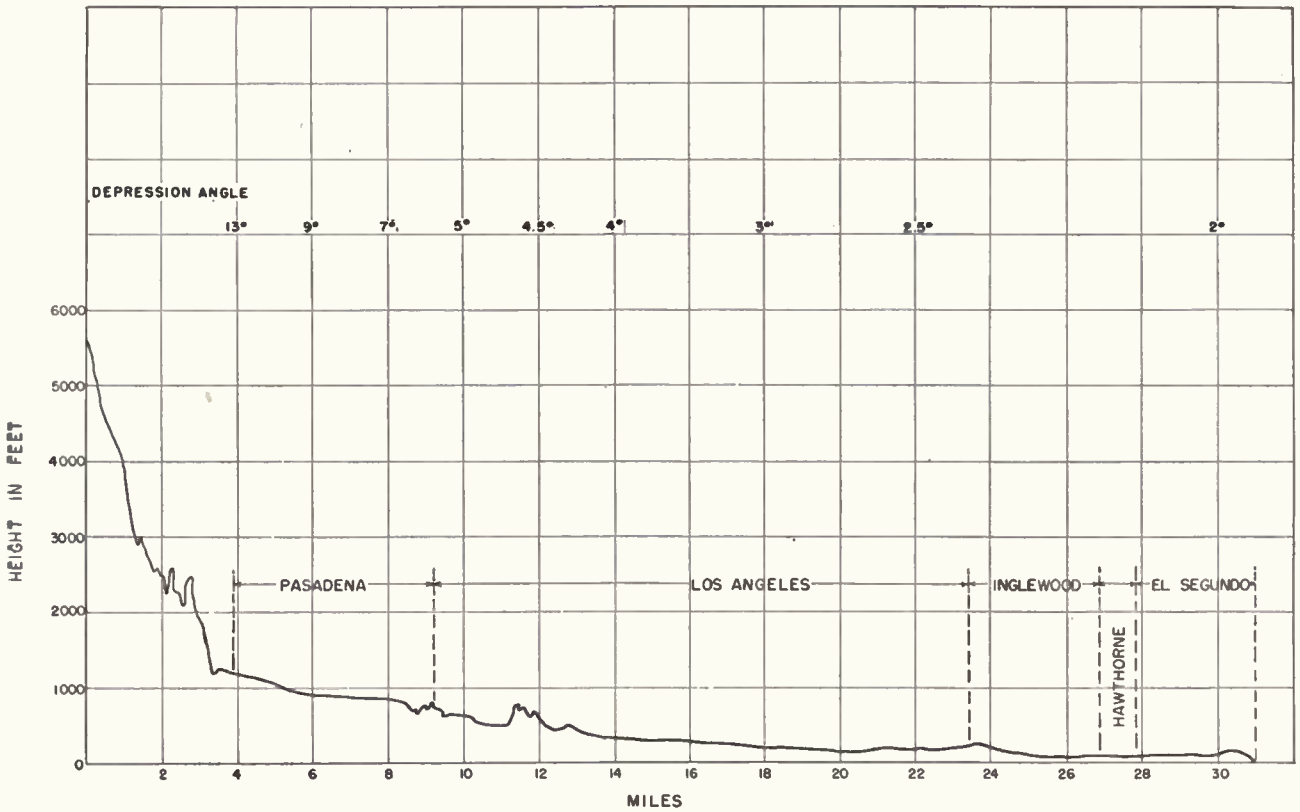

Fig. 7—Profile from Mount Wilson—azimuth 220° true.

As located on Mount Wilson this null coincides with the location of the city of Pasadena, an important service area. Furthermore, the extreme limit of service to the west is at the land's end, some 31 miles from Mount Wilson, and this is 2° below the point of maximum radiation of the standard 6-bay antenna. Fig. 7 is a typical profile from Mount Wilson with the areas to be served and depression angles marked thereon.

By various devious means we determine upon the field ratios and phase relations between the elements of a modified 6-bay antenna for Mount Wilson listed in Table I.

TABLE I

MODIFIED 6-BAY SUPERTURNSTILE ANTENNA

Bay	Field Ratio	Phasing-Degrees
1 (Bottom)	1	−35
2	1	− 5
3	2	0
4	2	+30
5	1	+35
6	1	+65

With this configuration the antenna produces a pattern shown in the solid line in Fig. 8. Here the dotted line represents the conventional 3-bay antenna with unity field ratios and excited in phase. Inspection of the two curves discloses the fact that the first null of the modified 6-bay antenna now occurs at approximately the same point as that of the 3-bay antenna and furthermore, that the peak of radiation from the array occurs at 2.5° below the horizontal plane, so that a maximum amount of energy is directed over the service area desired and the region of very low fields in the Pasadena area has been eliminated. Fig. 9 shows the computed fields westward from Mount Wilson for the standard 6-bay antenna and the modified antenna. The degree of improvement is apparent. It should also be noted that with the antenna modified in this fashion the power radiated below the horizontal plane is 2.1 times the power

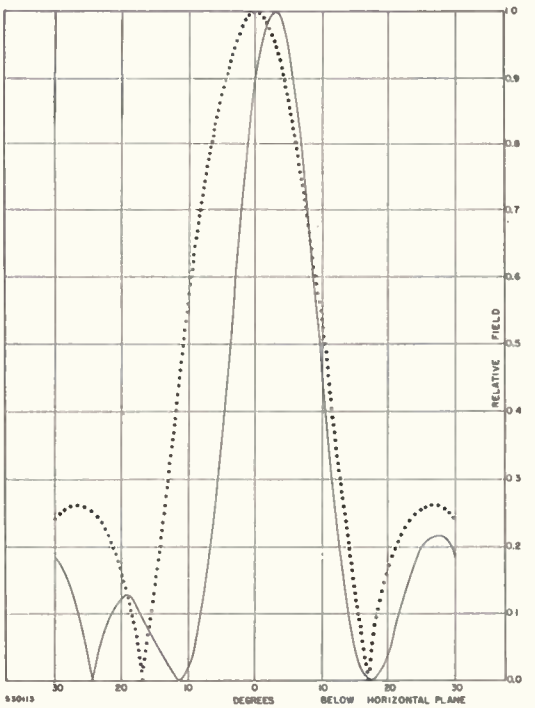

Fig. 8—Vertical plane field distribution standard 3-bay and modified 6-bay antennas, KECA-TV.

radiated above the horizon which is a marked improvement in efficiency.

The redistribution of energy results in decreasing the gain of the 6-bay antenna structure to approximately 80 per cent of the gain figure for the in-phase antenna; however, it is adequate to permit the use of a reasonably powered transmitter and still obtain the maximum effective radiated power permitted by the FCC.

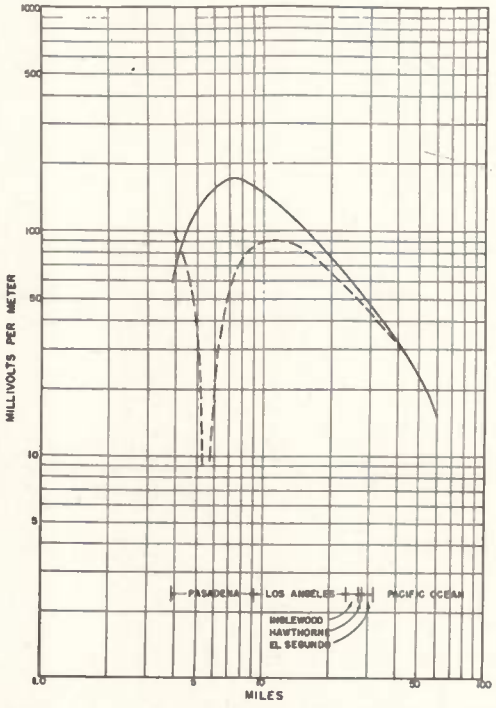

Fig. 9—Calculated field intensities for KECA-TV on Mount Wilson, with standard 6-bay and modified 6-bay antennas.

A different type of problem is encountered when we consider operation in the high frequency portion of the vhf band with antenna heights lower than the value at which a reduction in power is required. In this case it is desirable to maintain a figure for the over-all gain which will permit achieving an effective radiated power of 316 kw with a transmitter power of 50 kw. Taking into account transmission line losses, harness losses, etc., this means that a pattern gain of at least 7.5 must be maintained. A 12-bay antenna was selected as possessing sufficient elements to have a reasonable degree of flexibility and at the same time having a gain in the in-phase condition high enough so that the rephasing and redistribution of element power would not reduce it below the 7.5 figure.

Let us consider a specific case. It is proposed to operate KGO-TV, the American Broadcasting Company station in San Francisco, California, with an effective radiated power of 316 kw and to locate the antenna on a 700-foot tower which is in turn located on top of Mount Sutro. This gives an effective height for the antenna of 1,550 feet. Profiles were drawn in the various pertinent directions from the Mount Sutro site and the unattenuated field required to establish a minimum field of 150-mv/m along each profile to the point where attenuation takes control was determined. This value of 150 mv/m was selected as representing a value high enough to provide a good service signal throughout the heavily built-up areas of San Francisco and Oakland.

The equations for the pattern of a twelve-section antenna were then adjusted until the required unattenuated field was obtained. The following Table represents the combination finally arrived at:

TABLE II

MODIFIED 12-BAY SUPERTURNSTILE ANTENNA

Bay	Field Ratio	Phasing-Degrees
1 (Bottom)	1	0
2	1	0
3	1	0
4	1	0
5	2.4	−30
6	2.4	−30
7	2.4	+30
8	2.4	+30
9	1	0
10	1	0
11	1	0
12	1	0

Fig. 10 shows a portion of the vertical plane pattern of a standard 12-bay in-phase antenna and the modified pattern selected for Mount Sutro. The gain of the antenna thus modified is approximately 7.6, which will permit an effective radiated power of 316 kw with a 50-kw transmitter. The field produced by these two antennas was calculated and is shown on Fig. 11. It will be noted that the field does not drop below the 150 mv/m value until one is within 0.5 mile of the antenna, whereas the standard antenna shows numerous areas in which the field would be far below this value.

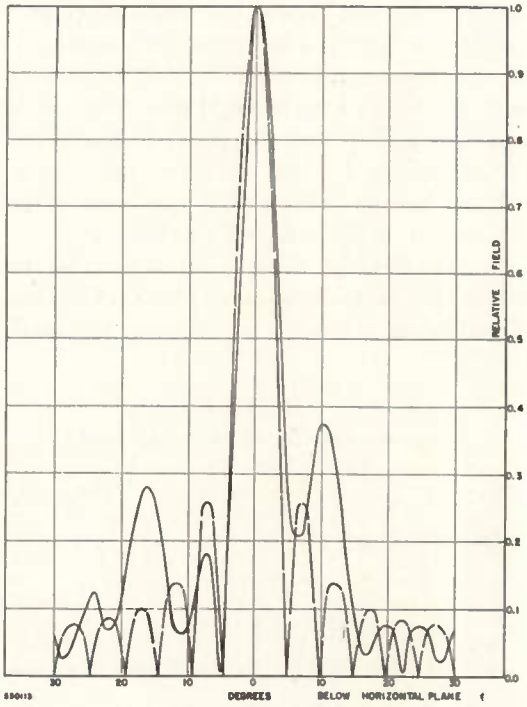

Fig. 10—Vertical plane field distribution standard 12-bay and modified 12-bay antennas, KGO-TV.

When we examine the problem of uhf coverage the factor of total available power assumes an important

position. Although the FCC authorizes effective radiated powers of up to 1,000 kw for uhf stations, present-day transmitters will not permit us even to approach this figure without absurdly high antenna gain figures. The problem presently is to find the best possible distribution for a total available power of 1,000 w to 12 kw in order that uhf stations going on the air in the immediate future will establish a field intensity coverage pattern which will permit them to approach competition with vh .

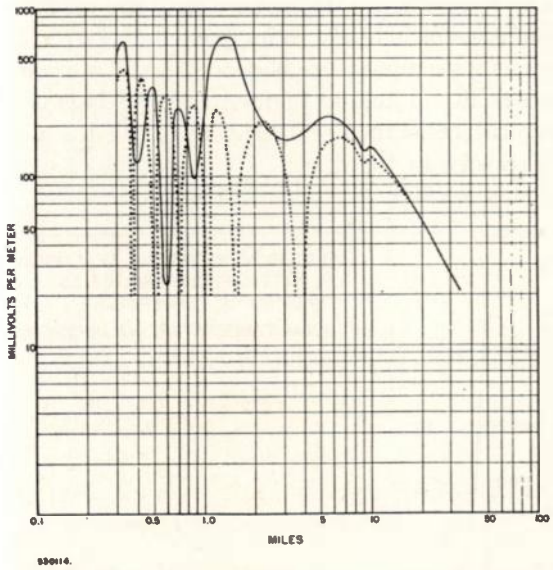

Fig. 11—Calculated field intensities for KGO-TV on Mount Sutro, with standard 12-bay and modified 12-bay antennas.

As an example of one such problem, we have WAKR-TV in Akron, Ohio. Here the transmitter is located on a rooftop in the center of the business district. This is desirable in that all of the high values of field intensity cover areas which need these values but it imposes an additional problem in that the presence of nulls or sharp decreases in field close in to the antenna would make the coverage ineffectual.

Several companies collaborated in devising a uhf antenna which would provide service for WAKR-TV. The one finally selected was a TFU-24-BM (slot-type radiator). This antenna is made up of sixteen sections and by dividing the power unequally between the upper ten and the lower six sections and suitably phasing the driving system, a directional pattern was achieved as shown in Fig. 12. It will be noted that the main beam is depressed one degree below the horizontal plane and the nulls are filled in down to at least fourteen degrees. The gain of this antenna computes to be 79 per cent of the standard untilted antenna; that is to say, 79 per cent of 24 or 19.

The computed field intensity[1] produced by this antenna with an effective height of approximately 350 feet and a total power input to the antenna of 1,000 w, is shown in Fig. 13. The close-in field out to the point of noticeable attenuation is maintained very close to 100 mv/m. While this is still a marginal field in a competitive market, it represents what is believed to be the best distribution of the available energy. It it the more acceptable since there is reason to believe that transmitter powers of 10 to 12 kw will be available within the next twelve months, which if used in connection with this antenna would increase the close-in field to an average of 300 mv/m. This compares favorably with the value of 100 to 200 mv/m used as a design minimum in one of our vhf illustrations.

[1] Where calculated or predicted field intensities are shown in this paper, the method employed in calculation was that prescribed by the FCC in its "Rules Governing Television Stations."

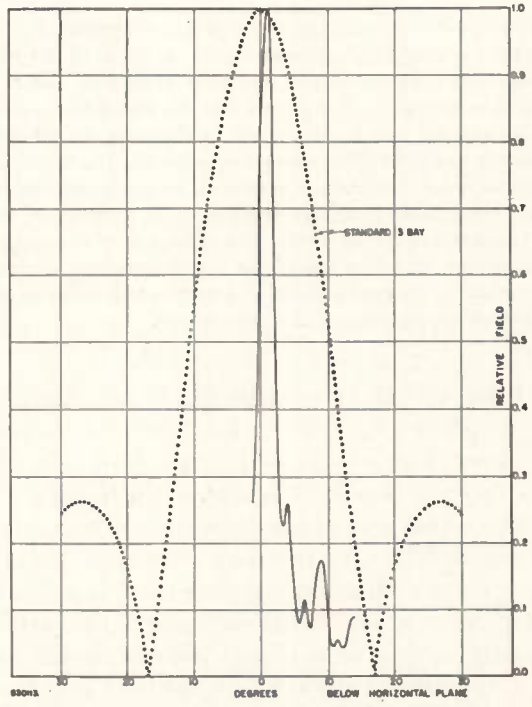

Fig. 12—Vertical plane field distribution standard 3-bay and modified 24-bay antennas, WAKR-TV.

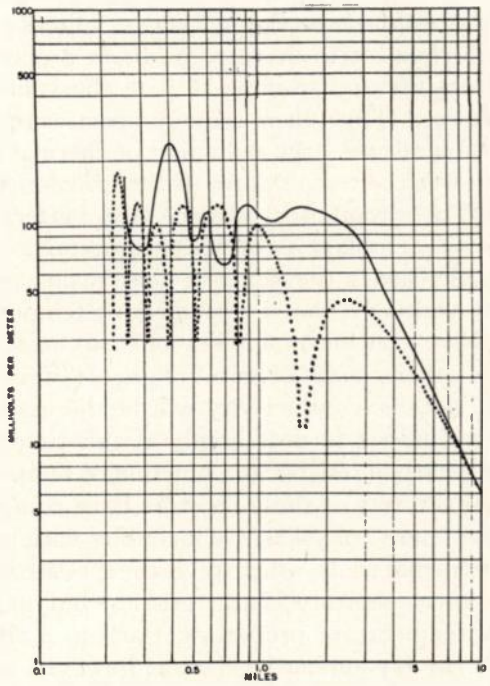

Fig. 13—Calculated field intensities for WAKR-TV with standard uhf and modified uhf antennas.

Low-Temperature Electronics*

CLAYTON A. SWENSON† AND ALFRED G. EMSLIE‡

Summary—**Considering potential practical applications of low-temperature phenomena, a general survey is made of the temperature dependence of noise and electrical resistance down to the region of Absolute Zero. It is shown that the signal-to-noise ratio in a metallic resistor can be increased by lowering its temperature, while cooling has little effect on carbon resistors. The theory of electrical conductance in metals is reviewed, and it is determined that whenever ohmic losses are important as in an inductance, decreasing the temperature can be a help. The phenomenon of superconductivity is treated, including use of the superconducting properties of certain metals in the construction of a very sensitive bolometer and in the frictionless suspension of a lead sphere.**

A GREAT DEAL of work has been done in recent years in the field of low-temperature physics, most of which can be found only in publications such as *Physical Review*, *Physica*, or *Philosophical Magazine*. The techniques of low-temperature research have developed sufficiently, however, (using commercially available helium liquefiers) so that there is an excellent opportunity for practical use for some of the results.

Among the items of potential interest to electronics' workers, the subject about which the least experimental data are available is that concerning the temperature dependence of electrical fluctuations or noise. The ordinary noise found in metallic resistors, commonly referred to as Johnson or thermal noise, follows a simple law which shows it to decrease in direct proportion to the temperature and to increase with the value of the resistance used. Thus, there is an immediate application of low temperatures—the reduction of thermal noise—and the lower the temperature of the grid resistor in a photo-electric circuit, for instance, the better. Where the noise in a resistor is of great importance, a considerable increase in the signal-to-noise ratio can be expected by making it metal and lowering temperature.

The above conclusion applies only to metallic resistors, for other sources of noise are important in carbon resistors. Metallic conductivity will be discussed later, but it is of interest to note briefly at this time the behavior of a carbon resistor as a function of temperature. The manufacturers of these resistors have complicated life for the "pure" physicist by including various types of unknown materials which give their resistors optimum room-temperature characteristics, but unpredictable low-temperature properties. Carbon, a semiconductor, is the predominating factor, however, and the properties of the resistors resemble those of a normal semiconductor.

Noise in semiconductors (and in carbon resistors) is found to be current and frequency dependent, and to behave much like shot noise in a vacuum tube.[1,2] Since it is practically independent of temperature, there is no advantage in cooling a carbon resistor. In fact, there is some disadvantage, since the resistance of semiconductors is so temperature dependent that the resistance of a carbon resistor is likely to increase by a factor of fifty or more as its temperature is decreased from room temperature to liquid helium temperatures.[3,4] This is shown in a typical resistance-temperature curve for a ½ watt carbon resistor in Fig. 1.

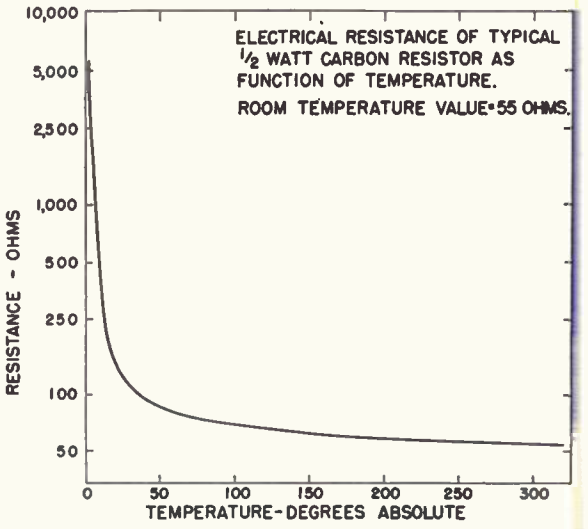

Fig. 1—The electrical resistance of a carbon resistor as a function of temperature.

These two factors, temperature-sensitive resistance and temperature-independent noise, would seem to lead to the conclusion that the transistor (made from a semiconductor, germanium) will have little to offer in the way of reduced noise as it is cooled. In fact, there seems to be little that can be done to overcome the shot noise difficulties inherent in vacuum tubes through the use of devices at low temperatures. The thermionic cathode could be replaced with a photo-electric cathode, or one using field emission, but the emission of electrons is still fundamentally a random process and leads to the shot effect type of noise. With modern refrigeration tech-

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[1] H. C. Montgomery, "Noise in semiconductors," *Bell Sys. Tech. Jour.*, vol. 31, p. 950; Sept., 1952.

[2] D. K. C. MacDonald and I. M. Templeton, "Fluctuations," *Proc. International Conference on Low Temperature Physics*, Oxford, p. 20; Aug., 1951.

[3] A. Brown, M. W. Zemansky, and H. A. Boorse, "A sensitive and reproducible thermometer in the range of 2° to 20°K," letter in *Phys. Rev.*, vol. 84, p. 1050; Dec. 1, 1951.

[4] J. R. Clement and E. H. Quinnell, "Characteristics of carbon-composition thermometers at low temperatures," *Proc. International Conference on Low Temperature Physics*, Oxford, p. 51; Aug., 1951.

niques, it is possible to operate hearing aid tubes, and their whole circuits, at temperatures approaching absolute zero, but unless this is done to reduce noise in a metallic grid resistor, or to provide extreme temperature stability, there would seem to be little value in this type of application.

A subject of greater potential interest is that of electrical conduction in metals. Ohmic losses in pure resistances cannot be altered, since there is nothing that can be done to reduce the heat developed when a current flows through a given resistance. The production of low loss reactive components for moderate frequencies is possible at low temperatures, however, and it is with respect to ideally nonresistive circuit components that the temperature dependence of the electrical resistance will be discussed.

First, some discussion about the theory should be given.[5] The distinguishing feature of a metal is not so much the strong inter-atomic forces which hold its atoms together, as the fact that each of these atoms is likely to lose control over one or more of its electrons. These electrons then belong to no single atom, and much like the atoms in a gas, they are free to wander about inside the metal with certain restrictions.

It is the guided wandering of these electrons that causes a net current to flow when an electric field (a potential difference) is put across the metal. Delving into the theory, it is found that electrons cannot lose energy by collisons with atoms which are arranged in a perfectly regular lattice. A first guess then would predict normally infinite conductivity.

However, lattices are not perfect in practice and are usually distorted in two ways. First, the lattice is being continually distorted by the thermal vibrations of the atoms, and these vibrations are able to interact with the electrons and to remove energy from them. This effect should go to zero as absolute zero is approached and the ideal metal should approach zero resistance as its temperature is lowered. In fact, the prediction can be made, in agreement with experiment, that the temperature dependence of the electrical resistance of pure metallic elements should be linear at high (room) temperatures, and should fall off as some higher power of the temperature (actually, T^5) near absolute zero. This is observed for a metal like sodium, which is rather simple and should obey the theory. It is roughly true for many other pure metals.

In practice, the electrical resistance does not generally extrapolate to zero at absolute zero, but approaches a small and finite value called the "residual resistance." This residual resistance is believed due to actual imperfections in the crystalline lattice, places where the regularity of the lattice has been destroyed (by crystal faults, grain boundaries, etc.), or due to actual impurity atoms present in the metal. These contributions should be temperature independent and should be a measure of the purity of the metal that is being used. The residual resistance is usually subtracted from the actual resistance to obtain the true temperature dependent part of the resistance. When this is done, the general conclusions given above seem to hold—namely, a linear temperature dependence at room temperature and a decrease with a higher power of the temperature at lower temperatures.

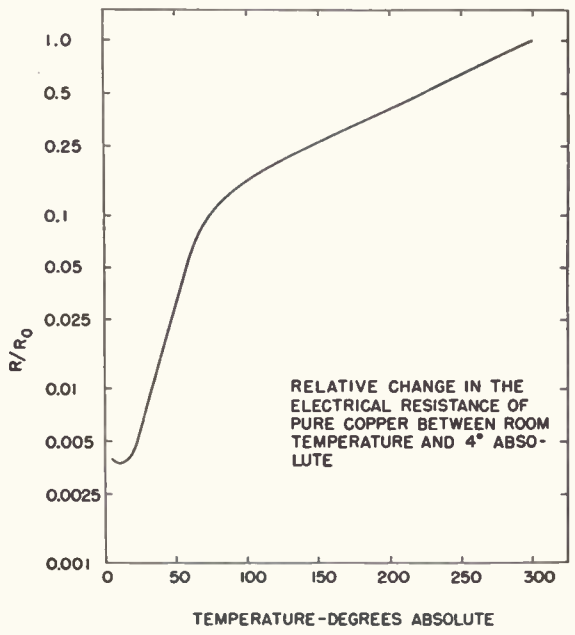

Fig. 2—The relative electrical resistance of pure copper as a function of temperature.

It is perhaps of interest to consider copper as a specific case of technical interest. A typical resistance-temperature curve for copper is shown in Fig. 2 in terms of the fraction of the room temperature resistance.[6] The upper part of the curve is much the same for most copper samples which are measured. Only when the resistance becomes very low, does the resistance due to the lattice imperfections and impurities become important, since it is quite small. An enlarged set of curves for various copper samples is shown in Fig. 3 for the lower part of the temperature range.[7]

Here it can be seen that the residual resistance depends very much on the history of the sample and can vary from 0.3 per cent of the room temperature value for highest purity annealed copper to about 0.6 or 0.9 per cent for the other two samples. Probably the most typical figure for untreated copper would be about 1 or 2 per cent, although it is subject to a considerable variation. Strangely enough, technical copper is often better from the standpoint of residual resistance than spectrographically pure copper, for reasons that are not

[5] D. K. C. MacDonald, "Properties of metals at low temperatures," *Prog. in Metal Phys.*, vol. 3, p. 55; 1952.

[6] D. K. C. MacDonald and R. Berman, "The thermal and electrical conductivity of copper," *Proc. Roy. Soc.*, ser. A, vol. 211, p. 122; 1952.

[7] D. K. C. MacDonald, "Minima of electrical resistance in gold and copper," *Proc. International Conference on Low Temperature Physics*, Oxford, p. 58; Aug., 1951.

apparent. Annealing is of great importance, but it is also important that no oxygen come into contact with the wire in the process.

One complication can be seen in both Figs. 2 and 3, and this complication cannot be explained by modern theory. Impurities should always add a constant value to the residual resistance, but these figures clearly show a minimum in the resistance of copper at about 15° Absolute for certain samples. This has been observed in other metals, also, and its cause is now believed to be due to the presence of minute traces of impurities that often cannot be detected.[8] In some cases, the resistance is found to be increasing rapidly with decreasing temperature and tends to become infinite at absolute zero, very contrary to present theoretical beliefs. This resistance minimum is just one contributing fact to the feeling that the theory of metallic conduction is not complete by any means.

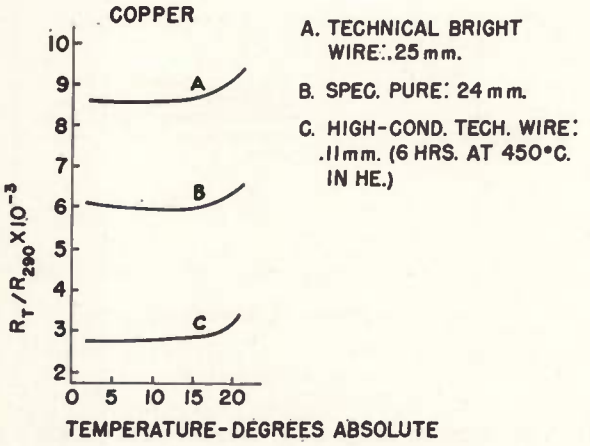

Fig. 3—The residual electrical resistance of various samples of pure copper.

There is also another interesting low-temperature effect which is connected with the impurity effect. Electrical resistance exists because of the interactions of the electrons and the lattice of a crystal or because of the deflections of the electron as it tries to go through the body. At low temperatures in a pure metal, the electron can travel a relatively long distance between collisions with atoms, and it is possible that it collides with the walls of the metal about as often as with atoms or internal impurities. Thus, the principal source of scattering will be the walls, and the dimensions of the sample will become important. In practice, this effect becomes predominant only when the distance traveled by the electron between atomic collisions is on the order of the dimensions of the sample and is about 0.1 millimeter at 4° Absolute. Nevertheless, it is important in thin films and fine wires and frequently must be taken into account.

The above discussion applies mostly to the pure metals but can be extended quite easily to the alloys. Here, the mixture of atoms which are not too similar acts as if there were considerable impurity present and has, consequently, a large residual resistance. In fact, the residual resistance in many cases is so large that it swamps the temperature-dependent term, and the electrical resistance is practically temperature independent. This is shown in Fig. 4 for some common alloys.[9] The resistance may show a slight tendency to decrease as the temperature is lowered, but not significantly.

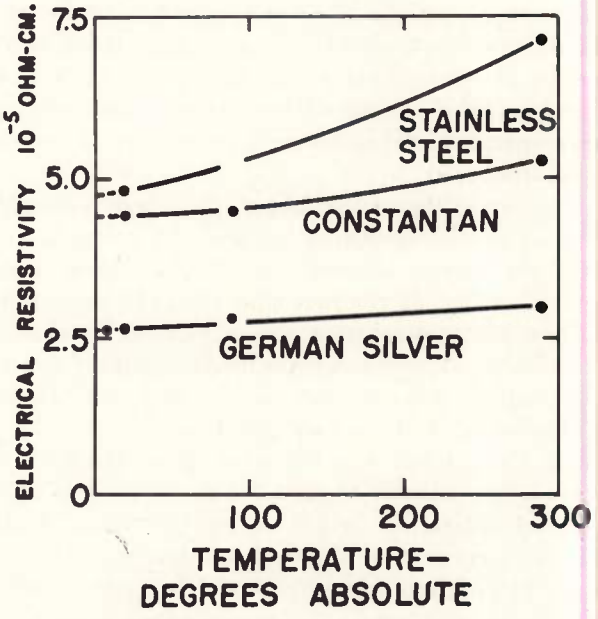

Fig. 4—The temperature dependence of the electrical resistance of alloys.

In general, then, it can be said that the pure metals are characterized by a fairly low electrical resistance, which decreases with temperature. As soon as any impurities are added to them (such as in brass, stainless steel, dural, etc.) the electrical resistance increases and tends to become almost temperature independent. Since the electrical resistance and thermal conductivity of metals are closely related, the poor thermal conductors can also be looked for among the alloys—and the good conductors among the pure metals. This does not mean, however, that some alloys cannot have lower resistance than specific metals at a given temperature.

What does all this add up to from a practical viewpoint? Whenever ohmic losses are important, as in an inductance, decreasing the temperature can be a help. Low losses and high inductance seem to be mutually exclusive for moderate frequencies, since high inductances require many turns of fine wire, and consequently, have a fairly high resistance. However, with the decrease in resistance expected on cooling, the Q of quite ordinary copper coils can be expected to rise appreciably. To show this some very crude, elementary experiments were carried out with a simple iron-cored audio-frequency inductance.

[8] D. K. C. MacDonald, "The minimum of electrical resistance at low temperatures," letter in *Phys. Rev.*, vol. 88, p. 148; 1952.

[9] R. Berman, "The thermal conductivity of some alloys at low temperatures," *Phil. Mag.*, vol. 42, pp. 642–650; June, 1951.

The results were surprisingly good. The Q and inductance of two coils were measured at 1,000 cps, together with the dc resistance, at room temperature (300° Abs.), liquid nitrogen temperature (77° Abs.), and liquid helium temperature (4° Abs.). The two coils behaved similarly, and the data for one are given in Fig. 5. The air-core inductance was about 1.18 henries, and could be increased to 8 henries by the addition of an iron core. The inductance proved to be practically independent of temperature. This coil had a room temperature resistance of 4,200 ohms.

The behavior of the Q was roughly what had been expected. Beginning with a value of about 1.75 for the air core, it increased to 12 in liquid nitrogen (a factor of 7), and to 88 in liquid helium (a factor of 50). Adding the iron-core increased the Q to 9 at room temperature but had no effect on the Q (although the inductance was increased) in liquid helium. Thus losses in the core were exceedingly important at these temperatures, and they proved to be very current sensitive, increasing as the measuring current was increased.

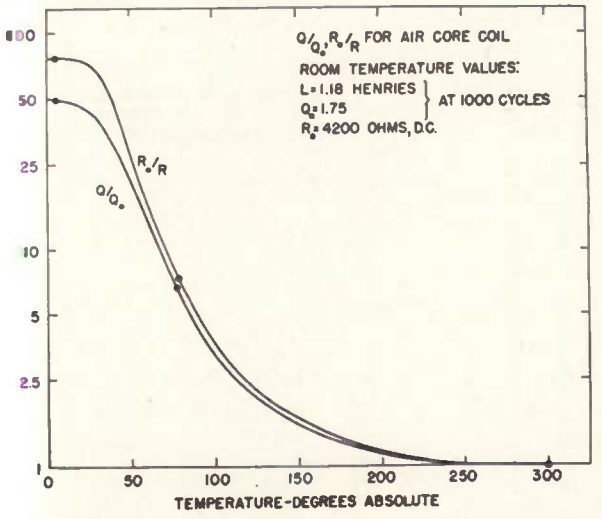

Fig. 5—The relative values of the Q at 1,000 cps and the dc resistance of an audio-frequency air-cored coil. (The smoothed curves were drawn through the three experimental points with Fig. 2 in mind.)

Another interesting effect, shown by the comparison of the two curves in Fig. 5, is that the Q did not increase as much as the dc resistance would lead one to expect. Thus, while the dc resistance decreased by a factor of 75, the Q increased by a factor of only 50. This effect is probably due to the skin effect, since the skin depth depends on the ratio f/R. Thus for copper, 1,000 cps at liquid helium temperature would correspond to room temperature and 75,000 cps. Actually, ordinary skin effect theory does not apply at these temperatures, and the surface impedance is increased even more than would be expected. It is worth emphasizing that, while Q's will tend to increase, they will not increase as much as the dc resistance decreases.

Dr. D. K. C. MacDonald has suggested that one possible application here would be in the manufacture of transformers of extremely high turns ratio for the matching of low-impedance sources to high-impedance amplifiers. Because of the relatively low resistance of copper at these temperatures, very fine wire could be used in many turns without loss of efficiency. Similarly, compensating mutual inductances are often used in low-temperature work to measure small changes in inductance. At room temperature, these mutual inductances are large and unwieldy because of the current that they must carry. Also because of their size, they are very susceptible to pickup. Dr. R. J. Benzie at Oxford used a compensating coil immersed in liquid nitrogen, which was relatively small and free from pickup, and yet could carry fairly large currents without heating up because of the low resistance of the copper. Other applications of the same type should suggest themselves to people working in fields where ohmic losses are a problem.

One of the most spectacular low-temperature effects is also of interest in this respect—namely, the phenomenon of superconductivity.[10] It was first discovered by the Dutch physicist, Kamerlingh Onnes, during an investigation of the behavior of electrical resistance with temperature. While gold, silver, copper, and many other metals behaved properly and gave finite residual resistances, mercury, lead, tin, and indium seemed to deviate from this pattern and become abruptly resistanceless at definite temperatures which were characteristic of the metals themselves. There was no indication from the behavior of the metal above the transition point that this phenomenon was about to take place, and no general criterion for predicting superconducting properties has been discovered since. It is definitely not believed that all metals will become superconducting at absolute zero.

Original resistance-temperature curve for mercury, as obtained by Kamerlingh Onnes, is shown in Fig. 6, page 414, with abrupt transition from a residual resistance to zero resistance at about 4° Absolute.[11] Lead, tin, and indium also become superconducting in this region near 4°, while aluminum and zinc have transition points at lower temperatures. The exact temperature tends to depend on the purity of the metal and its state of annealing.

Popularly, the superconductors are identified with the fact that once a current of several amperes is started in a ring of metal, the current will persist indefinitely, with no ohmic losses. This sounds most promising, and immediately vast horizons are opened up, principally in the field of lossless transmission of electrical power. Unfortunately, there are two limitations. First, almost all the pure metallic superconductors exist in the liquid helium region or at lower temperatures, although the pure metal, niobium, has a transition point at about 10° Absolute. Alloys do exist with higher transition temperatures (up to 17° Abs.), but they are generally not pure

[10] "Four Lectures on Low Temperature Physics," Pergamon Press, London, Eng., pp. 95–132; 1952.

[11] H. Kamerlingh Onnes, "Superconductivity of mercury," *Leiden Comm.*, vol. 124, p. 23; 1911.

superconductors in the true sense of the word and exhibit some more complicated effects. In general, the main disadvantage of superconductors is that they exist only in a temperature region in which it is extremely difficult to work on a large scale. Their main practical use, therefore, would seem to be on a laboratory scale, where zero resistance is important.

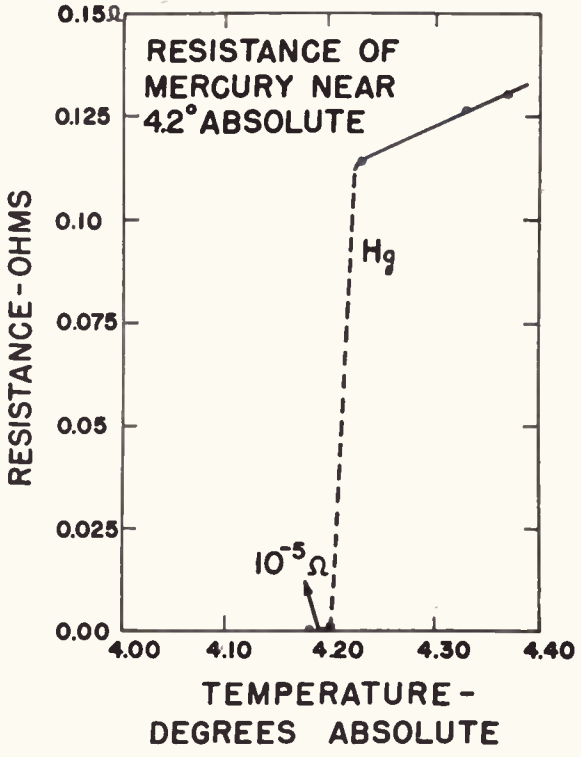

Fig. 6—The electrical resistance of mercury near its superconducting transition point.

A second limitation to the use of superconductors concerns the maximum current which a superconductor can carry. Experimentally, it has been found that superconductivity can be destroyed by placing the specimen in a sufficiently large magnetic field. The critical magnetic field below which the metal is superconducting, and above which it is normal, is a function of temperature. Typical curves of critical field vs. temperature are shown in Fig. 7. When too large a current is passed through a superconductor, the magnetic field at the surface of the superconductor may become larger than the critical field for that temperature, and the superconductivity will be destroyed. This is the basic mechanism for the current-limiting phenomenon.

Superconductivity is not well understood theoretically, although a quite good macroscopic theory does exist. This theory, an adaptation of ordinary electromagnetic theory, is capable of predicting effects but gives no insight as to why they occur.[12] It is found, for instance, that no magnetic field exists inside a superconductor (it has zero permeability), and there are indications that this is as basic a property of superconductors as the phenomenon of zero resistance. This turns out to be the most reliable way of detecting the superconducting transition, since it is independent of electrical leads going into the sample itself. In practice, the sample is placed at the center of a set of mutual inductance coils, and the superconducting transition is marked by a change in the mutual inductance of the system.

Electromagnetic theory requires that if the magnetic lines of force are to be expelled from the system ($B=0$), then the electric field inside the system must also be zero. The lack of an electric field means that potential differences are impossible in a superconductor, which leads to the original observation of zero resistance. Physically, it is not possible to have these two fields, B and E, go to zero abruptly at the surface of the metal. In fact, they fall off exponentially in a short distance known as the penetration depth (about 10^{-5} mm). The penetration depth becomes important in two cases: when the size of the specimen is of the same order of magnitude as the penetration depth, or when the skin thickness (identified with high-frequency resistance) becomes of this order of magnitude at high frequencies.[13,14,15]

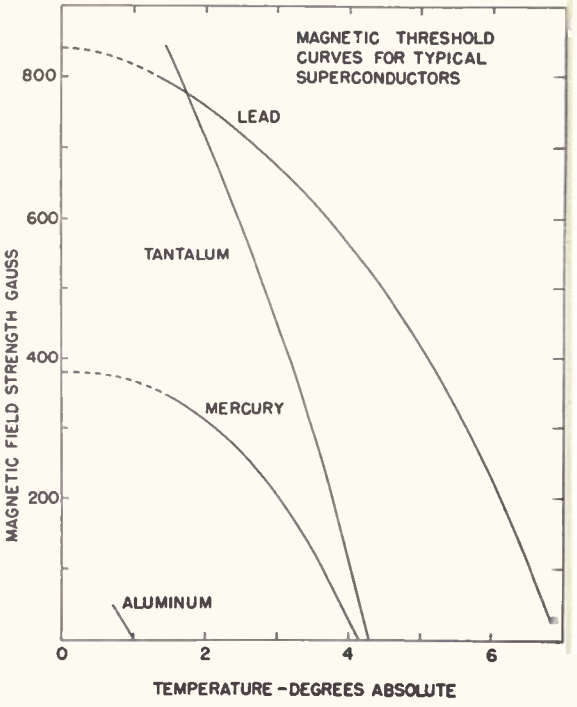

Fig. 7—The magnetic threshold curves for various typical superconductors.

These two are equivalent when the metal within the penetration depth is thought of as containing a combination of both normal and superconducting electrons with only the normal electrons being involved in ohmic losses.

[12] F. London, "Superfluids, Vol. I. Macroscopic Theory of Superconductivity," John Wiley and Sons, Co., New York, N. Y.; 1950.

[13] E. Maxwell, P. M. Marcus, and J. C. Slater, "Surface impedance of normal and superconductors at 24,000 megacycles per second," *Phys. Rev.*, vol. 76, pp. 1332–47; Nov. 1, 1949.

[14] A. B. Pippard, "The high frequency resistance of superconductors," *Physica*, vol. 15, pp. 40–44; April, 1949.

[15] I. Simon, "Surface impedance of superconducting tin, mercury, and lead at 9200 mc/sec," *Phys. Rev.*, vol. 77, pp. 384–387; Feb. 1, 1950.

Thus, at high frequencies, the resistance of a superconductor is not zero but has a finite value.

This is shown in Fig. 8, where the Q of a coaxial resonator made from lead is shown as a function of temperature. The frequency corresponds to a wavelength of about 3 centimeters. If the metal were to remain normal the Q would approach a constant value of about 3,400 at absolute zero due to the residual resistance. This would be observed if the cavity were placed in a magnetic field, for instance. But due to the superconducting transition, the Q increases to a value of about 80,000 at absolute zero—not infinity, since the penetration depth is finite and the skin resistance is not zero. Q's of the order of 10 million have been observed, although no practical use has been made of them as yet.

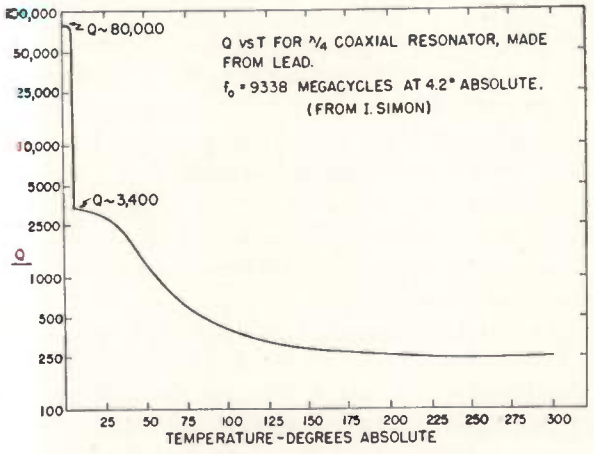

Fig. 8—The Q of a microwave lead resonator as a function of temperature.

There are two interesting devices which have been used or suggested which utilize the superconducting properties of metals. The first is a very sensitive bolometer due to Andrews[16] at Johns Hopkins University. This instrument makes use of the very large change in the resistance of a superconductor over a relatively small temperature interval along its transition curve (see Fig. 6). Thus, if the resistor can be biased to be in the middle of the transition curve, the addition of a small amount of heat will change its temperature slightly, and the voltage across it (for constant current) very much. This is the basis of the superconducting bolometer as devised by Andrews. In practice, an alloy of niobium, niobium nitride (NbN), was used, which had a transition point near 15° Absolute. The reason was partly that it is easier to work with liquid hydrogen than with liquid helium. The main drawback with this apparatus seems to be that it takes very accurate temperature control, and unexpected noise is found along the transition curve in the system which is not superconducting and yet not normal.

A second device using superconductors was suggested recently by I. Simon,[17] when he showed that a superconducting lead sphere could be placed in rotation and suspended in space by a suitable arrangement of coils. This is a truly frictionless suspension, since eddy current effects can be eliminated, leaving only the friction due to the helium gas necessary to keep the sphere in thermal contact with the surrounding liquid helium. The idea is only of experimental interest at present, although it has been used to measure the viscosity of gaseous helium in the neighborhood of 4° Absolute.

The above discussions make no pretense at being complete and have included only a few selected low-temperature topics. This rough survey of the basic behavior of noise and electrical resistance at low temperatures has been made with the hope that a knowledge of these phenomena might aid in the solution of some existing problems, or might stimulate further research of a practical nature.

Acknowledgment

The authors wish to express their appreciation to Dr. D. K. C. MacDonald, Dr. I. Simon, and Dr. H. O. McMahon for interesting discussions which led to much of this paper.

[16] D. H. Andrews, R. M. Milton, and W. de Sorbo, "A fast superconducting bolometer," *Jour. Opt. Soc. Amer.*, vol. 36, pp. 518–524; Sept., 1946.

[17] I. Simon, "Forces acting on superconductors in magnetic fields," *Jour. Appl. Phys.*, vol. 24, pp. 19–24; Jan., 1953.

Correction

The following error in the paper, "A Survey of Digital Computer Memory Systems," by J. P. Eckert, Jr., which appeared on pages 1393–1406 of the October, 1953 issue of the PROCEEDINGS OF THE I.R.E., has been brought to the attention of the editors by the author.

Page 1397, under heading "Electrostatic Systems," sentence commencing on line 13 should read: "Prior to this, McConnell at MIT did some work recording radar signals."

This reference also should have been listed and detailed in the Bibliography: McConnell, R. A., "Video storage by secondary emission from simple mosaics," PROC I.R.E., vol. 35, pp. 1258–1264; Nov., 1947.

Transistor Reliability Studies*

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Summary—**Factors affecting the useful life of transistors are discussed. From consideration of their structure, long life might well be expected, and has been shown to occur for usual laboratory conditions. However, under high humidity or high temperature, life is sometimes greatly shortened. Mechanisms of the effects are described. Corrective measures are under way.**

EXPECTATION of great reliability is one of the attractive prospects accompanying the use of transistors. Since the fundamental mechanisms involve rugged solids, one might think a transistor should last forever. Nevertheless a rather short life is sometimes encountered, and demands explanation. This report summarizes some mechanisms which have been found to cause failure in particular types of transistors, and draws conclusions as to the extent and means whereby the situation can be improved. It should be clear that the conditions described in large part have been corrected, and that units now being produced are much better than some of the test curves shown.

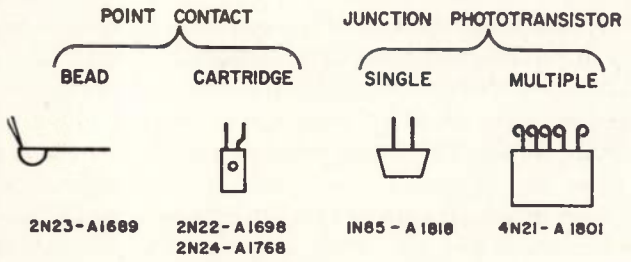

Fig. 1—Transistor types tested.

Tests on early transistors at Bell Telephone Laboratories showed quite excellent life and led to extrapolated values of average life which were remarkably high, such as 70,000 hours. These transistors were encapsulated with waxlike materials. The tests simulated an early telephone application, and were made under ordinary laboratory conditions. It is now known that shelf tests under high temperature and humidity conditions would have shown much shorter life. Investigations to elucidate and improve conditions have been carried out by means of a variety of experiments described below and by means of accelerated life tests on various transistors, including point-contact types and grown-junction diodes and triodes in both cartridges and plastic beads. Outline sketches of studied units are shown in Fig. 1.

Field experience and accelerated life tests on some of these units will first be described. Then mechanisms explaining the behavior will be outlined, then measures appropriate to improve the behavior. Finally, conclusions will be drawn.

* Decimal classification: R282.12. Reprinted from CONVENTION RECORD OF THE I.R.E., part 6, pp. 9–14; 1953.
† Bell Telephone Lab., Murray Hill, N. J.
‡ Hughes Aircraft Co., Culver City, Calif.

FIELD EXPERIENCE OF SOME EARLY TRANSISTORS

When early transistors were exposed to the pitiless gaze of various circuit users, a number of ailments were found which can be tabulated here in a sort of transistor "toxicology."

1. A very gradual drift in the characteristics with time. Particularly affected are the reverse currents of the collectors, both point and junction. This disease was the factor which limited the life to 70,000 hours in the original life tests; since it ordinarily takes a long time to become appreciable, it is known as the "slow death."

2. A gradual development with time of what appears to be a leakage path between the collector and emitter. Not very noticeable in most point-contact transistors, this disease is more virulent in junction transistors, particularly grown types which normally have very high resistance levels; it shows up as a variable floating potential on the emitter when the current is cut off. Since the ailment concerns emitter current cutoff conditions, it is called "sleeping sickness."

3. In some point-contact transistors the current multiplication-factor alpha may become markedly reduced, particularly at low voltages. Though normally rare, this occurrence has at times reached an incidence as high as 25 per cent for some types. Since this disease may occur quickly without previous warning, it goes by the name "sudden death."

4. Sometimes loss of alpha has occurred prior to receipt of the transistor by the customer. Such units are declared "dead on arrival."

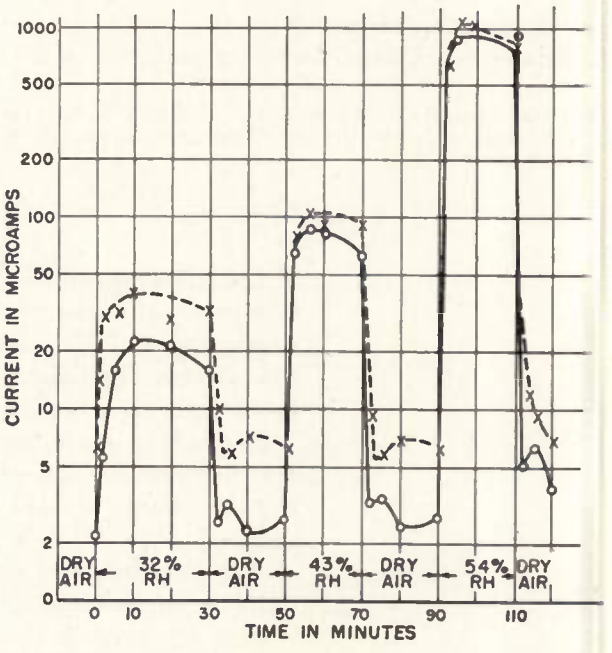

Fig. 2—Effects of water vapor on a bare *n-p-n* junction bar.

Mechanisms Explaining Changes in Characteristics

The "slow death" is explained in terms of ions resident on the germanium surface. A number of different kinds of ions can change the surface conditions of the semiconductor; in particular, water vapor is a common contaminant which can cause considerable changes.

The effects of moisture on germanium surfaces may be shown by exposing an entirely unprotected *n-p-n* junction bar to various humidities, as in Fig. 2. The saturation current in this bar was about 2 microamperes originally in dry air; on exposure to 32 per cent relative humidity, the saturation current increases in a few minutes by an order of magnitude. Return to dry air substantially restores the original condition. Still higher humidities cause markedly larger effects, which are still reversible by drying out. The dotted curve shows very similar results with the applied voltage reversed, indicating that the moisture effects are the same at both junctions.

When the finished transistor is encapsulated in a moisture-resistant wax and plastic, the effects are greatly slowed, as shown in Fig. 3 for A1818 phototransistors of an early design. This unit contains a p-n junction bar enclosed first in wax and then in a thin transparent plastic block of precise dimensions. Under ordinary laboratory conditions, life tests have shown no significant trends in a year of operation, but a small increase in reverse saturation I_d current appears during the second year. However, in 100 per cent humidity, as shown in the figure, the plastic coating protected these units for only about 200 hours at 25 degrees C. For each temperature rise of 10 degrees C., speed of the effect doubles in agreement with the speed of water vapor diffusion through the plastic and along the leads into the bar.

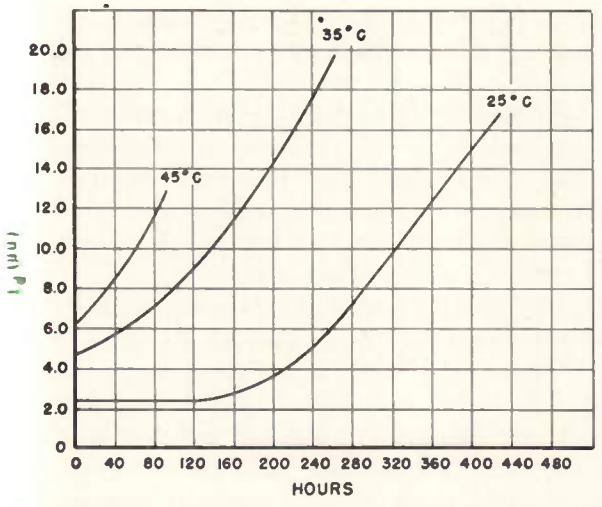

Fig. 3—Effects of water vapor on A1818 phototransistors with inadequate protective encapsulation.

The original characteristics are usually substantially recoverable by drying out the units in an oven, as shown in Tables I and II below. The history of six A1818 phototransistors and of three elements of an A1801 multiple phototransistor shows that the effect of moderate exposure to humidity was reversed by a bakeout for 96 hours at 55 degrees C.

TABLE I

Recovery of Phototransistors on Heating
A1818 Phototransistors

Unit No.	Volts	Initial I_d	Final After Humidity Run	I_d After Bake Out
25 degrees C. 100 per cent R.H.				
E 418 B	10	2.8μa	14.5μa	4.7μa
E 180 C	10	2.6μa	13.5μa	2.8μa
35 degrees C. 100 per cent R.H.				
E 431 B	10	4.6μa	22.0μa	2.0μa
E 181 B	10	4.8μa	17.5μa	5.2μa
45 degrees C. 100 per cent R.H.				
E 433 B	10	6.8μa	18.4μa	4.8μa
E 200 C	10	5.6μa	17.5μa	5.2μa

TABLE II

Recovery of Phototransistors on Heating
A1801 Phototransistors

Unit No.	Volts	Initial I_d	Final After Humidity Run	I_d After Bake Out
65 degrees C. 100 per cent R.H.				
39	10	3.3μa	10.4μa	3.8μa
17	10	2.8μa	14.0μa	1.8μa
34	10	3.6μa	13.4μa	2.3μa

These tests show that "slow death" can be caused by moisture at the junctions. However, the possibility of other ions being involved also, either harmfully or helpfully, is certainly not excluded. It is noteworthy that the sensitivity to moisture may vary considerably from unit to unit and even may vary with time for a particular unit. It may be possible to develop insensitive surfaces which will not show the effect.

Similar effects have occurred in point-contact transistors. The effect of shelf-aging five bead-type M1689 transistors, prior to design changes, in 97 per cent humidity, is shown in Fig. 4. The collector current I_c (0, −40) drawn at zero emitter current and −40-volts collector voltage, under these high humidity conditions shows a sharp change, usually an increase, rather suddenly after about 400 hours.

Point-contact transistors of cartridge type are slower in showing the effect, since they are protected by thicker protective coatings. Fig. 5 shows the average value of I_c (0, −40) for several A1698 cartridge units of older design, in which only minor fluctuations appear. The average of this parameter stays continuously within the type limit (straight dotted line) for at least 2,500 hours

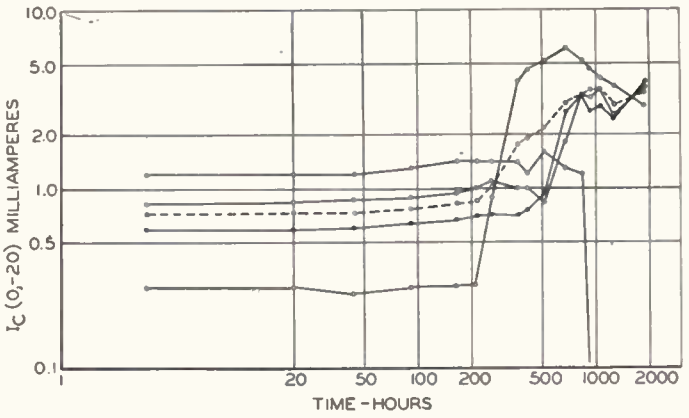

Fig. 4—Shelf aging of bead M1689 transistors at 25 degrees C. in 97 per cent humidity prior to design changes.

even at 55 degrees C. Before too much optimism is engendered by this apparent immunity to "slow death," however, let it be recorded that several of these units, encountered "sudden death," which is now to be described. As will appear below, the "sudden death" produces a sharp change, mainly in the parameter V_e (1, −2). Units showing this effect were excluded from the averages of Fig. 5.

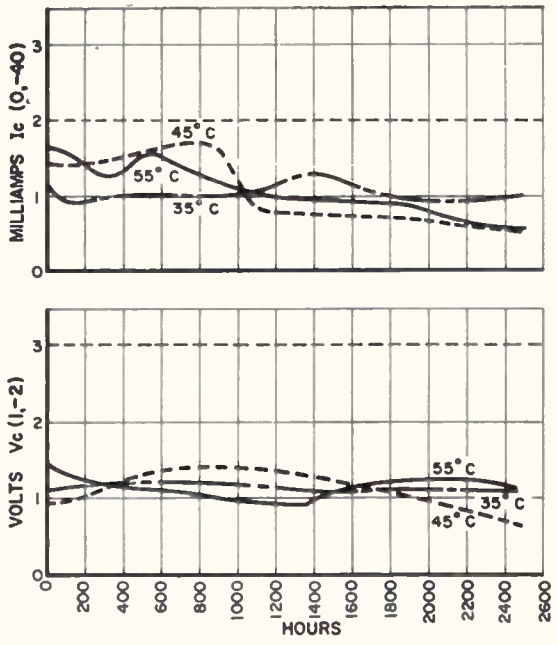

Fig. 5—Effect of 100 per cent relative humidity on A1698 transistors prior to design changes.

The electrical condition which goes by the name of "sudden death" is described in Fig. 6, in which the dashed curves depict the collector characteristics of a normal M1689 bead point-contact transistor. The solid lines are the characteristics of a particular unit after "sudden death" has occurred. It is seen that the current amplification factor alpha, corresponding to the spacing between curves, is substantially less than before, particularly at low voltages. Note that a particularly sensitive measure of this effect is a large increase in the parameter V_e (1, −2), the collector voltage at currents of 1 milliampere in the emitter and 2 milliamperes in the collector. It is both curious and significant that "sudden death" often produces practically no change in the collector diode characteristic line (I_e=0); the change is only in alpha, the transfer characteristic between emitter and collector diodes.

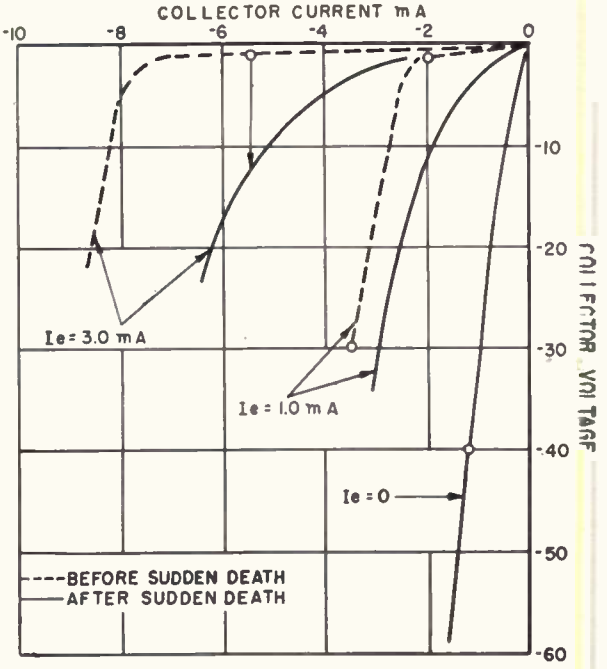

Fig. 6—"Sudden death" in point contact transistors.

By experiments in a manipulator it has been found that the symptoms of sudden death can be induced by a surprisingly small mechanical change in the angle of the collector point of a lightly formed transistor. Fig. 7 shows the results of such an experiment, in which the collector points held in a manipulator were rocked somewhat less than 1 degree in all four directions, meanwhile measuring I_e (0, −40) and V_e (1, −2). The rocking was accomplished through a lateral motion of 0.002 inches of the manipulator in which the collector lead was clamped at a distance of some 0.180 inches from the point. Meanwhile the collector point remained continuously in contact with the surface, which was rough enough to prevent lateral motion of the point.

In Fig. 7, the manipulator transistors are labeled 1 to 4 and the directions of rocking are called x, x', y, and y'. Each transistor was measured in two conditions, A and B. In the A condition, the minute rocking described was more often than not sufficient to cause V_e (1, −2) to exceed its 3-volt limit. On restoration to the original position, the transistor usually recovered satisfactory performance. The B condition shows that heavier forming makes the characteristics practically insensitive to rocking, at the expense of some increase in I_e (0, −40).

To complete the explanation of the disease it is necessary to account for the motion of the pins to which the collector wire is secured. This finding was rather unexpected, in view of the fact that units of

this sort have withstood vibration and shock of many thousand g with no apparent effect. The actual contacts of course are not mathematical points but are rather small areas, which in some units may be as large as $\frac{1}{2}\times 3$ mils in size. Contact force is supplied by the deflection of a spring section in the wire structure. In bead units, the upper end of the wire beyond the spring is directly anchored in the beading plastic; in cartridge units, the wire is swaged into the stem lead, which in turn is held by a plastic insulator.

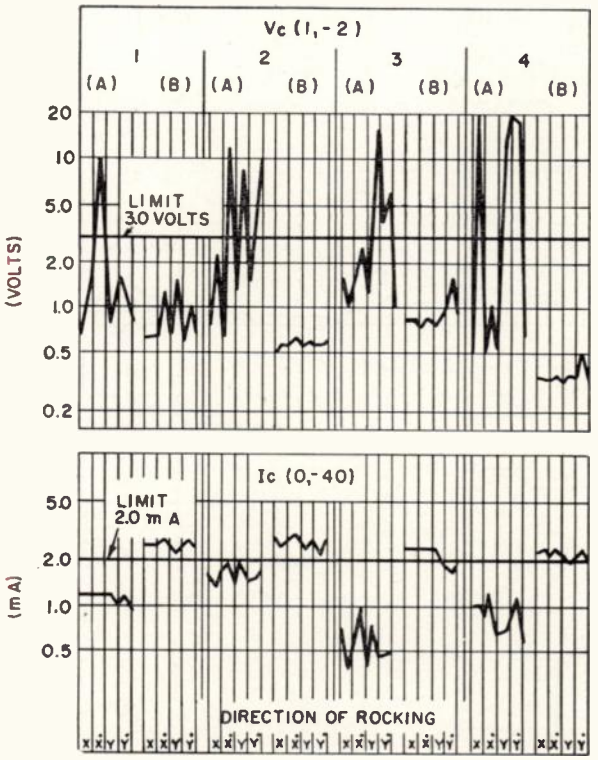

Fig. 7—Effects of rocking collector point.

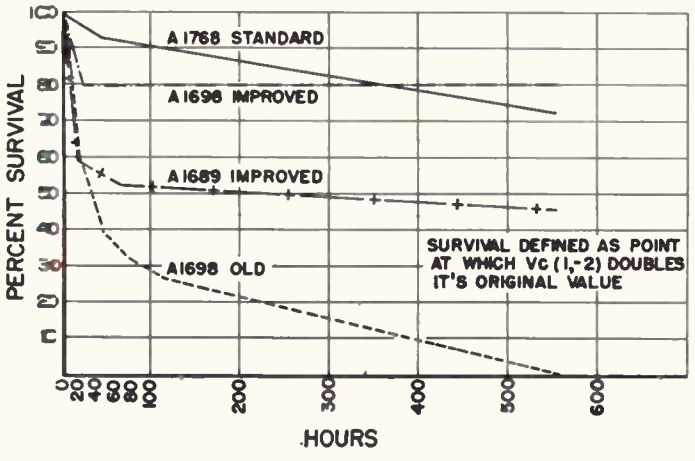

Fig. 8—Accelerated life tests on point contact transistors.

From the above rocking experiment one sees that rather small motions of the supporting insulator would be sufficient to cause "sudden death." Such motions might be caused by mechanical force on the unit or by temperature cycling; from the life test evidence, one concludes that high humidity can also do the job. Most probably the humidity mechanism involves warping of the plastic insulator, caused by moisture absorption.

Results on units which incorporate improvements based on the above mechanism are shown in the "*B*" data of Fig. 7. In Fig. 8 are shown the results of accelerated life tests (100 per cent relative humidity at 65 degrees C.) on units of several types, including a direct comparison showing the improvement in the 1698 transistors. If one considers that failures in the first few hours might be weeded out by factory aging tests, the preliminary conclusion might be that the improved resistance to the effects of humidity may be entirely satisfactory, at least for 1698's.

Ability to withstand temperature cycling is also improved. Fig. 9 shows the effect of cycling fifteen of the old M1689 beaded units from room temperature to 70 degrees C., with a fifteen-minute dwell at each end.

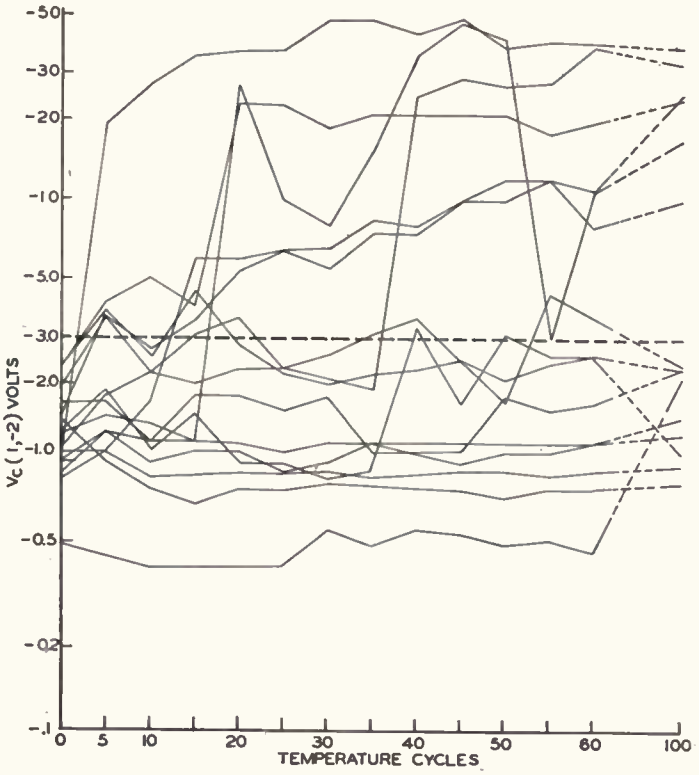

Fig. 9—Temperature cycling of old M1689 units.

Nine of the fifteen units showed symptoms of sudden death at some time during the test, and six were still dead at the end of the test. (The remaining three had been resurrected by continuation of the cycling.) On the other hand, for five units improved by heavier forming, Fig. 10 shows that sudden death did not occur at any time during 100 temperature cycles.

Life tests on point contact transistors in dry air give confirmatory evidence for the foregoing mechanisms and also give encouraging forecasts for the life of units in which the effects will have been prevented by design changes. For these tests, cartridge transistors of the old 1698 type were enclosed with a desiccant in containers

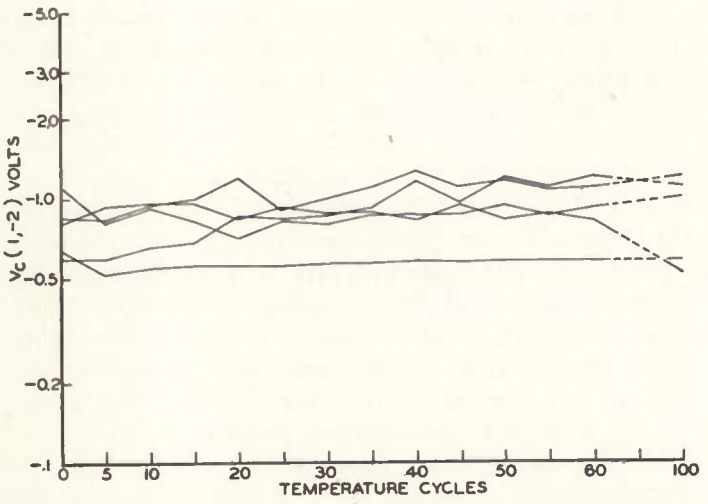

Fig. 10—Temperature cycling of improved M1689 units.

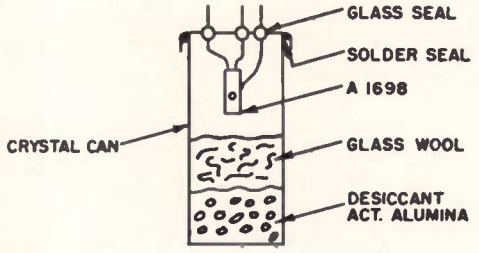

Fig. 11—Construction of test fixture to provide dry atmosphere.

of glass-metal construction (Fig. 11). These dry units were then shelf-aged 2,800 hours under normal laboratory conditions, 25 degrees C. and about 50 per cent relative humidity, with the results on I_o (0, −40), V_c(1, −2), and α(1, −30) shown in Figs. 12–14.

Since the changes in 2,800 hours were minor, the test was then altered to provide more severe conditions. While still in the desiccated cans, one-third of the units were subjected to 65 degrees C. and 100 per cent RH; one-third to 25 degrees C. and 50 per cent RH, a continuation of the original test; and the remainder to 65 degrees C., 0 per cent RH. The new conditions had the effect of testing the tightness of some of the cans, and also temperature-cycled some of the units, since all measurements were made at 25 degrees C. Examination of the figures shows that changes in the units were small and that sudden death and slow death did not occur.

A comparison of the dry units with other unimproved A1698's subjected to 100 per cent humidity is shown in Fig. 15. None of the dry units has failed to date. This test shows that humidity was a factor (sometimes primary and sometimes secondary) in the behavior of these units previous to the design changes.

The last malady, "sleeping sickness," is described in Fig. 16. A normal biasing potential is applied to the collector of the transistor, and the "floating" potential appearing at the open-circuited emitter is measured with a high-impedance voltmeter. This floating potential should have the value $-(kT/q)$ ln $(1-\alpha)$ or about 0.05 volt for normal units at 25 degrees C. However, if the unit has "sleeping sickness," the emitter floating potential may drift to much higher values, approaching the collector potential in severe cases. The unit acts as if a surface leakage resistance were developing, connecting the emitter and collector regions by a conducting channel across the surface of the base layer.[1] Such a channel may be induced in several ways, a common one being exposure to high humidity.

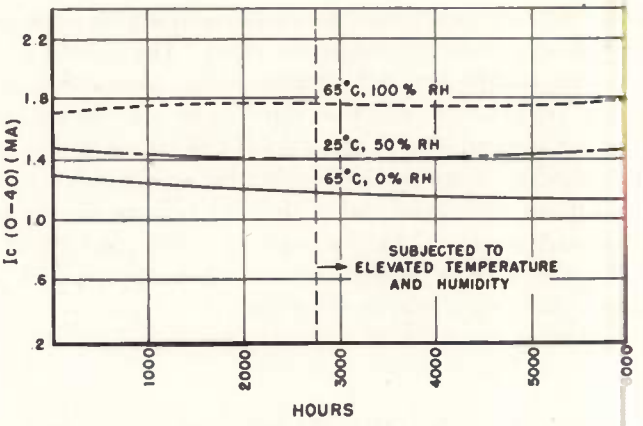

Fig. 12—Life test on A1698 units in dry air.

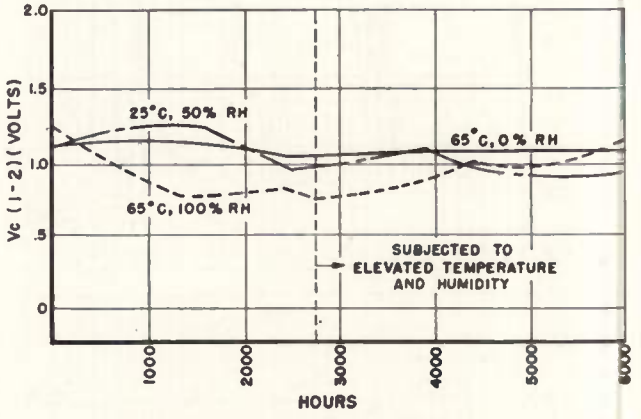

Fig. 13—Life test on A1698 units in dry air.

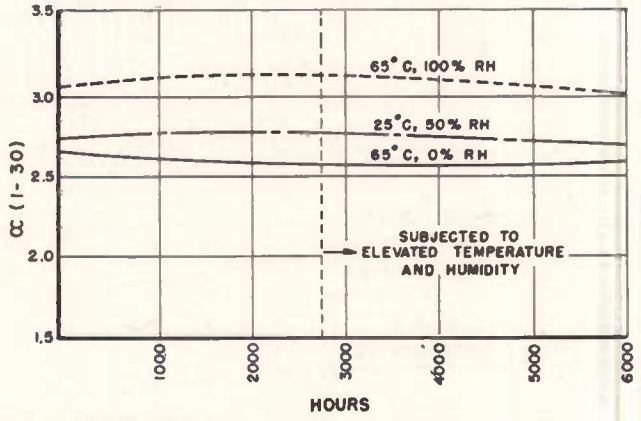

Fig. 14—Life test on A1698 units in dry air.

[1] This effect is discussed by W. L. Brown in "N-type conductivity on p-type germanium," *Phys. Rev.*, vol. 91, pp. 518–527; August 1, 1953.

Fig. 16 also shows the results of tests on two groups of grown *n-p-n* junction transistors. The top curve shows the behavior of M1752 units encased in a wax-plastic assembly, while the lower curve is for similar units sealed in a glass-metal container. One sees that the sealed units are improved, but not entirely free of the problem. Besides sealing, it appears necessary also to take more stringent precautions against sealing up water or other ions inside the units when they are made. At present the difficulty is most severe for grown junction units because the leakage path across the base layer is very short.

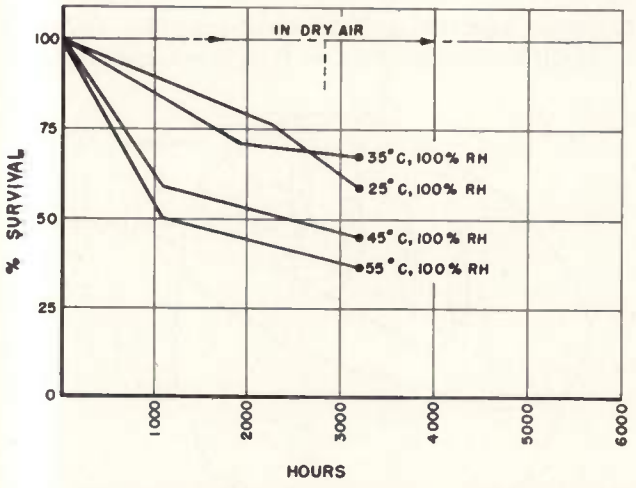

Fig. 15—Effect of humidity on A1698 units prior to design changes.

More recent tests indicate that it is possible by proper cleaning and surface treatment to eliminate "sleeping sickness" even for the most severe accelerated tests. However, further experiments of greater statistical significance are required before one can say with certainty that sleeping sickness has been eliminated.

In alloy junction transistors having longer leakage paths, and in point-contact transistors having lower impedance levels, it is possible that for many purposes adequate protection against sleeping sickness may be obtainable with proper plastic encapsulation without resort to glass-metal seals.

Conclusions

A number of mechanisms affecting reliability of transistors has been studied. They involve:

1. Susceptibility of transistor surfaces to ionic contamination, either originally present, or penetrating the unit encapsulation during life.
2. Mechanical stability, including not only the point rocking discussed in detail above, but also more obvious troubles like solder failures or loss of pressure contacts.
3. Electrical and chemical processing of the units during and after fabrication.

Many of the troubles can be caused by humidity, either directly by affecting surface conductivity of the semiconductor, or indirectly by affecting mechanical stability.

In terms of these mechanisms, improvements in design are being currently applied. While great improvements in reliability are indicated, more extensive tests and design changes are needed.

Recommendations

As is obvious from consideration of the mechanisms, there is no one panacea which will assure absolute reliability. Hermetic sealing against the effects of external humidity should be helpful, but one must also consider a variety of other factors, such as proper initial condition of the surface, and proper mechanical and electrical design and processing.

We believe that for many purposes completely adequate reliability can be obtained at lower cost from proper plastic encapsulation. There is evidence that the plastic is penetrated by moisture not so much through its bulk as along the electrode leads, and considerable improvement seems possible by using better lead seals and more tightly-bonding plastics. Much can also be done to reduce the sensitivity of transistor surfaces to moisture. The degree of protection obtainable by such means is being determined.

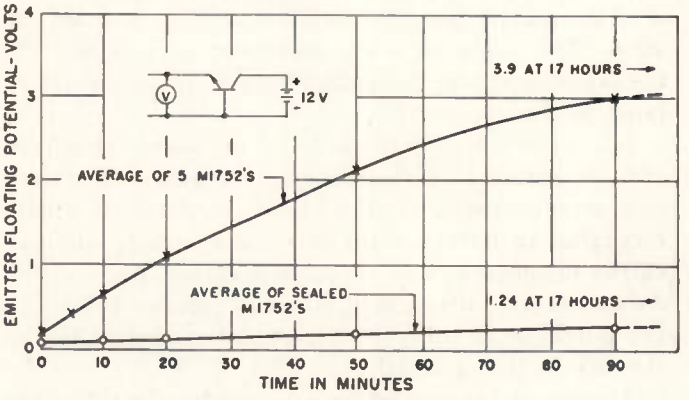

Fig. 16—Improvement of emitter floating potential by sealing *n-p-n* grown junction transistors.

At this writing it appears probable, though not certain, that for the greatest protection for very long periods against extreme humidity, hermetic sealing will be necessary. In any case, immediate use of glass-metal hermetic seals is being inaugurated for a number of the most sensitive transistor types. However, in view of the greater size and greater complexity of such seals, and in view of the fact that there exist difficulties not curable by these means, we believe properly designed plastic-encapsulated units are still competitive for many uses.

At these laboratories a broad program is still continuing for study of the mechanisms of change in transistors. It is expected that further modifications leading to improved reliability will be forthcoming.

Acknowledgment

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Resistivity Measurements on Germanium for Transistors*

L. B. VALDES†, MEMBER, IRE

Summary—**This paper discusses a laboratory method which has been found very useful for measuring the resistivity of the semiconductor germanium. The method consists of placing four probes that make contact along a line on the surface of the material. Current is passed through the outer pair of probes and the floating potential is measured across the inner pair. There are seven cases considered, the probes on a semi-infinite volume of semiconductor material and the probes near six different types of boundaries. Formulas and curves needed to compute resistivity are given for each case.**

Introduction

THE PROPERTIES of the bulk material used for the fabrication of transistors and other semiconductor devices are essential in determining the characteristics of the completed devices. Resistivity and lifetime[1] (of minority carriers) measurements are generally made on germanium crystals to determine their suitability. The resistivity, in particular, must be measured accurately since its value is critical in many devices. The value of some transistor parameters, like the equivalent base resistance,[2] are at least linearly related to the resistivity.

Many conventional methods for measuring resistivity are unsatisfactory for germanium because it is a semiconductor and metal-semiconductor contacts are usually rectifying in nature. Also there is generally minority carrier injection by one of the current carrying contacts. An excess concentration of minority carriers will affects the potential of other contacts[3] and modulate the resistance of the material.[4]

The method described here overcomes the difficulties mentioned above and also offers several other advantages. It permits measurement of resistivity in samples having a wide variety of shapes, including the resistivity of small volumes within bigger pieces of germanium. In this manner the resistivity on both sides of a *p-n* junction can be determined with good accuracy before the material is cut into bars for making devices. This method of measurement is also applicable to silicon and other semiconductor materials.

The basic model for all these measurements is indicated in Fig. 1. Four sharp probes are placed on a flat surface of the material to be measured, current is passed through the two outer electrodes, and the floating potential is measured across the inner pair. If the flat surface on which the probes rest is adequately large and the crystal is big the germanium may be considered to be a semi-infinite volume. To prevent minority carrier injection and make good contact, the surface on which the probes rest may be mechanically lapped.

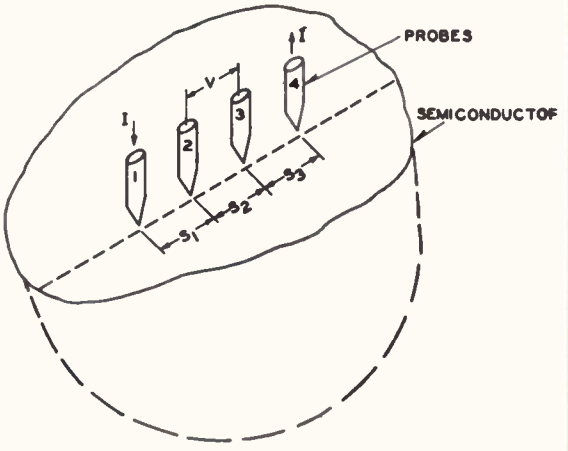

Fig. 1—Model for the four probe resistivity measurements.

The experimental circuit used for measurements is illustrated schematically in Fig. 2. A nominal value of probe spacing which has been found satisfactory is an equal distance of 0.050 inch between adjacent probes. This permit measurement with reasonable currents of *n*- or *p*-type germanium from 0.001 to 50 ohm-cm.

The simple case of four probes on a semi-infinite volume of germanium, which has been solved previously by W. Shockley and others,[5] is repeated here for complete-

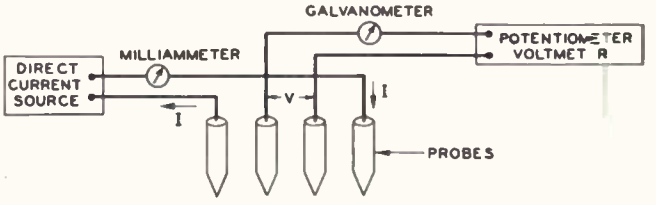

Fig. 2—Circuit used for resistivity measurements.

* Decimal classification: R282.12. Original manuscript received by the Institute, March 26, 1953; revised manuscript received, August 14, 1953.

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

[1] L. B. Valdes, "Measurement of minority carrier lifetime in germanium," PROC. I.R.E., vol. 40, pp. 1420–1423; November, 1952.

[2] L. B. Valdes, "Effect of electrode spacing on the equivalent base resistance of point-contact transistors," PROC. I.R.E., vol. 40, pp. 1429–1434; November, 1952.

[3] J. Bardeen, "Theory of relation between hole concentration and characteristics of germanium point contacts," *Bell Sys. Tech. Jour.*, vol. 29, pp. 469–495; October, 1950.

[4] W. Schockley, G. L. Pearson, J. R. Haynes, "Hole injection in germanium-quantitative studies and filamentary transistors," *Bell Sys. Tech. Jour.*, vol. 28, pp. 344–366; July, 1949.

[5] The author has been informed that this method is the same as used in earth resistivity measurements. Some of the more pertinent references in that field are:

(a) F. Ollendorff, "Erdstrome," Julius Springer, Berlin, Germany; 1928.

(b) J. Riordan and E. D. Sunde, "Mutual impedance of grounded wires for horizontally stratified two-layer earth," *Bell Sys. Tech. Jour.*, vol. 12, pp. 162–177; April, 1933.

(c) E. D. Sunde, "Earth Conduction Effects in Transmission Systems," D. Van Nostrand Co., Inc., New York, N. Y., pp. 47–51; 1949.

ness. Three cases of plane boundaries parallel and perpendicular to the surface where the measurement is made are solved for both conducting and nonconducting boundaries.

In order to use this four probe method in germanium crystals or slices it is necessary to assume that:

1. The resistivity of the material is uniform in the area of measurement.
2. If there is minority carrier injection into the semiconductor by the current-carrying electrodes most of the carriers recombine near the electrodes so that their effect on the conductivity is negligible. (This means that the measurements should be made on surfaces which have a high recombination rate, such as mechanically lapped surfaces.)
3. The surface on which the probes rest is flat with no surface leakage.
4. The four probes used for resistivity measurements contact the surface at points that lie in a straight line.
5. The diameter of the contact between the metallic probes and the semiconductor should be small compared to the distance between probes.
6. The boundary between the current-carrying electrodes and the bulk material is hemispherical and small in diameter.
7. The surfaces of the germanium crystal may be either conducting or nonconducting.

(a) A conducting boundary is one on which a material of much lower resistivity than germanium (such as copper) has been plated.

(b) A nonconducting boundary is produced when the surface of the crystal is in contact with an insulator.

The derivation of equations is considered in the appendix. Only the final results for each of the cases are presented here.

Case 1. Resistivity Measurements on a Large Sample

The model for a semi-infinite volume of material is in Fig. 1. This is approximated by a large sample, such as a crystal or part of it. The resistivity is computed as

$$\rho = \frac{V}{I}\,\frac{2\pi}{\left(\frac{1}{s_1}+\frac{1}{s_3}-\frac{1}{s_1+s_2}-\frac{1}{s_2+s_3}\right)} \qquad (1)$$

where

V = floating potential difference between the inner probes, volts
I = current through the outer pair of probes, amps
s_1 s_2, s_3 = point spacing, in cm
ρ = resistivity in ohm-cm.

When $s_1 = s_2 = s_3 = s$ (1) simplifies to:

$$\rho = \frac{V}{I}\,2\pi s. \qquad (2)$$

Case 2. Resistivity Probes Perpendicular to a Nonconducting Boundary

The model is in Fig. 3. The boundary is perpendicular to the surface where the measurement is made and is also nonconducting. The probes are perpendicular to the boundary. The resistivity may be calculated from

$$\rho = \rho_0 F_2\left(\frac{l}{s}\right) \qquad (3)$$

where ρ_0 is computed from (2) and $F_2(l/s)$ is plotted in Fig. 4. The distance l between the nearest probe and the boundary and the point spacing s are defined in Fig. 3.

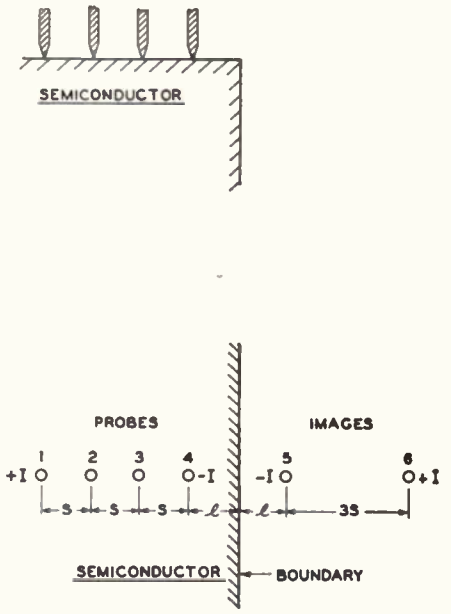

Fig. 3—Resistivity probes perpendicular to a boundary.

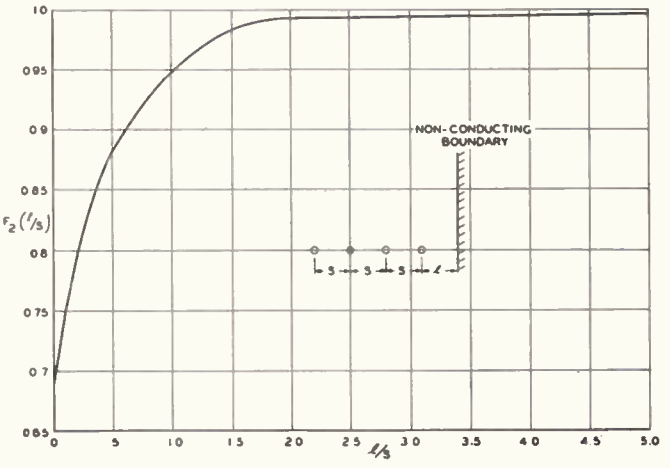

Fig. 4—Correction factor for probes perpendicular to a nonconducting boundary.

The resistivity ρ_0 may be computed from (1) if the point spacings differ but are approximately equal to within 5 per cent.

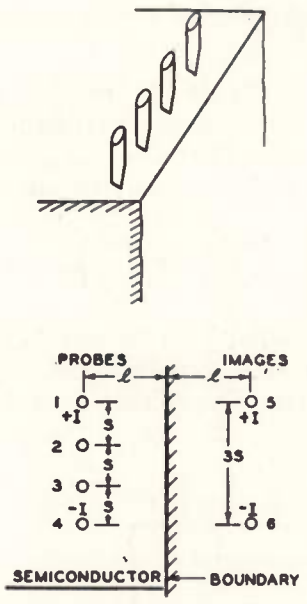

Fig. 5—Resistivity probes parallel to a boundary.

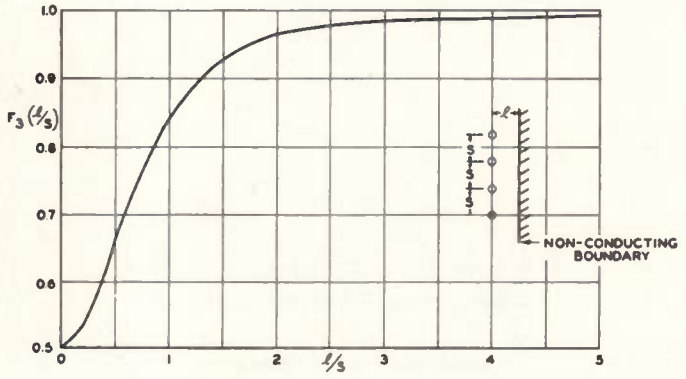

Fig. 6—Correction factor for probes parallel to a nonconducting boundary.

Case 3. Resistivity Probes Parallel to a Nonconducting Boundary

The model is given in Fig. 5. It is the same as for Case 2 except that the probes are parallel to the boundary. The resistivity is

$$\rho = \rho_0 F_3\left(\frac{l}{s}\right) \tag{4}$$

where ρ_0 is defined as before and $F_3(l/s)$ is plotted in Fig. 6.

Case 4. Resistivity Probes Perpendicular to a Conducting Boundary

This differs from Case 2 only in that the boundary is a good conductor of electricity. Such boundary might be obtained by plating that face of the semiconductor with a metal such as copper. The model of Fig. 3 essentially describes the geometry of this case. The resistivity is

$$\rho = \rho_0 F_4\left(\frac{l}{s}\right) \tag{5}$$

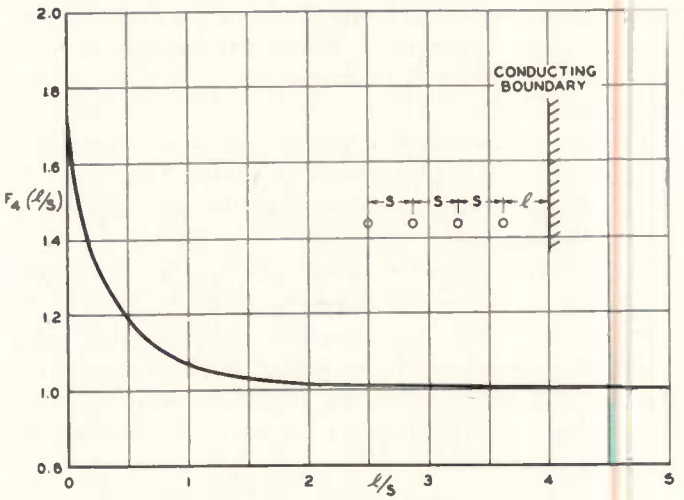

Fig. 7—Correction factor for probes perpendicular to a conducting boundary.

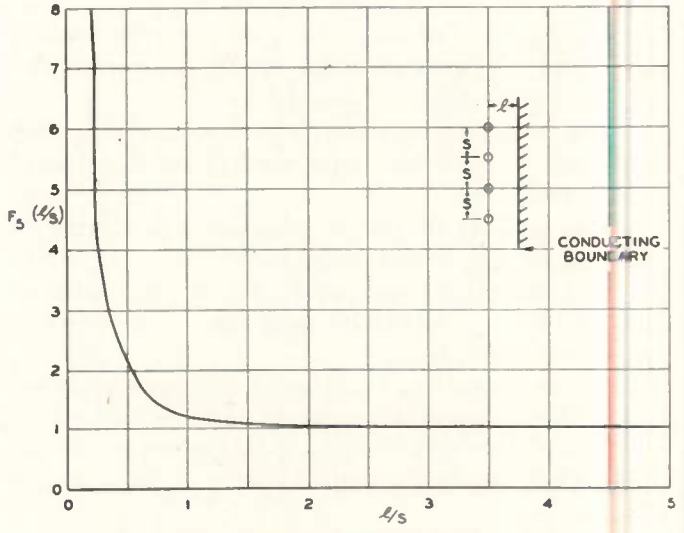

Fig. 8—Correction factor for probes parallel to a conducting boundary.

where ρ_0 is defined as before and $F_4(l/s)$ is plotted in Fig. 7.

Case 5. Resistivity Probes Parallel to a Conducting Boundary

The same model of Fig. 5 is applicable except for the conducting type of boundary. The resistivity is

$$\rho = \rho_0 F_5\left(\frac{l}{s}\right) \tag{6}$$

and $F_5(l/s)$ is plotted in Fig. 8.

Case 6. Resistivity Measurements on a Thin Slice-Conducting Bottom Surface

Fig. 9 shows the resistivity probes on a die of material. If the side boundaries are adequately far from the probes the die may be considered to be identical to a slice. For this case of a slice of thickness w and with a conducting bottom surface the resistivity is computed

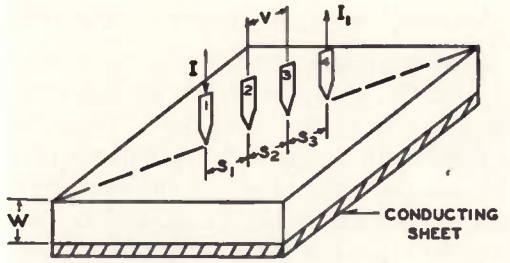

Fig. 9—Resistivity measurements on a thin slice.

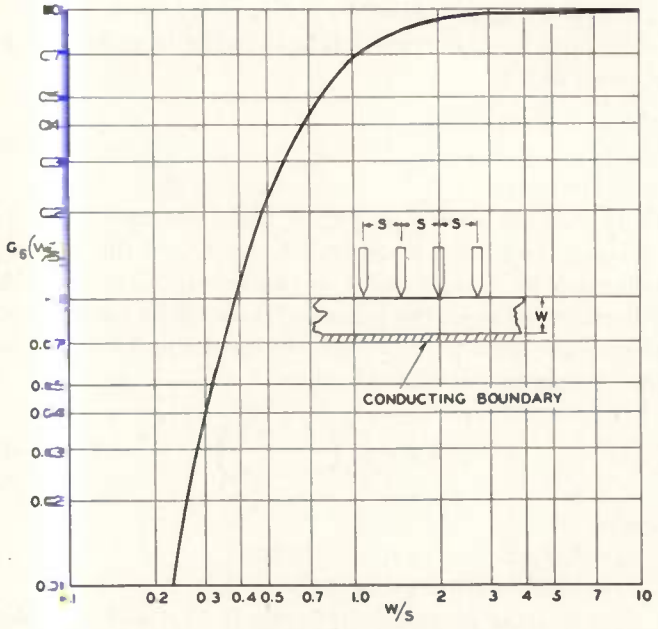

Fig. 10—Correction divisor for probes on a thin slice with a conducting bottom surface.

by means of the divisor $G_6(w/s)$ of Fig. 10 as:

$$\rho = \frac{\rho_0}{G_6\left(\frac{w}{s}\right)}. \qquad (7)$$

This method is not recommended for w/s very small.

Case 5. Resistivity Measurements on a Thin Slice-Nonconducting Bottom Surface

For the case of a nonconducting bottom on a slice like that of Case 6, the resistivity is computed from

$$\rho = \frac{\rho_0}{G_7\left(\frac{w}{s}\right)}. \qquad (8)$$

The function $G_7(w/s)$ is shown in Fig. 11 and ρ_0 is obtained as defined previously from either (1) or (2).

Results

An experimental check has been obtained from Cases 2 and 3. These are the two cases where a nonconducting side boundary was considered. Case 2 is for the probes perpendicular to the boundary and the agreement between theory and experiment is shown in Fig. 12. Case 3 assumes the probes to be parallel to the boundary and the results are in Fig. 13. Both of these curves show the uncorrected value of resistivity ρ_0 calculated from (1) at different ratios l/s. These values are shown as circled dots. The corrected experimental values ρ obtained by means of (3) and (4) and of Figs. 4 and 6 are shown by crosses. Using the mean value of ρ smooth curves have been drawn to indicate the way that ρ_0 should vary if the material is of constant resistivity and there is almost perfect agreement between theory and experiment. All experimental values shown on these curves represent the average of four readings taken with probes spaced about 0.050 inch apart.

The effect of the current flowing through the outer probes on the measured resistivity has been investigated by one set of readings taken on a 6.3 ohm-cm sample. The results are plotted in Fig. 14. The 37 per cent reduction in resistivity at 100 ma is believed to be

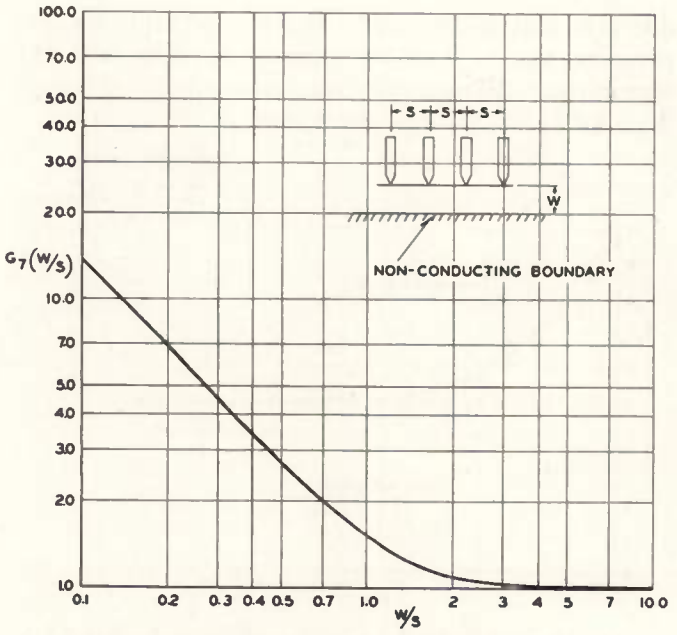

Fig. 11—Correction divisor for probes on a thin slice with a nonconducting bottom surface.

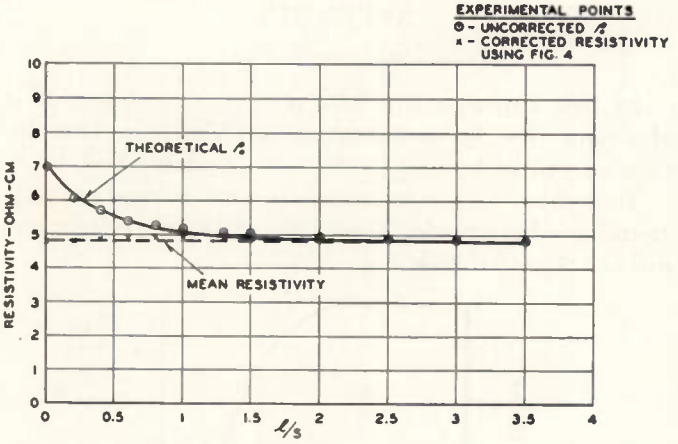

Fig. 12—Experimental check for resistivity probes perpendicular to a nonconducting boundary.

due primarily to heating of the sample, since its temperature was at least 30° C above ambient with this high current. All resistivity measurements are ordinarily done with 1 ma through the probes.

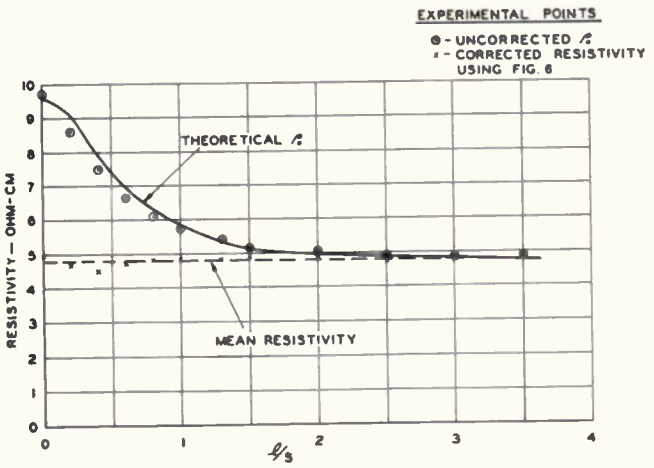

Fig. 13—Experimental check for resistivity probes parallel to a nonconducting boundary.

APPENDIX

Deviation of Formulas

In this appendix are derived the formulas used in this paper for the computation of resistivity. This discussion will be limited to the four point method.

To treat the various models several general assumptions have been made. These were stated in the text and are not repeated here.

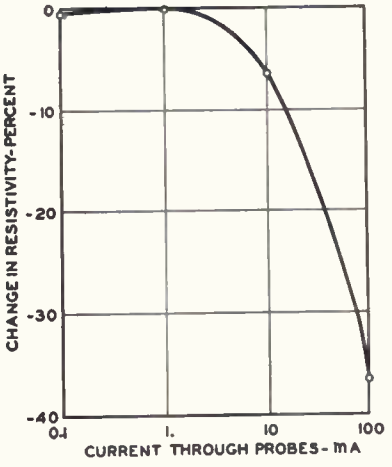

Fig. 14—Effect of current on measured resistivity.

Seven different geometries will be analyzed here. These are: 1. the four point probes on a semi-infinite volume of germanium; 2. the probes near a nonconducting boundary and perpendicular to it; 3. the probes parallel to a nonconducting boundary; 4. the probes perpendicular to a conducting boundary; 5. the probes parallel to a conducting boundary; 6. the probes on a thin slice where the bottom surface is conducting; and 7. the probes on a slice with a nonconducting bottom surface.

Case 1. Resistivity Measurements on a Large Sample

One added boundary condition is required to treat this case, namely, that the probes are far from any of the other surfaces of the sample and the sample can thus be considered a semi-infinite volume of uniform resistivity material. Fig. 1 shows the geometry of this case. Four probes are spaced s_1, s_2, and s_3 apart. Current I is passed through the outer probes (1 and 4) and the floating potential V is measured across the inner pair of probes 2 and 3.

The floating potential V_f a distance r from an electrode carrying a current I in a material of resistivity ρ is given by[6]

$$V_f = \frac{\rho I}{2\pi r}. \tag{9}$$

In the model shown in Fig. 1 there are two current-carrying electrodes, numbered 1 and 4, and the floating potential V_f, at any point in the semiconductor is the difference between the potential induced by each of the electrodes, since they carry currents of equal magnitude but in opposite directions. Thus:

$$V_f = \frac{\rho I}{2\pi}\left(\frac{1}{r_1} - \frac{1}{r_4}\right) \tag{10}$$

where:

r_1 = distance from probe number 1.

r_4 = distance from probe number 4.

The floating potentials at probe (2) V_{f_2} and at probe (3) V_{f_3} can be calculated from (10) by substituting the proper distances as follows:

$$V_{f_2} = \frac{\rho I}{2\pi}\left(\frac{1}{s_1} - \frac{1}{s_2 + s_3}\right) \tag{11}$$

$$V_{f_3} = \frac{\rho I}{2\pi}\left(\frac{1}{s_1 + s_2} - \frac{1}{s_3}\right). \tag{12}$$

The potential difference V between probes is then

$$V = V_{f_2} - V_{f_3} = \frac{\rho I}{2\pi}\left(\frac{1}{s_1} + \frac{1}{s_3} - \frac{1}{s_2 + s_3} - \frac{1}{s_1 + s_2}\right) \tag{13}$$

and the resistivity ρ is computable as

$$\rho = \frac{V}{I}\,\frac{2\pi}{\left(\frac{1}{s_1} + \frac{1}{s_3} - \frac{1}{s_1 + s_2} - \frac{1}{s_2 + s_3}\right)}. \tag{14}$$

When the point spacing is equal, that is $s_1 = s_2 = s_3 = s$, the above simplifies to

$$\rho = \frac{V}{I} 2\pi s. \tag{15}$$

[6] L. B. Valdes, "Effect of electrode spacing on the equivalent base resistance of point-contact transistors," Proc. I.R.E., vol. 40, pp. 1429–1434, eq. (26); November, 1952.

Case 2. Resistivity Probes Perpendicular to a Nonconducting Boundary

Consider as a boundary condition that the reflecting (nonconducting) boundary is a plane perpendicular to the plane which describes the surface of the material on which the resistivity measurements are done. For this case the line on which the resistivity probes lie is assumed to be perpendicular to the line which corresponds to the intersection of the planes of the top surface and the boundary.

The relation between measured voltage and current in the four point probes and the resistivity of the material can be obtained from the model of Fig. 3. Images of the current sources are used and the floating potential at the potential probes is calculated as before. If the boundary is reflecting (nonconducting) the images are current sources of the same sign. Equal spacing between probes is assumed for simplicity.

The floating potential V_{f_2} at probe (2) is given by

$$V_{f_2} = \frac{\rho I}{2\pi}\left(\frac{1}{s} - \frac{1}{2s} - \frac{1}{2l+2s} + \frac{1}{2l+5s}\right). \quad (16)$$

Similarly the floating potential V_{f_3} at probe (3) can be obtained and the difference is

$$V = \frac{\rho I}{2\pi s}\left(1+\frac{s}{2l+s} - \frac{s}{2l+2s} - \frac{s}{2l+4s} + \frac{s}{2l+5s}\right). \quad (17)$$

The resistivity then is

$$\rho = \frac{V}{I} 2\pi s \frac{1}{\left(1+\frac{s}{2l+s} - \frac{s}{2l+2s} - \frac{s}{2l+4s} + \frac{s}{2l+5s}\right)} \quad (18)$$

or

$$\rho = \rho_0 F_2\left(\frac{l}{s}\right) \quad (19)$$

where ρ_0 is the resistivity computed from (15) and $F_2(l/s)$ is the function plotted in Fig. 4.

TABLE I

l/s	$F_2(l/s)$	$F_3(l/s)$	$F_4(l/s)$	$F_5(l/s)$
1	0.69	0.5	1.82	∞
2	0.79	0.533	1.365	8.07
5	0.882	0.658	1.182	2.08
10	0.947	0.842	1.060	1.232
20	0.992	0.965	1.010	1.038
30	0.996	0.997	1.004	1.003
100	0.9995	0.9996	0.0005	1.0004

The resistivity ρ_0 may be obtained from (14) when the probe spacings s_1, s_2, and s_3 are approximately equal. Table I shows the values of $F_2(l/s)$ used for plotting Fig. 4. The function was calculated from

$$F_2\left(\frac{l}{s}\right) = \frac{1}{1+\frac{1}{1+2l/s} - \frac{1}{2+2l/s} - \frac{1}{4+2l/s} + \frac{1}{5+2l/s}}. \quad (20)$$

Case 3. Resistivity Probes Parallel to a Nonconducting Boundary

The boundary of the surface on which the measurements are made is assumed to lie on a nonconducting plane perpendicular to the surface. The four probes are assumed to lie in a line parallel to the intersection of the two planes.

The model for this case is in Fig. 5. As in the previous case, the images are current sources of the same sign since the boundary is reflecting or nonconducting. The floating potentials V_{f_2} and V_{f_3} at probes (2) and (3) are identical in form so that the difference is

$$V = V_{f_2} - V_{f_3} = 2V_{f_2} \quad (21)$$

and

$$\rho = \frac{V}{I} 2\pi s \frac{1}{2\left(\frac{1}{2} + \frac{s}{\sqrt{s^2+(2l)^2}} - \frac{s}{\sqrt{(2s)^2+(2l)^2}}\right)} \quad (22)$$

or

$$\rho = \rho_0 F_3\left(\frac{l}{s}\right). \quad (23)$$

The resistivity ρ_0 can be computed from (14) or (15) and

$$F_3\left(\frac{l}{s}\right) = \frac{1}{\left[1+\frac{2}{\sqrt{1+\left(\frac{2l}{s}\right)^2}} - \frac{1}{\sqrt{1+\left(\frac{l}{s}\right)^2}}\right]}. \quad (24)$$

This function is tabulated in Table I and plotted in Fig. 6.

Case 4. Resistivity Probes Perpendicular to a Conducting Boundary

The same geometry as in Case 2 is assumed here, except that the boundary is conducting and the sign of the images has to be reversed. Fig. 3 serves as a model, except that image (5) is positive and image (6) is negative. Then

$$\rho = \frac{V}{I} 2\pi s \frac{1}{\left(1-\frac{s}{2l+s} + \frac{s}{2l+2s} + \frac{s}{2l+4s} - \frac{s}{2l+5s}\right)} \quad (25)$$

and

$$F_4\left(\frac{l}{s}\right) = \frac{1}{1-\frac{1}{1+\frac{2l}{s}} + \frac{1}{2+\frac{2l}{s}} + \frac{1}{4+\frac{2l}{s}} - \frac{1}{5+\frac{2l}{s}}}. \quad (26)$$

The function $F_4(l/s)$ is tabulated in Table I and plotted in Fig. 7.

Case 5. Resistivity Probes Parallel to a Conducting Boundary

The same geometry as in Case 3 is assumed here, except that the boundary is conducting and the sign of the images must be reversed. Therefore, Fig. 5 is the model with image (5) negative and image (6) positive. From this

$$\rho=\frac{V}{I}\,2\pi s\,\frac{1}{2\left(\frac{1}{2}-\frac{s}{\sqrt{s^2+(2l)^2}}+\frac{s}{\sqrt{(2s)^2+(2l)^2}}\right)}. \tag{27}$$

The function

$$F_5\left(\frac{l}{s}\right)=\frac{1}{\left[1-\frac{2}{\sqrt{1+\left(\frac{2l}{s}\right)^2}}+\frac{1}{\sqrt{1+\left(\frac{l}{s}\right)^2}}\right]} \tag{28}$$

is tabulated in Table I and plotted in Fig. 8.

Case 6. Resistivity Measurements on a Thin Slice-Conducting Bottom Surface

Two boundary conditions must be met on this case; the top surface of the slice must be a reflecting (non-conducting) surface and the bottom surface must be an absorbing (conducting) surface. Since the two boundaries are parallel a solution by the method of images requires for each current source an infinite series of images along a line normal to the planes and passing through the current source.

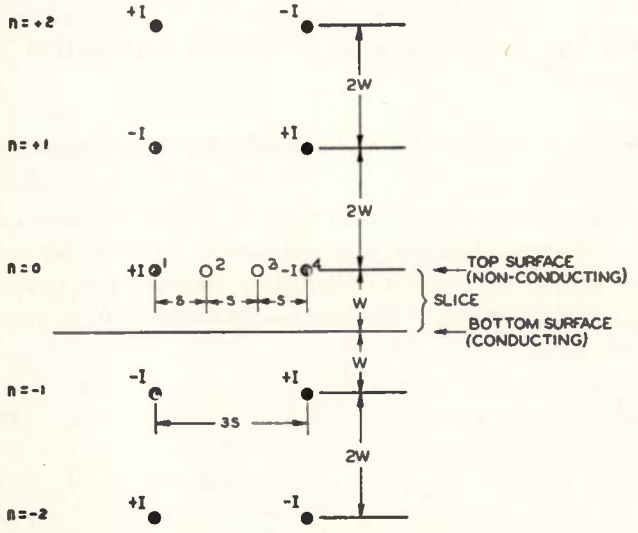

Fig. 15—Images for the case of the resistivity probes on a thin slice with a conducting bottom surface.

The model for this case is shown in Fig. 9. The side surfaces of the die are assumed to be far from the area of measurement and, therefore, only the effect of the bottom surface needs to be considered. In this analysis equal probe spacing s shall be assumed. The width of the slice is w. The array of images needed is indicated in Fig. 15, where the polarity and spacings of the first few images are as shown.

The floating potential V_{f_2} at electrode (2) is:

$$V_{f_2}=\frac{\rho I}{2\pi}\left[\sum_{n=-\infty}^{n=\infty}(-1)^n\frac{1}{\sqrt{s^2+(2nw)^2}}-\sum_{n=-\infty}^{n=\infty}(-1)^n\frac{1}{\sqrt{(2s)^2+(2nw)^2}}\right]. \tag{29}$$

Likewise, the floating potential at electrode (3) can be obtained and

$$V=\frac{\rho I}{2\pi}\left[\frac{1}{s}+\sum_{n=1}^{n=\infty}(-1)^n\frac{4}{\sqrt{s^2+(2nw)^2}}-\sum_{n=1}^{n=\infty}(-1)^n\frac{4}{\sqrt{(2s)^2+(2nw)^2}}\right]. \tag{30}$$

The resistivity then becomes

$$\rho=\frac{\rho_0}{G_6\left(\frac{w}{s}\right)}. \tag{31}$$

Where resistivity ρ_0 is computable from (15), and (14) can be used if the point spacings are different, but approximately equal. The function $G_6(w/s)$ is computed from[7]

$$G_6\left(\frac{w}{s}\right)=1+4\,\frac{s}{w}\sum_{n=1}^{n=\infty}(-1)^n\left\{\frac{1}{\sqrt{\left(\frac{s}{w}\right)^2+(2n)^2}}-\frac{1}{\sqrt{\left(2\,\frac{s}{w}\right)^2+(2n)^2}}\right\}, \tag{32}$$

which is tabulated in Table II and plotted in Fig. 10.

TABLE II

W/S	$G_6(W/S)$	$G_7(W/S)$
0.100	0.0000019	13.863
0.141	0.00018	9.704
0.200	0.00342	6.931
0.333	0.0604	4.159
0.500	0.228	2.780
1.000	0.683	1.504
1.414	0.848	1.223
2.000	0.933	1.094
3.333	0.983	1.0228
5.000	0.9948	1.0070
10.000	0.9993	1.00045

[7] A neat way of summing the series has been suggested by A. Uhlir, to be published.

Case 7. Resistivity Measurements on a Thin Slice-Nonconducting Bottom Surface

The model for these measurements is like for Case 6, except that the bottom surface of the slice is nonconducting. This means that all the images of Fig. 15 have the same charge as the current source. Thus all the images on a row have equal charges and (30) describes the potential difference across the inner pair of probes if $(-1)^n$ is removed from the equation. Then,

$$\rho = \frac{\rho_0}{G_7\left(\frac{w}{s}\right)} \tag{33}$$

where

$$G_7\left(\frac{w}{s}\right) = 1 + 4\frac{s}{w}\sum_{n=1}^{n=\infty}\left\{\frac{1}{\sqrt{\left(\frac{s}{w}\right)^2 + (2n)^2}} - \frac{1}{\sqrt{\left(2\frac{s}{w}\right)^2 + (2n)^2}}\right\}. \tag{34}$$

This function $G_7(w/s)$ is tabulated in Table II and plotted in Fig. 11. For smaller values of w/s the function $G_7(w/s)$ approaches the case for an infinitely thin slice, or

$$G_7\left(\frac{w}{s}\right) = \frac{2s}{w}\ln 2. \tag{35}$$

ACKNOWLEDGMENT

This paper presents the curves needed to compute resistivity under a large number of different experimental conditions. The accuracy of the results is better than 2 per cent in practically all cases. The assistance of F. R. Keene, who carried out most of the computations of tables and who obtained the experimental results, is gratefully acknowledged. The author is also indebted to A. Uhlir, W. J. Pietenpol, and R. M. Ryder.

A General RLC Synthesis Procedure*

LOUIS WEINBERG†

The following paper is published on the recommendation of the IRE Professional Group on Circuit Theory. —*The Editor.*

Summary—**Any physically realizable RLC transfer function—impedance, admittance, or dimensionless ratio—can be realized within a multiplicative constant by the synthesis procedures presented in this paper. The form of network achieved is a lattice with these significant features: (a) it may have any desirable termination; (b) it contains no mutual inductance; (c) every inductance in the network appears with an associated series resistance so that, in building the network, low-Q coils may be used.**

In addition, the lattice arms are of so simple a form relative to each other that many of the achieved lattices are amenable to reduction to unbalanced networks. Further, in case of a transfer admittance, reduction can be achieved with the use at most of real transformers i.e., transformers with winding resistance, finite magnetizing inductance, and a coupling coefficient smaller than one.

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I. INTRODUCTION

THERE IS A wide variety of existing synthesis procedures, but as anyone conversant with the synthesis field fully realizes much remains to be done. The inadequacy of available procedures shows up particularly in a broad field of communications, namely, synthesis for prescribed transient response.[1] In this synthesis both magnitude and phase are important, so that the methods for realizing a prescribed magnitude of transfer function are inapplicable. Up to the present time one of the principal procedures that could be used for the realization of both minimum-phase and nonminimum-phase transfer functions has been the one that yields a constant-resistance lattice. This type of lattice suffers from many disadvantages. In general each of the

[1] D. F. Tuttle, Jr.: "Network Synthesis for Prescribed Transient Response," Sc.D. thesis in electrical engineering, M.I.T., Cambridge, Mass.; 1948.

arms requires close-coupled or lossless coils. An important disadvantage, too, for those cases in which an unbalanced form of network is definitely preferable, is that the series and cross arms are so complicated relative to each other that without the use of ideal transformers reduction to an unbalanced form is virtually impossible.

The lattice-synthesis procedures treated in this paper realize a given transfer function within a multiplicative constant. No restriction other than physical realizability is placed on the function to be realized. Among the advantages claimed for the final lattice are that it contains no mutual inductance and all its coils are lossy, i.e., every inductance may be associated with a series resistance. In addition, the arms are of so simple a form as to render the lattice amenable to reduction to an unbalanced network. For the case of a transfer admittance, moreover, reduction can *always* be achieved with the use at most of real transformers, i.e., transformers with winding resistance, finite magnetizing inductance, and a coupling coefficient smaller than one.

The dimensions of the transfer function to be realized depend, of course, on the type of system in which the synthesized network is to be used. If the driving force, for example, is a pentode which approximates a current source, a transfer impedance is needed. On the other hand, a transfer admittance is necessary for a voltage-source drive like a cathode follower and a current output. Finally a ratio of output to input voltage calls for the synthesis of a dimensionless transfer function. The procedures show how each of these functions may be realized with the practical terminations of an open circuit, a resistance, or a parallel resistance-capacitance combination.

Fig. 1—Two terminal-pair network.

II. The Synthesis Problem Presented

In Fig. 1 is shown a general passive, two terminal-pair network which may be completely characterized by the pair of simultaneous equations

$$\begin{aligned} I_1 &= y_{11}E_1 + y_{12}E_2 \\ I_2 &= y_{21}E_1 + y_{22}E_2, \end{aligned} \tag{1}$$

or by the inverse set

$$\begin{aligned} E_1 &= z_{11}I_1 + z_{12}I_2 \\ E_2 &= z_{21}I_1 + z_{22}I_2, \end{aligned} \tag{2}$$

where $z_{12}=z_{21}$ and $y_{12}=y_{21}$ by reciprocity, and the y's and z's are the familiar short-circuit admittances and open-circuit impedances, respectively. Useful relations may be derived from the above basic equations. For example, for a pure resistance load of one ohm, since numerically $E_2=-I_2$, the second of (1) yields

$$Y_{12} \equiv \frac{I_2}{E_1} = \frac{y_{12}}{1+y_{22}}, \tag{3}$$

an analogous relation being derived for Z_{12} from the second of (2).

For the load an open circuit, that is, with $I_2=0$, (2) yields

$$K \equiv \frac{E_2}{E_1} = \frac{z_{12}}{z_{11}}; \tag{4}$$

while for a short-circuit load, from (1) may be obtained

$$\frac{I_2}{I_1} = \frac{y_{12}}{y_{11}}. \tag{5}$$

In this paper the two terminal-pair network is particularized to the lattice shown in Fig. 2, for which

$$\begin{aligned} z_{11} &= \tfrac{1}{2}(Z_b + Z_a) \\ z_{12} &= \tfrac{1}{2}(Z_b - Z_a), \end{aligned} \tag{6}$$

an analogous set of relations holding true for the y's.

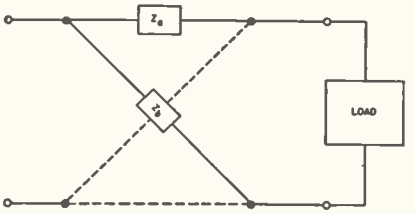

Fig. 2—RLC lattice network.

Thus, with the substitution of (6), (4) becomes

$$K = \frac{Z_b - Z_a}{Z_b + Z_a}, \tag{7}$$

whereas substitution of the y's for the lattice transforms (3) to

$$Y_{12} = \frac{\tfrac{1}{2}(Y_b - Y_a)}{1 + \tfrac{1}{2}(Y_b + Y_a)}. \tag{8}$$

It is well known, furthermore, that the general form for any of the transfer functions is a rational function given by the quotient of polynomials

$$\begin{aligned} &\frac{a_n s^n + a_{n-1}s^{n-1} + \cdots + a_0}{b_m s^m + b_{m-1}s^{m-1} + \cdots + b_0} \\ &\quad = \frac{(s-s_1)(s-s_3)\cdots}{\dfrac{b_m}{a_n}(s-s_2)(s-s_4)\cdots} \qquad n \leqq m \\ &\quad = \frac{p(s)}{Hq(s)}, \end{aligned} \tag{9}$$

where q must be a Hurwitz polynomial (have all its zeros in the left half-plane), and H is a positive constant.

III. Open-Circuited Lattice Realization for $K = E_2/E_1$

It is desired to realize the given transfer function

$$K = \frac{E_2}{E_1} = \frac{p}{Hq} \tag{10}$$

as an open-circuited lattice with as large a gain (that is, as small an H) as possible. The given function may therefore be set to (7); this yields

$$K = \frac{p}{Hq} = \frac{Z_b - Z_a}{Z_b + Z_a}. \tag{11}$$

We break q into the sum of two polynomials, so that

$$q = q_1 + Aq_1', \tag{12}$$

where q_1 is a Hurwitz polynomial, A is a positve real constant, and q_1' is the derivative of q_1. This can always be done as was previously demonstrated.[2] After dividing the numerator and denominator of the resulting K by q_1 to obtain

$$K = \frac{\dfrac{p}{q_1}}{H\left(1 + \dfrac{Aq_1'}{q_1}\right)}, \tag{13}$$

we expand p/q_1 into partial fractions. Its residues are in general positive or negative real for real poles, complex for complex poles. A similar expansion of Aq_1'/q_1 makes the total denominator of K

$$H\left(1 + \frac{Aq_1'}{q_1}\right) = H\left(k_0^{(d)} + \frac{k_1^{(d)}}{s - s_1} + \frac{k_2^{(d)}}{s - s_2} + \cdots\right), \tag{14}$$

where $k_0^{(d)} = 1$, and all the residues $k_\mu^{(d)}$ for $\mu \neq 0$ are equal to A. If $(Z_b - Z_a)$ and $(Z_b + Z_a)$ are also thought of as expanded in partial fractions, the residues of like terms of p/q_1 and $(Z_b - Z_a)$ may be equated as may those of $H(1 + Aq_1'/q_1)$ and $(Z_b + Z_a)$. We thus obtain

$$\begin{aligned} &k_\mu^{(b)} - k_\mu^{(a)} = k_\mu^{(n)} \quad (\mu = 0, 1, 2, \cdots, m, \\ &k_\mu^{(b)} + k_\mu^{(a)} = Hk_\mu^{(d)} \quad \text{where } m \text{ is degree of } q), \end{aligned} \tag{15}$$

where the superscripts a, b, n, and d refer respectively to Z_a, Z_b, and the numerator and denominator of K, while the subscript μ designates the poles $s_\mu = -\sigma_\mu + j\omega_\mu$, or $s_\mu = -\sigma_\mu - j\omega_\mu$, or $s_\mu = -\sigma_\mu$ in which σ_μ and ω_μ are positive.

Solving (15) for the unknown Z_a and Z_b residues as indicated in (16) below

$$\begin{aligned} k_\mu^{(b)} &= \tfrac{1}{2}(Hk_\mu^{(d)} + k_\mu^{(n)}) \\ \alpha_\mu^{(b)} + j\beta_\mu^{(b)} &= \tfrac{1}{2}(Hk_\mu^{(d)} + \alpha_\mu^{(n)} + j\beta_\mu^{(n)}) \\ k_\mu^{(a)} &= \tfrac{1}{2}(Hk_\mu^{(d)} - k_\mu^{(n)}) \\ \alpha_\mu^{(a)} + j\beta_\mu^{(a)} &= \tfrac{1}{2}(Hk_\mu^{(d)} - \alpha_\mu^{(n)} - j\beta_\mu^{(n)}) \end{aligned} \tag{16}$$

finally yields

$$\begin{array}{ll} \mu \neq 0 & \mu \neq 0 \\ \alpha_\mu^{(b)} = \tfrac{1}{2}(HA + \alpha_\mu^{(n)}) & \alpha_\mu^{(a)} = \tfrac{1}{2}(HA - \alpha_\mu^{(n)}) \\ \beta_\mu^{(b)} = \tfrac{1}{2}\beta_\mu^{(n)} & \beta_\mu^{(a)} = -\tfrac{1}{2}\beta_\mu^{(n)} \\ \underline{\mu = 0} & \underline{\mu = 0} \\ \alpha_0^{(b)} = \tfrac{1}{2}(H + \alpha_0^{(n)}) & \alpha_0^{(a)} = \tfrac{1}{2}(H - \alpha_0^{(n)}). \end{array} \tag{17}$$

For negative real poles the requirement that the residues, $\alpha_\mu^{(a)}$ and $\alpha_\mu^{(b)}$, be real and positive when used in conjunction with the above (17), gives as the condition to be satisfied for $\mu \neq 0$

$$-1 \leqq \frac{\alpha_\mu^{(n)}}{HA} \leqq 1, \tag{18}$$

and for $\mu = 0$ we substitute the constant unity for A. This, of course, is the same as the condition that arises in the Bower-Ordung RC synthesis,[3] since for an RC lattice the poles must all be real. In the general synthesis considered here, however, the complex poles must also be provided for. The real parts of the residues in these complex poles must not only be positive, but must also be equal to or greater than a positive constant c_μ which is defined below. This is seen by application to the residues of Z_a of the condition for realizability that is derived in Appendix I (with a similar result holding for Z_b):

$$\frac{|\beta_\mu^{(a)}|}{\alpha_\mu^{(a)}} \leqq \frac{\sigma_\mu}{\omega_\mu} \tag{19}$$

or

$$\begin{aligned} \alpha_\mu^{(a)} &\geqq \frac{|\beta_\mu^{(a)}|\,\omega_\mu}{\sigma_\mu} = c_\mu \text{ (positive constant)} \\ &\geqq \frac{\tfrac{1}{2}|\beta_\mu^{(n)}|\,\omega_\mu}{\sigma_\mu} = c_\mu. \end{aligned} \tag{20}$$

When the constant c_μ is substituted in those relations of (17) for which $\mu \neq 0$, the conditions to be satisfied become

$$1 \geqq \frac{2c_\mu - \alpha_\mu^{(n)}}{HA} \tag{21}$$

and

$$1 \geqq \frac{2c_\mu + \alpha_\mu^{(n)}}{HA}. \tag{22}$$

[2] L. Weinberg: "Synthesis of Unbalanced RLC Networks," *Proc. National Electronics Conference*, vol. 8; 1952; also published in *Jour. Appl. Phys.*, pp. 300–306; March, 1953.

[3] J. L. Bower and P. F. Ordung: "The synthesis of resistor-capacitor networks," Proc. I.R.E., pp. 263–269; March, 1950.

We need satisfy only the stronger of the above two inequalities for any specific complex pole. If $\alpha_\mu^{(n)}$ is positive, (22) is the stronger and must be used to determine the minimum value of H; if $\alpha_\mu^{(n)}$ is negative, we use (21). Therefore, to summarize the two steps for the complex poles, we must determine first the c_μ for each pole and then the value of H necessary to satisfy the stronger of (21) and (22).

By satisfaction also of (18) for the real poles we may thus tabulate the necessary value of H for each pole. In doing this we may use the equals sign in (18), (21), and (22); then we choose a value of *H greater than* the largest required value, which automatically guarantees the satisfaction of the condition for each pole with the *inequality* sign. This is necessary, as is pointed out in Appendix I, in order that every inductance appear with an associated resistance and that each of the partial-fraction components (complex conjugate poles taken in pairs) of

$$Z_a = k_0^{(a)} + \frac{k_1^{(a)}}{s - s_1} + \frac{k_2^{(a)}}{s - s_2} + \cdots + \frac{k_m^{(a)}}{s - s_m}$$

and

$$Z_b = k_0^{(b)} + \frac{k_1^{(b)}}{s - s_1} + \frac{k_2^{(b)}}{s - s_2} + \cdots + \frac{k_m^{(b)}}{s - s_m} \tag{23}$$

be positive real. Then Z_a and Z_b may be realized by inspection in the Foster manner for two-element kind networks; and as Appendix I demonstrates, every inductance will have an associated series resistance.

One final point, useful in the subsequent procedures, is made regarding the constant term in (23). As is obvious from (17) for $\mu = 0$, we *always* obtain this constant term. Furthermore, if the degree of p is less than that of q, then

$$\alpha_0^{(b)} = \alpha_0^{(a)} = \tfrac{1}{2} H, \tag{24}$$

while if the degrees of p and q are equal,

$$\alpha_0^{(b)} = \alpha_0^{(a)} + 1 = \tfrac{1}{2}(H + 1). \tag{25}$$

IV. Procedure Using the Integral of q

An alternate procedure is obtained by making use of the integral of q, designated by $q^{(-1)}$. This method has the drawback that it may not always work; it works only when $q^{(-1)}$ is Hurwitz. It can be easily shown that the integral of a Hurwitz polynomial, unlike the derivative, is not always Hurwitz. It is a simple matter in any particular problem, however, to form the integral, choosing the arbitrary constant conveniently, and then check for Hurwitz character.[4] The advantage of the method is that it generally yields a network with a higher gain and coils of lower Q.

The steps in the procedure for synthesizing an open-circuited lattice begin with

$$K = \frac{E_2}{E_1} = \frac{Z_b - Z_a}{Z_b + Z_a} = \frac{p}{Hq} = \frac{\dfrac{p}{q^{(-1)}}}{\dfrac{Hq}{q^{(-1)}}}. \tag{26}$$

Now $q/q^{(-1)}$ is obviously of the same form as q_1/q_1, if $q^{(-1)}$ is Hurwitz; and the procedure follows along the same lines as the previous one except for a few minor differences. The differences to be noted are that the constant term in the partial-fraction expansions, i.e. for $\mu = 0$, is nonexistent and the constant A is equal to one.

For this method the useful equations that correspond to (17) are

$$\begin{array}{ll} \underline{\mu \neq 0} & \underline{\mu \neq 0} \\ \alpha_\mu^{(b)} = \tfrac{1}{2}(H + \alpha_\mu^{(n)}) & \alpha_\mu^{(a)} = \tfrac{1}{2}(H - \alpha_\mu^{(n)}) \\ \beta_\mu^{(b)} = \tfrac{1}{2}\beta_\mu^{(n)} & \beta_\mu^{(a)} = -\tfrac{1}{2}\beta_\mu^{(n)}, \end{array} \tag{27}$$

which yield for the real poles the condition corresponding to (18),

$$-1 \leqq \frac{\alpha_\mu^{(n)}}{H} \leqq 1, \tag{28}$$

and for the complex poles yield the inequalities that correspond respectively to (21) and (22),

$$1 \geqq \frac{2c_\mu - \alpha_\mu^{(n)}}{H} \tag{29}$$

and

$$1 \geqq \frac{2c_\mu + \alpha_\mu^{(n)}}{H}. \tag{30}$$

The use of the above equations, along with the definition of c_μ given in (20), allows further for a synthesis to be carried out.

V. Realization of Transfer Impedance in Form of Terminated Lattice

Resistance Termination

To obtain the transfer impedance

$$Z_{12} = \frac{E_2}{I_1} = \frac{p}{Hq} \tag{31}$$

as a resistance-terminated lattice, it is possible to make use of the dual form of (3),

$$Z_{12} = \frac{z_{12}}{1 + z_{22}} = \frac{\tfrac{1}{2}(Z_b - Z_a)}{1 + \tfrac{1}{2}(Z_b + Z_a)}. \tag{32}$$

[4] E. A. Guillemin: "The Mathematics of Circuit Analysis," John Wiley & Sons, Inc., New York, N. Y.; 1949.

A procedure similar to that shown for the open-circuited lattice then allows us to make the necessary identifications for direct synthesis of the resistance-terminated lattice. The network obtained, however, is the same as the one obtained by application of the reciprocity theorem and well-known lattice transformations to the open-circuited lattice. For simplicity of demonstration we will therefore consider the method for realizing the open-circuited lattice as the basic one from which the other desirable forms of network are easily derived.

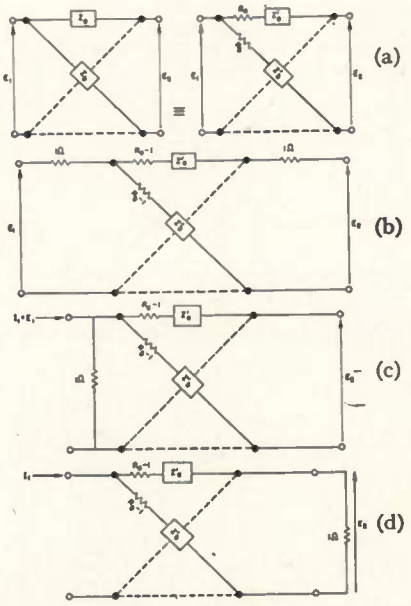

Fig. 3—Steps in the conversion of open-circuited lattice for which $K=E_2/E_1=p/Hq$ to resistance-terminated lattice with $Z_{12}=E_2/I_1=p/Hq$.
(a) Open-circuited lattice for which $K=E_2/E_1=p/Hq$.
(b) Lattice equivalent to that in (a).
(c) Lattice after application of Norton's Theorem.
(d) Lattice given by application of reciprocity theorem, where $Z_{12}=E_2/I_1=p/Hq$.

Since, as observed previously, a series resistance is always present in each arm of the open-circuited lattice, we derive an equivalent lattice[4,5] by removing one ohm from each arm, then convert to a current source by use of Norton's theorem, and finally by use of the reciprocity theorem obtain the desired network. The sequence of steps beginning with the previously realized open-circuited lattice for which

$$K=\frac{E_2}{E_1}=\frac{p}{Hq} \tag{33}$$

is illustrated in Fig. 3. The one-ohm series resistance at the output terminals is omitted in Fig. 3(c) because the output is open-circuited.

An improvement in gain can be effected by removing more than one ohm from each of the arms. Suppose, for the sake of illustration, that p is of lower degree than q so that, as was noted in (24), $R_a=R_b=H/2$. Then we can remove R_a and follow the same sequence of steps as in Fig. 3 to obtain a network terminated in $R_a=H/2$ with the transfer function

$$Z_{12}=\frac{E_2}{I_1}=\frac{1}{2}\,\frac{p}{q}. \tag{34}$$

It is pointed out, finally, that if we stop at the step given by Fig. 3(c), we realize a transfer impedance in the form of an open-circuited lattice, where the shunt resistance is useful in the instrumentation of the network since it may represent the finite internal resistance of the current source.

Parallel RC Termination

In the instrumentation of a practical circuit it is often useful to have a shunt capacitance at the input or output of a network. The immediately preceding method showed how to obtain a resistance termination for the lattice network; this part demonstrates the realization of a parallel RC termination.

The following artifice is restricted in that it can only be used for those transfer functions in which the degree of p is less than the degree of q. Suppose that the q_1 given by the $q=q_1+Aq_1'$ breakdown (similar remarks applying to the $q^{(-1)}$ procedure) has at least one negative real zero given by $(s+a)$. (If it does not and we wish to employ this technique we multiply numerator and denominator of the transfer function by a linear term $(s+b)$ to obtain a new denominator from which q_1 is determined.) Letting the impedance of the desired termination be $k/(s+a)$, we may then write for the RC-terminated lattice (using the additional constant k for convenience):

$$\begin{aligned} Z_{12}&=\frac{E_2}{I_1}=\frac{kp}{Hq}\\ &=\frac{kp}{H(q_1+Aq_1')}=\frac{\dfrac{kp}{q_1}}{H\left(1+\dfrac{Aq_1'}{q_1}\right)}\\ &=\frac{\dfrac{kp}{(s+a)q_2}}{H\left(1+\dfrac{Aq_1'}{q_1}\right)}. \end{aligned} \tag{35}$$

Multiplying the transfer function by $(s+a)/k$ yields

$$\begin{aligned} K&=(s+a)Z_{12}/k\\ &=\frac{\dfrac{p}{q_2}}{H\left(1+\dfrac{Aq_1'}{q_1}\right)}. \end{aligned} \tag{36}$$

[4] E. A. Guillemin: "RC-Coupling Networks," M.I.T. Rad. Lab. Report no. 43; October 11, 1944.

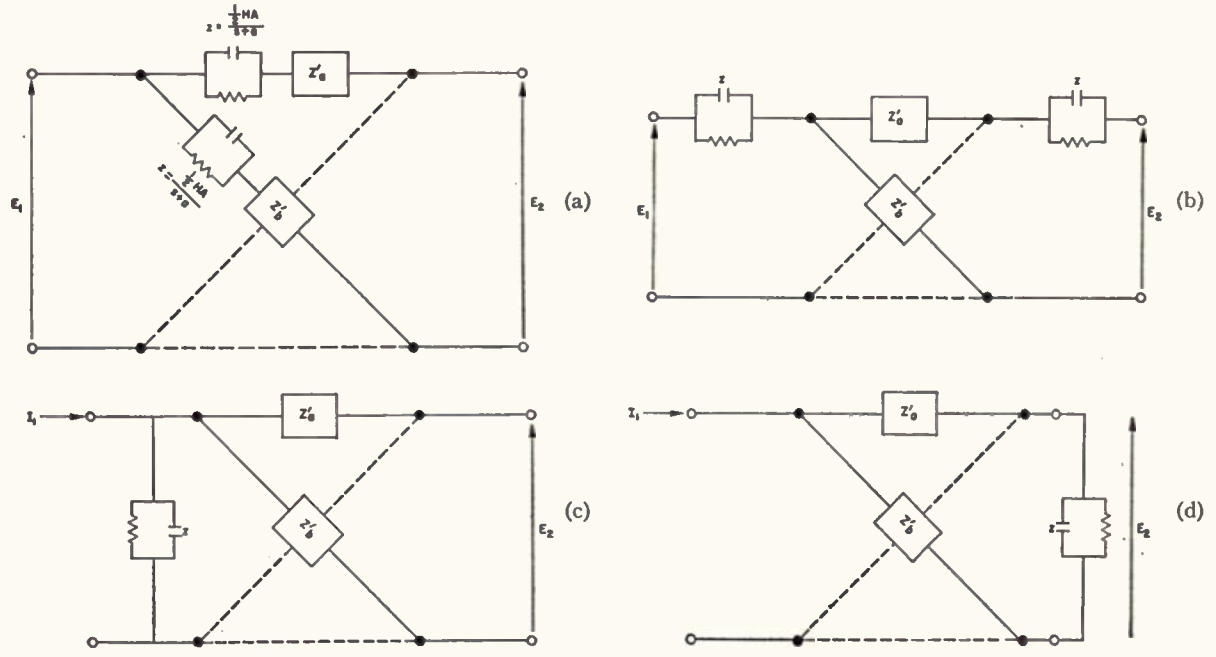

Fig. 4—Steps in the conversion of an open-circuited lattice for which $K=E_2/E_1=(s+a)p(Hq)$ to an RC-terminated lattice for which $Z_{12}=E_2/I_1=Ap/(2q)$.
a) Open-circuited lattice for which $K=(s+a)p/(Hq)=E_2/E_1$.
b) Lattice equivalent to that in (a).
c) Lattice after application of Norton's Theorem, where $I_1=2(s+a)E_1/(HA)$.
d) Lattice after application of reciprocity theorem with $Z_{12}=E_2/I_1=(1/2)HAE_2/(s+a)E_1=(HA/2)K/(s+a)=Ap/(2q)$.

Now this function K is realized as the voltage ratio of an open-circuited lattice by the previously demonstrated method. Since the numerator residue in the pole $s=-a$ is zero, it is clear by inspection of (17) that the residues for Z_a and Z_b in the pole are each equal to $HA/2$. The network which thus has the form shown in Fig. 4(a) can be transformed by removal of the RC combination from each arm, after which the successive applications of Norton's Theorem and reciprocity give

$$Z_{12}=\frac{E_2}{I_1}=\frac{kp}{Hq}=\frac{1}{2}\,\frac{Ap}{q} \tag{37}$$

for the final desired form of network. The pertinent equations and steps are indicated in Fig. 4.

To summarize the procedure, we consider the given function as $Z_{12}=kp/Hq$ (this is what we finally achieve, where $k=1/2(HA)$). We then multiply this by $(s+a)/k$ where $(s+a)$ is a factor of q_1 (or of $q^{(-1)}$, if we are using the integral method). We then synthesize the resulting function as the voltage ratio for an open-circuited lattice, after which the steps illustrated in Fig. 4 will give $Ap/(2q)$ as the desired transfer impedance for the RC-terminated lattice.

VI. Realization of Transfer Admittance as a Resistance-Terminated Lattice

As pointed out previously for the dual case, it is possible by use of the equation

$$Y_{12}=\frac{y_{12}}{1+y_{22}}=\frac{\frac{1}{2}(Y_b-Y_a)}{1+\frac{1}{2}(Y_b+Y_a)} \tag{38}$$

to develop the proper identifications for the synthesis procedure. The simpler approach, however, is to consider the entire procedure for realizing an open-circuited lattice carried over to the dual problem; that is, instead of first synthesizing for a voltage ratio by means of an open-circuited lattice, we now realize

$$\frac{I_2}{I_1}=\frac{y_{12}}{y_{11}} \tag{39}$$

in the form of a short-circuited lattice. The dual of the previous remarks on Z_{12} now apply so that we obtain a final transfer function

$$Y_{12}=\frac{I_2}{E_1}=\frac{1}{2}\,\frac{p}{q} \tag{40}$$

for the same problem treated previously. The final lattice with a resistance termination will be of the form shown in Fig. 5.

In Appendix II, where the general problem of the reduction of lattices to unbalanced forms is taken up, it is shown that the use of *real* transformers *always* allows the lattice obtained for Y_{12} to be converted to an unbalanced network.

VII. Illustrative Example

We desire to synthesize within a constant multiplier the given

$$Y_{12} = \frac{p}{q} = \frac{1}{(s^2 + 0.76536s + 1)(s^2 + 1.84776s + 1)}$$

as a resistance-terminated lattice. The given Y_{12} represents the Butterworth approximation to the normalized low-pass filter with $n=4$, where the Butterworth functions are $B_{2n}(\omega^2) = 1+\omega^{2n}$ and

$$|Y_{12}|^2_{s=j\omega} = \frac{1}{|q(s)|^2}\bigg|_{s=j\omega} = \frac{1}{1+\omega^{2n}}.$$

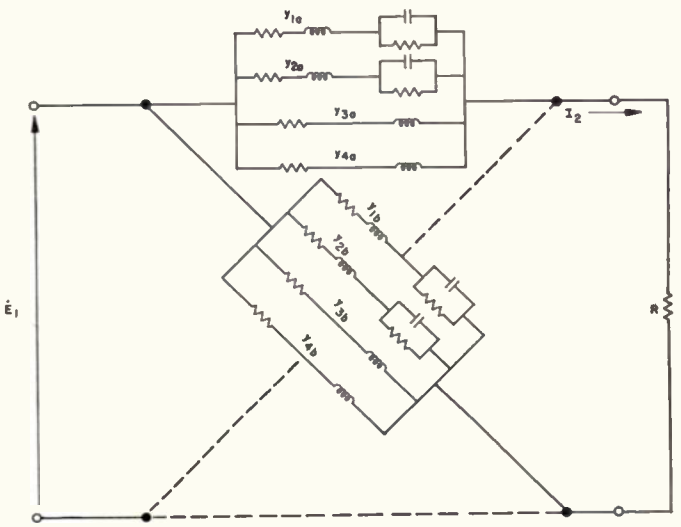

Fig. 5—Lattice for which $Y_{12}=1/2(p/q)$.

We first realize

$$\frac{I_2}{I_1} = \frac{p}{Hq} = \frac{Y_b - Y_a}{Y_b + Y_a}$$

as a short-circuited lattice and then by transformations obtain the desired Y_{12}.

The breakdown $q=q_1+Aq_1'$ for $A=0.1$ yields

$$q_1=(s^2+1.61976s+0.813459)(s^2+0.59336s+0.975703)$$

and therefore

$$\frac{1}{q_1} = \frac{0.44509 - j1.0854}{s + 0.80988 - j0.39693} + \frac{-0.44509 + j0.21483}{s + 0.29668 - j0.94217} + \text{conjugates.}$$

By inspection we see

$\alpha_1^{(n)} =$	0.44509	$\beta_1^{(n)} =$	-1.0854
$\alpha_2^{(n)} =$	-0.44509	$\beta_2^{(n)} =$	0.21483
$\sigma_1 =$	0.80988	$\omega_1 =$	0.39693
$\sigma_2 =$	0.29668	$\omega_2 =$	0.94217.

We find from

$$c_\mu = \frac{\frac{1}{2}\,|\beta_\mu^{(n)}|\,\omega_\mu}{\sigma_\mu}$$

that

$$c_1 = 0.2660$$
$$c_2 = 0.3411.$$

By use of (21) and (22), we obtain

$$H_1 \geqq 9.8$$
$$H_2 \geqq 10.3.$$

We choose $H=25$. Therefore

$\alpha_0^{(b)} =$	12.5	$\alpha_0^{(a)} =$	12.5
$\alpha_1^{(b)} =$	1.47255	$\alpha_1^{(a)} =$	1.02746
$\alpha_2^{(b)} =$	1.02746	$\alpha_2^{(a)} =$	1.47255
$\beta_1^{(b)} =$	-0.5427	$\beta_1^{(a)} =$	0.5427
$\beta_2^{(b)} =$	0.10742	$\beta_2^{(a)} =$	-0.10742.

The lattice arms are then given by

$$Y_a = 12.5 + \frac{2.05492(s + 0.60022)}{s^2 + 1.61976s + 0.813459} + \frac{2.94510(s + 0.36541)}{s^2 + 0.59336s + 0.975703}$$

$$Y_b = 12.5 + \frac{2.94510(s + 0.95617)}{s^2 + 1.61976s + 0.813459} + \frac{2.05492(s + 0.19818)}{s^2 + 0.59336s + 0.975703}.$$

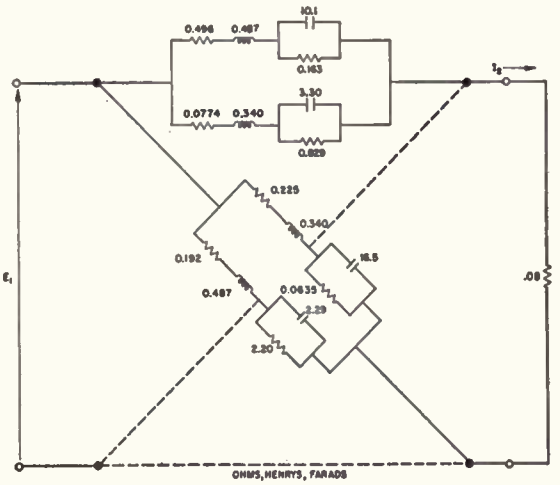

Fig. 6—Network realized in the illustrative example where $Y_{12}=1/2(p/q)$.

The application of the necessary transformations gives the lattice shown in Fig. 6 for which

$$Y_{12} = \frac{1}{2}\,\frac{p}{q} = \frac{1}{2}\,\frac{1}{(s^2+0.76536s+1)(s^2+1.84776s+1)}.$$

VIII. Conclusion

A method has been demonstrated for realizing any transfer function by a lattice network. Each inductance used has an associated series resistance so that lossy coils may be used in building the network. General methods for reducing a lattice to an unbalanced form have been considered, and it was shown that if real transformers are allowed, i.e., transformers with winding resistance, finite magnetizing inductance, and a coupling coefficient smaller than one, then the lattice realizing Y_{12} is always reducible to an unbalanced network.

Appendix I

Partial Fraction Expansion of a Positive Real RLC Function

For the driving-point functions of two-element kind networks the Foster method of synthesis can be successfully applied, that is, a partial-fraction expansion may be made to yield terms each of which is positive real and therefore realizable by inspection. Though the method breaks down for general RLC functions, it may work for a particular type of RLC function. In this appendix we determine the necessary conditions on the residues for the success of the method;[6] we determine, in addition, the conditions necessary so that a series resistance can be associated with every inductance; and finally we indicate the application of the above to the the synthesis treated in this paper.

The partial-fraction expansion of a positive real RLC impedance

$$Z = \frac{k_1}{s - s_1} + \frac{k_2}{s - s_2} + \frac{k_3}{s - s_3} + \frac{k_4}{s - s_4} + \qquad (41)$$

can be written more explicitly as

$$\begin{aligned} Z &= \frac{\alpha_1 + j\beta_1}{s + \sigma_1 - j\omega_1} + \frac{\alpha_1 - j\beta_1}{s + \sigma_1 + j\omega_1} + \frac{R}{s + a} + \\ &= \frac{2\alpha_1\left(s + \sigma_1 - \dfrac{\beta_1\omega_1}{\alpha_1}\right)}{s^2 + 2\sigma_1 s + \omega_0^2} + \frac{R}{s + a} + \qquad (42) \\ &= z_1 + z_2 + \end{aligned}$$

so that the typical terms z_1 and z_2 become apparent. To simplify exposition we have assumed only simple poles; we also have considered an impedance, the dual of the following remarks of course applying to an admittance. In (42), σ_1 and ω_1 are real and positive constants, the constants R, α_1, and β_1 are real, and $\omega_0^2 \equiv \sigma_1^2 + \omega_1^2$. In order for terms like z_2 (with negative real poles) to be positive real and hence separately realizable as a simple driving-point impedance, it is clear that R must be positive. However, the condition on the residue of the partial-fraction component z_1 containing complex poles is not so obvious, except that it is necessary for α_1 to be positive. By application to z_1 of the well known test for positive real character,[4] the additional condition is found to be

$$\frac{|\beta_1|}{\alpha_1} \leqq \frac{\sigma_1}{\omega_1}. \qquad (43)$$

In words, then, the condition for the existence of a positive real partial-fraction component z_1 is that α_1 be positive and the angle formed by the imaginary axis and the radius from the origin to the pole be greater than or equal to the angle of the residue of the pole. This is illustrated in Fig. 7 where angle $\phi = \tan^{-1}(\beta_1/\alpha_1)$ must be less than or equal to the angle $\psi = \tan^{-1}(\sigma_1/\omega_1)$, or the residue may lie anywhere in the crosshatched portion of the plane.

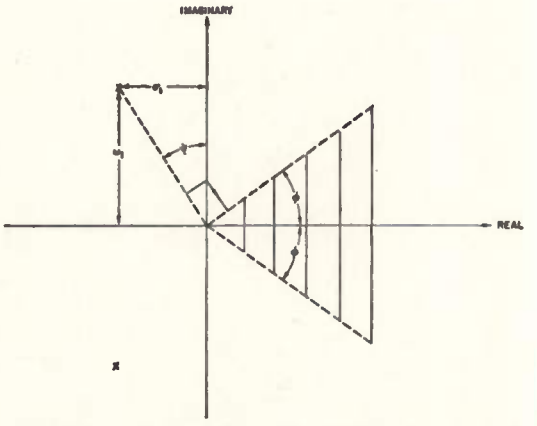

Fig. 7—Graphical illustration of pole and residue relationship.

It is desired that the inductance used in the synthesis of z_1 have a series resistance associated with it. Since β_1 may be positive or negative, two cases arise. For $\beta_1 > 0$, if (43) is satisfied with the equality sign, then $\beta_1\omega_1/\alpha_1 = \sigma_1$ and the quantity $(\sigma_1 - \beta_1\omega_1/\alpha_1)$ appearing in (42) equals zero, so that a perfect coil is required. (For the admittance case all of the dissipation would be associated with the inductance.) On the other hand, if the inequality sign is used, $(\sigma_1 - \beta_1\omega_1/\alpha_1)$ is greater than zero, which guarantees a lossy coil. For the case of a negative β_1, it is possible to associate all of the dissipation with the coil by making $\sigma_1 - \beta_1\omega_1/\alpha_1 = 2\sigma$, with satisfaction of (43) by the equals sign. Again, satisfaction with the inequality sign allows some dissipation to be associated with the coil. Thus, in both cases ($\beta_1 > 0$, $\beta_1 < 0$), use of the inequality sign calls for an impedance containing a lossy coil; and since as is demonstrated in this paper, the lattice synthesis procedure requires that the lattice arms Z_a and Z_b have β's which are equal numerically but of opposite sign, it is mandatory that the inequality sign be used if it is desired to obtain lossy coils in both arms.

[6] These conditions were pointed out by E. A. Guillemin in his Network Synthesis Course at M.I.T.

The synthesis procedure of this paper guarantees that Z_a and Z_b have positive real partial-fraction components with lossy coils, and thus the arms may be realized by inspection in the Foster manner.

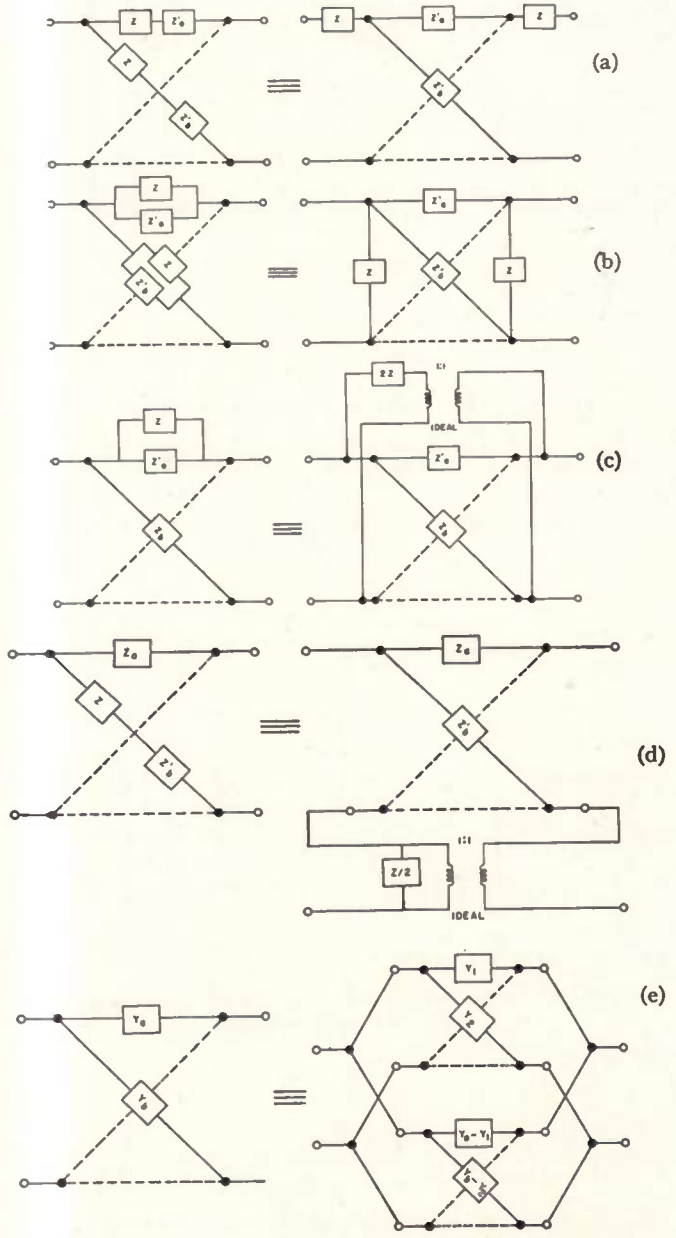

Fig. 8—Methods for the conversion of lattices to unbalanced networks.
(a) Removal of a series impedance from each arm.
(b) Removal of a shunt impedance from each arm.
(c) Removal of a shunt impedance from series arm.
(d) Removal of a series impedance from the cross arm.
(e) A single lattice decomposed into two lattices in parallel.

Appendix II

Reduction of Lattices to Unbalanced Networks

An unbalanced form of network, that is, one with a common ground from the input to the output terminals, is definitely preferred to a lattice. The problem of converting a lattice to an unbalanced network is therefore an important one. This appendix first considers general lattice equivalents, then passes on to reduction of the special types of lattices that may arise from the synthesis procedure of this paper, and finally shows that the reduction of a network realizing a Y_{12} is always possible with at most real transformers.

Lattices may be transformed to unbalanced networks in the following ways:[3,5]

(a) A series impedance may be removed from both Z_a and Z_b and placed in series with both the input and output terminals (see Fig. 8(a)).

(b) A shunt impedance may be removed from both Z_a and Z_b and placed in shunt with both the input and output terminals (see Fig. 8(b)).

(c) A shunt impedance may be removed from the series arm Z_a and considered as a bridge across the remainder of the lattice (see Fig. 8(c)). The ideal transformer that is necessary may be removed when the remainder of the lattice has been transformed to an unbalanced network.

(d) A series impedance may be removed from the cross arm as shown in Fig. 8(d). Again the ideal transformer becomes unnecessary when the remainder of the lattice has been transformed to an unbalanced network.

(e) A lattice may be broken into a group of parallel lattices (see Fig. 8(e)).

In any specific problem, to reduce a lattice to an unbalanced form may require the application of a succession of the above methods, or it may first be necessary to resynthesize a lattice arm before one of the methods may be successfully applied. Thus considerable ingenuity is called for. However, the types of lattices that are realized in this paper, since they contain the same poles in both arms, are reducible in a large number of problems. We mention below a few of the forms which can be recognized as reducible in general.

If all the residues in the real poles of one arm are larger than the corresponding residues in the other arm and, in addition, the coefficients of the numerators of the pairs of complex poles in the first arm are larger than the corresponding coefficients of the poles in the other arm, then the lattice is immediately reducible to an L network. Since this is a restricted form of network, the residues rarely will have this desired distribution.

A completely general form of unbalanced network is given if, at any stage of the lattice reduction process, the lattice arms can be resynthesized into the ladder forms shown in Fig. 9, where G represents a conductance and Z an impedance. A two-element kind network is used for simplicity of illustration. How to bring this about in general is a matter requiring further investigation,[7] but when it is possible the lattice containing these arms can be reduced to the form shown in Fig. 10.

[7] After this had been written (in 1951), it came to the writer's attention that O. Aberth was conducting an investigation into this problem for RC networks as his Master's thesis research at M.I.T.

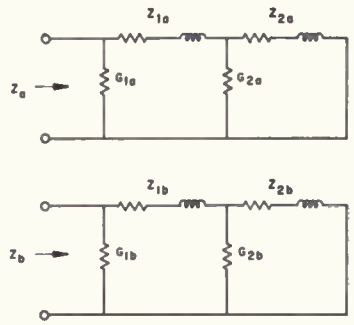

Fig. 9—Desirable form of lattice arm impedances, where $G_{na} \geqq G_{nb}$ and $Z_{nb} \geqq Z_{na}$.

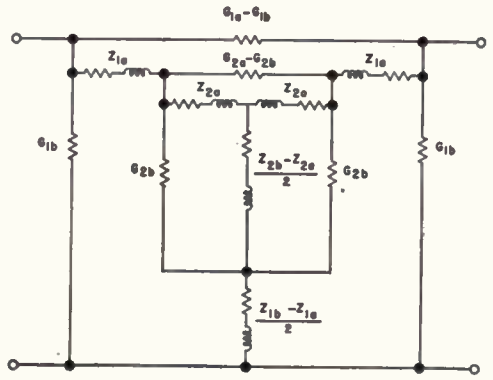

Fig. 10—Unbalanced network corresponding to lattice with arms given in Fig. 9.

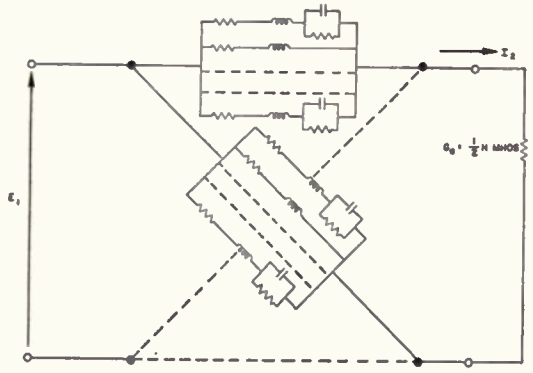

Fig. 11—Six-pole lattice to be reduced to unbalanced network.

For the lattice obtained in the synthesis of Y_{12}, it is always possible to effect a reduction to an unbalanced network if one of the residues in a real pole is very large, specifically, large enough for the method to be applied. As an example of the reduction procedure when a large residue is present, consider the six-pole lattice shown in Fig. 11, whose arms are given by

$$Y_a = y_{1a} + y_{2a} + y_{3a} + y_{4a}$$

$$= \frac{2s+1}{s^2+2s+5} + \frac{3s+4}{s^2+4s+6} + \frac{8}{s+2} + \frac{2}{s+4} \qquad (44)$$

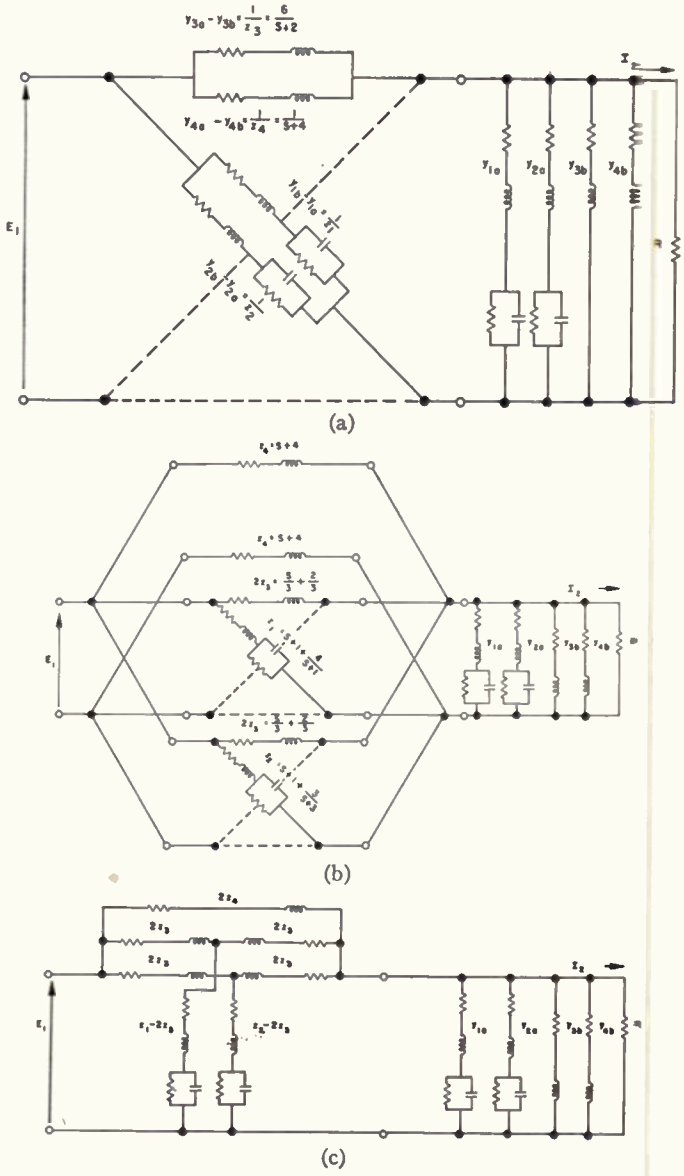

Fig. 12—Steps in reduction of lattice that contains a sufficiently large residue.

and

$$Y_b = y_{1b} + y_{2b} + y_{3b} + y_{4b}$$

$$= \frac{3s+2}{s^2+2s+5} + \frac{4s+7}{s^2+4s+6} + \frac{2}{s+2} + \frac{1}{s+4}. \qquad (45)$$

Because of the large residue in the admittance y_{3a}, the lattice can be reduced. First we remove from each arm the shunt branches y_{1a}, y_{2a}, y_{3b}, y_{4b} (see Fig. 12(a)). Since the drive is a voltage source the shunt branches may be omitted from the input terminals. Then we split the remaining lattice into two parallel lattices with a bridging branch, as shown in Fig. 12(b). Finally these may be transformed, as shown in Fig. 12(c), to obtain a bridged twin-tee network with a complicated load.

As a last resort, if all other methods fail, it is possible by the use of *real* transformers always to reduce any lattice obtained in the synthesis of Y_{12} by the methods and in the way presented here in this paper. The procedure is explained below.

As is well known, any lattice may be reduced to an unbalanced form by the use of an ideal transformer. The process calls for a rotation of the output terminals so that the cross arms become series arms. In order to compensate for this rotation an ideal transformer providing a phase reversal is necessary. The procedure is illustrated in Fig. 13.

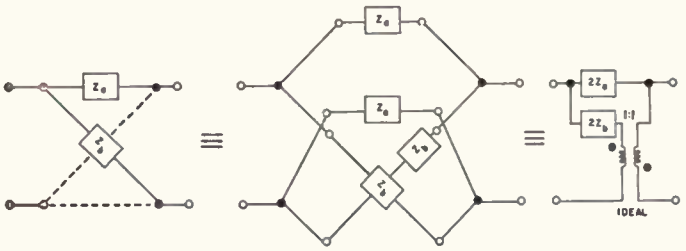

Fig. 13—Reduction of a lattice by use of ideal transformer.

However, we wish to use only real transformers, i.e., transformers possessing leakage inductance, winding resistance, finite-magnetizing inductance, and core loss; the equivalent circuit of such a real transformer is an ideal one with a series resistance and inductance and a shunt resistance and inductance. We shall now show that the form of network realized in the synthesis of a transfer admittance automatically provides the necessary series and shunt branches.

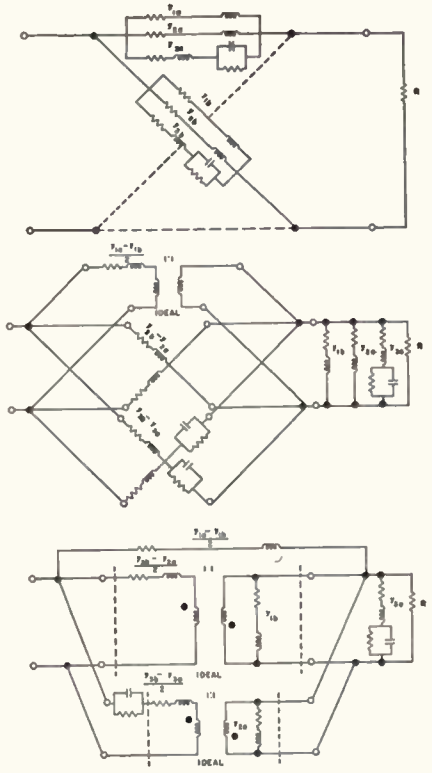

Fig. 14—Reduction of a lattice by use of real transformers.

The general procedure is best explained by a simple example. Without loss of generality suppose we consider the lattice shown in Fig. 14(a) where the numerator of y_{1a} is larger than the numerator of the same pole in the cross arm, but the numerators of y_{2b} and y_{3b} are larger than the corresponding numerators of y_{2a} and y_{3a}. This is the form of lattice obtained in realizing a transfer admittance. We may remove y_{1b}, y_{2a}, and y_{3a} from both arms and divide the lattice into a group of lattices as shown in Fig. 14(b). Then the final rotation step and use of ideal transformers, which now have series and shunt inductances, may be carried out as in Fig. 14(c). Each of the networks within the broken lines represents the equivalent circuit of a real transformer.

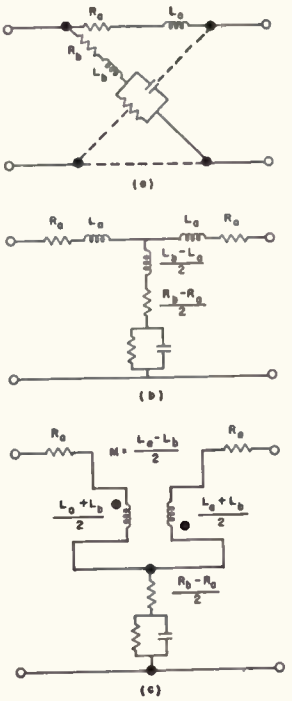

Fig. 15—Reduction of a lattice by use of mutual inductance.

If it is desired to use physically realizable mutual inductance without a core-loss resistance, we may use another method that is often applicable. The lattices that realize a transfer admittance may be divided into components similar to that shown in Fig. 15(a). It is seen that the only requirement for the reduction of this component lattice is that R_b be greater than R_a. For, by removing R_a and L_a from each arm, we then obtain the network shown in Fig. 15(b). If (L_b-L_a) is negative, we can use the mutual inductance form of network shown in 15(c) of the figure for practical realization. All that remains is to show how to realize a sufficiently small R_a. This happy circumstance often comes about naturally in the realization of a transfer admittance. If it does not, it can often be brought about by the multiplication of numerator and denominator of Y_{12} by $(s+a)$ as the first step in the synthesis procedure, where a is a sufficiently small positive constant.

Thus, in summary, we see that we may often arrive at an unbalanced form of network by the methods presented in this appendix. Moreover, if we allow realizable mutual inductance in the synthesis of Y_{12}, we can *always* obtain an unbalanced network.

Normalized Phase and Gain Derivatives as an Aid in Evaluation of FM Distortion*

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Summary—**Second and third order harmonic distortions are evaluated which are generated in a network of known transfer functions when subjected to an undistorted FM input signal. From the general Carson and Fry formula for the response of the network, both quasi-stationary and nonquasi-stationary terms can be evaluated with the aid of normalized derivatives of the gain and phase function of the network. These normalized functions can be calculated by adding contributions of every significant pole and zero of the transfer function. The distortions are evaluated in discrete terms and presented in a form which lends itself well to quantitative discussion. One numerical example is included.**

INTRODUCTION

THE PROBLEM of distortionless transmission of frequency-modulated signals through electric networks is of considerable practical importance. Methods of evaluation of the shape of the signal after passing through amplifier circuits were presented by Frantz[1] and Giacoletto.[2] The method used by both these authors involved the presentation of the input signal in the form of a spectrum composed of carrier and sidebands and then operation on every spectral component separately by altering its magnitude and phase in accordance with the transfer function of the network. The evaluation of the output signal is based on the subsequent synthesis of the individual spectral components. The above-referenced publications do not provide a direct method of evaluation of distortion or intermodulation components generated by the network. Recently a method of evaluation of distortion components from the output spectrum was published by Robin[3] and Poincelot,[4] the network being assumed in the form of a single-tuned circuit.

Alternatively it is possible to evaluate directly distortion and intermodulation terms generated in the network using the concept of "instantaneous" frequency defined (with necessary qualifications) as the time rate of change of phase[5,6] and assuming "quasi-stationary" operation of the circuit.[7,8,9,10,11,12,13]

The assumption of "quasi-stationary" network behavior means that the instantaneous amplitude and frequency at the output can be derived from the instantaneous amplitude and frequency at the input by multiplication by the value of the transfer function applicable to that frequency. A component of the network response is neglected in this case which is due to the circumstance that the frequency at the input to the network does not vary infinitely slowly. There is theoretical and experimental evidence that, at least in a number of cases of practical importance the "quasi-stationary" approach is well justified.[7,9,14] On the other hand the desirability of exact approach is pointed out by some authors.[3,4]

If a frequency-selective network is subjected to a forcing voltage of rapidly varying frequency, the response of the network can be evaluated from its known transfer function. A few publications dealing with this problem have also appeared in the literature.[15,16]

Conditions prevailing in the case of networks intended to handle FM signals with low harmonic distortion are usually such that quasi-stationary analysis is applicable at lower values of FM deviation and at lower modulation frequencies. The author has previously published[12] a method applicable to quasi-stationary conditions and enabling the calculation of distortion and intermodulation terms introduced by a network of known transfer function. This paper will provide a short summary of the mentioned method and

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[1] W. J. Frantz, "The transmission of frequency modulated wave through a network," PROC. I.R.E., vol. 34, pp. 114–125; March, 1946.

[2] L. J. Giacoletto, "Network transmission of a frequency modulated wave," PROC. I.R.E., vol. 35, pp. 1105–1106; October, 1947.

[3] L. Robin, "Calcul de la distorsion harmonique en modulation sinusoidale de frequence et de phase. Etude detaillee d'un quadripole avec circuit antiresonnant," *Annales des Telecommunications*, t. 7, pp. 459–465; Novembre, 1952.

[4] P. Poincelot, "La notion de frequence instantanee," *L'Onde Electrique*, Avril, 1953.

[5] J. Shekel, "Instantaneous Frequency," PROC. I.R.E., vol. 41, page 548; April, 1953.

[6] J. J. Hupert, "Instantaneous Frequency," PROC. I.R.E., vol. 41, page 1188; September, 1953.

[7] S. Fast, "Crosstalk in frequency and phase modulated radio relays used in conjunction with multichannel telephone equipment," Signal Corps Engineering Labs., *Technical Memorandum M 1259*, October 11, 1949.

[8] J. Fagot, "Methode permettant de calculer en modulation de frequence la distortion non-lineaire en fonction des variations du temps de propagation d'un amplificateur," *Annales de Radioelectricite*, t. 5, pp. 1–4; Janvier, 1950.

[9] L. J. Libois, "Distorsions non-lineaires en modulation de frequence," *Cables et Transmissions*, 4ºA, pp. 297–307; Octobre, 1950.

[10] T. Vellat, "Der empfang frequenzmodulierten wellen," *Elektrische Nachrichten Technik*, bd. 18, pp. 61–96; April, 1951.

[11] A. T. Starr and T. H. Walker, "Microwave radio links," *Jour. of the Inst. of Elec. Eng.*, vol. 99, pp. 241–255; September, 1952.

[12] J. J. Hupert, "A method of evaluation of the quasi-stationary distortion of FM signals in tuned interstages," Proc. NEC 1952, vol. 8, pp. 445–461; January, 1953.

[13] E. Hölzler, "Über die Wirkung von Verzerrungen bei der Übertragung frequenzmodulierter, Schwingungen," *Elektrische Nachrichten Technik*, bd. 18, h. 5, pp. 106–117.

[14] J. J. Hupert and A. M. Reslock, "A method of alignment of band-pass amplifiers," PROC. I.R.E., vol. 41, pp. 1668–1671; November, 1953.

[15] G. Hok, "Response of linear resonant systems to excitation of a frequency varying linearly with time," *Jour. Appl. Phys.*, vol. 19, pp. 242–250; March, 1948.

[16] J. Marique, "The response of RLC resonant circuits to EMF of sawtooth varying frequency," PROC. I.R.E., vol. 40, pp. 945–950; August, 1952.

also will deal with its generalization beyond the applicability of the quasi-stationary analysis.

In contrast to several above-cited references, the answer is obtained in the form of discrete distortion terms and it can be obtained for any network, once its transfer function is known.

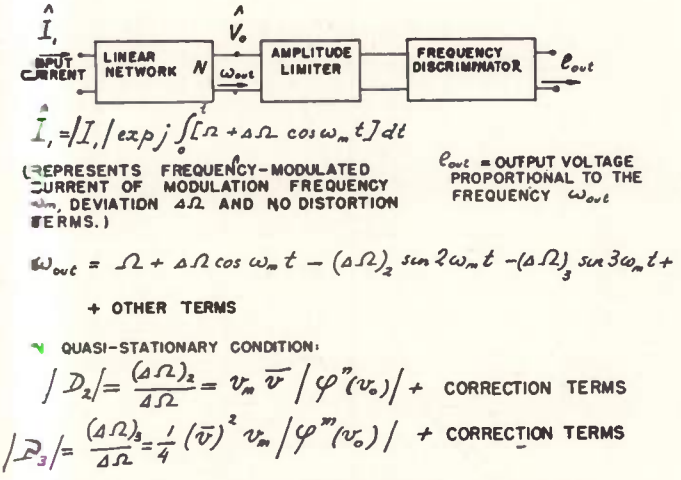

Fig. 1—Summary of results for the quasi-stationary case.

An ideal amplitude limiter and frequency discriminator are assumed to follow the studied network which is considered to be perfectly linear (see Fig. 1). These assumptions are highly significant and are frequently (although not always) reasonably well approximated in practice in the vicinity of the pass-band center. Quantitative ideas as to desirable excellency of limiting can be found in a paper by Vellat.[10] Frequency discriminator is understood as a device providing instantaneous voltage proportional to the time rate of change of the argument of the sum phasor of all significant spectral components of the output voltage.[6] In simpler words the discriminator is expected to respond to instantaneous output frequency. This qualification is necessary since the author has found at least one reference[3] where the harmonic distortion was defined in terms of instantaneous phase of the output voltage.

Summary of Results of the Quasi-Stationary Analysis

It has been demonstrated theoretically[12] and experimentally[14] that second and third harmonic distortion can be with approximation represented by (1) and (2) below:

$$|D_2| = \omega_m \left| \tfrac{1}{2}(d^2\phi/d\omega^2)_{\omega=\Omega}\Delta\Omega + \frac{1}{24}(d^4\phi/d\omega^4)_{\omega=\Omega}\Delta^3\Omega \right| \quad (1)$$

$$|D_3| = \tfrac{1}{8}\omega_m \Delta^2\Omega (d^3\phi/d\omega^3)_{\omega=\Omega} \quad (2)$$

where:

ω_m = modulation frequency
$\Delta\Omega$ = frequency deviation
Ω = carrier frequency
$\phi = f(\omega)$: phase-versus-frequency dependence of the network
(frequencies expressed in angular scale).

As it is well known[17] the transfer function of the network can be expressed in terms of its poles and zeros:

$$Z(s) = \frac{\hat{V}_0(s)}{\hat{I}_1(s)} = (\text{constant}) \frac{\prod_{i=1}^{i=s} (s - s_i)}{\prod_{k=1}^{k=m} (s - s_k)} \quad (3)$$

where s_k and s_i are position of poles and zeros in the complex: $s = \alpha + j\omega$ plane. The variable s can be treated as a generalized complex variable for the purpose of determination of poles and zeros but in the case of real frequency ω the transfer impedance $Z(\omega)$ can be obtained from $Z(s)$ by setting $s = j\omega$.

The above transfer function of band-pass networks encountered in FM equipment usually assumes simpler form:

$$Z(s) = (\text{constant}) \frac{s^g}{\prod_{k=1}^{k=m} (s - s_k)} \quad (4)$$

with a multiple zero at the origin. Poles of function (4) the number of which is m will be found to be grouped in conjugate pairs. For "narrow-band" analysis the effect of zeros and of one half of poles may be neglected and substituted by a constant factor,[17] but the generality of the case discussed is not affected whether or not the "narrow-band" procedure is followed.

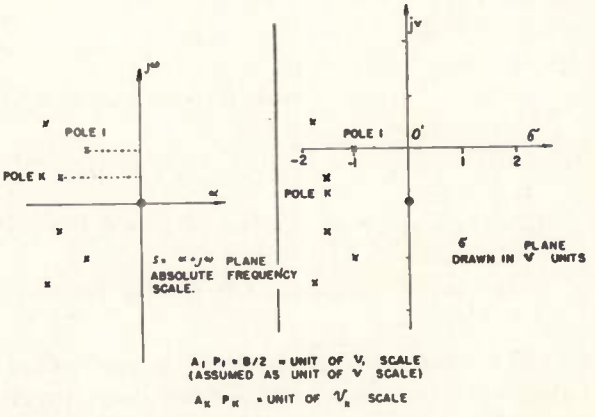

Fig. 2—Transformation of the complex plane from absolute to normalized scale.

Let us change the scale and shift the origin of the $s = \alpha + j\omega$ plane as shown in Fig. 2. We have shifted the origin vertically to a point opposite pole (1) chosen as a reference, introducing the concept of "reference-bandwidth" B such that $B/2$ equals the absolute value of abscissa of pole (1) in the scale in which it appears in the original s plane, i.e., in the scale of frequency. The

[17] D. L. Trautman, "The application of conformal mapping to the synthesis of band-pass networks," *Proc. of the Symposium on Modern Network Synthesis*, April, 1952.

unit of frequency of the new plane equals $B/2$. The location of the new origin O' is quite arbitrary and may be dictated by computational convenience (for example O' may be located in the pass-band center, in a position opposite reference pole as in Fig. 2, or in any other convenient fashion).

Let us next normalize modulation frequency and frequency deviation to B by the introduction of symbols:

$$\bar{v} = 2\Delta\Omega/B \tag{5}$$

which we shall call "*normalized FM swing,*"[18]

$$v_m = \omega_m/B \tag{6}$$

named "*normalized modulation frequency*" and

$$v_0 = 2(\Omega - \omega_0)/B \tag{7}$$

called "*normalized carrier-frequency departure*"[18] (from the chosen origin of the normalized plane). Symbol ω_0 denotes the ordinate of new origin O' in the original scale. The new normalized plane $\sigma + jv$ is drawn in "units v_1" as shown in Fig. 2. It was previously shown[12] that:

$$d^n\phi/d\omega^n = (2/B)^n d^n\phi/dv^n \tag{8}$$

and:

$$d^n\phi/dv^n = \sum_{i=1}^{i=z} \frac{1}{|\sigma_i|^n} \frac{d^n}{dv_i} [\arctan (v_i - v_{ii})] - \sum_{k=1}^{k=m} \frac{1}{|\sigma_k|^n} \frac{d^n}{dv_k} [\arctan (v_k - v_{kk})] \tag{9}$$

where:

z = total applicable number of zeros
m = total applicable number of poles
i, k = "running" subscripts
v_{kk} = ordinate of pole k with respect to origin O' in units of v_k scale (see Fig. 2)
v_k = variable ordinate from O' expressed in units of "v_k" scale.
$|\sigma_k|$ = absolute value of abscissa of pole k in units of "v" scale.
The use of subscripts i is similar for the case of zeros.

The form in which (9) is presented is best suited for graphical construction of $d^n\phi/dv^n$ for cases involving several poles and/or zeros. We shall call variable v *normalized frequency variable.* In terms of this variable we can express *normalized instantaneous frequency* at the input to the network as:

$$v = v_0 + \bar{v} \cos \omega_m t. \tag{10}$$

The curves $\phi_1(x) =$ arc tan x and d^n arc tan x/dx^n = function (x) shown in Figs. 3, 4, 5, 6, and 7 represent, after proper scale change (and with sign reversal in the case of a pole), contributions of a simple pole, or a

[18] "Standards on antennas, modulation systems and transmitters," Proc. I.R.E., pp. 7–15; 1948.

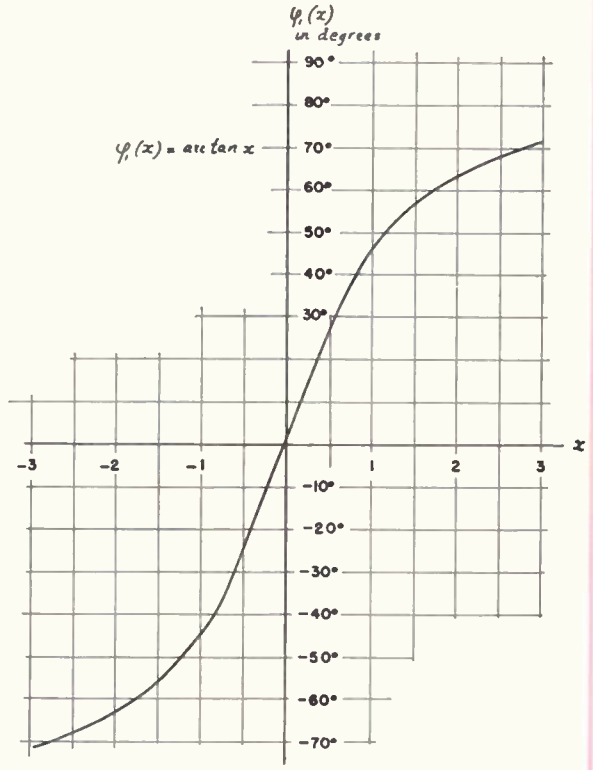

Fig. 3—Normalized contribution of a simple pole to the phase-frequency function.

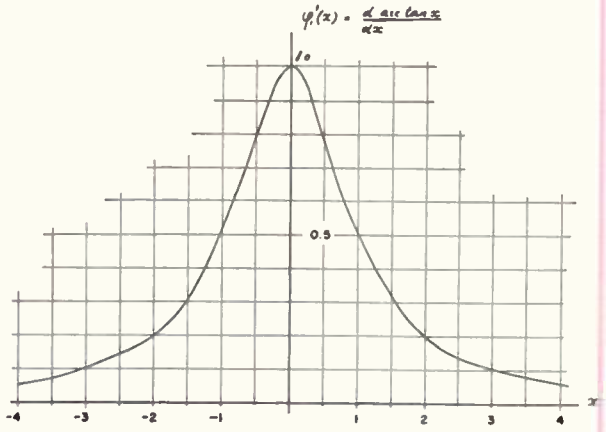

Fig. 4—Normalized contribution of a simple pole to the delay-frequency function.

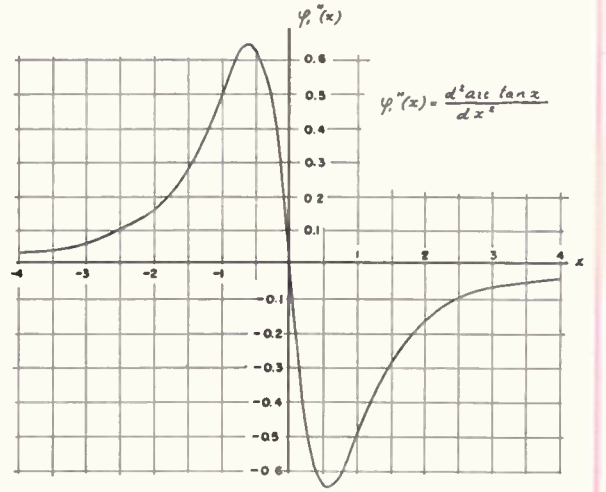

Fig. 5—Normalized contribution of a simple pole to the second derivative of the phase-frequency function.

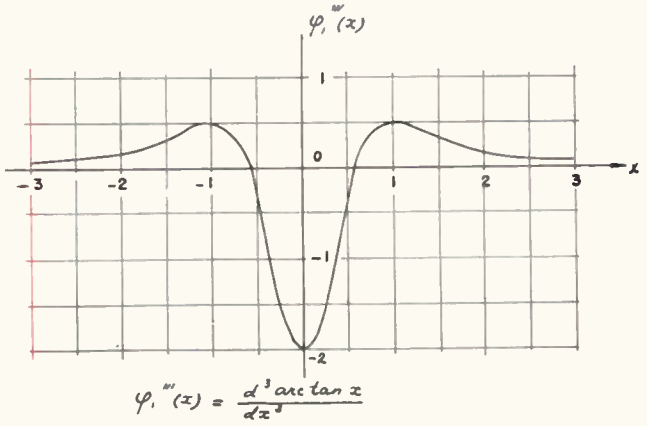

Fig. 6—Normalized contribution of a simple pole to the third derivative of the phase-frequency function.

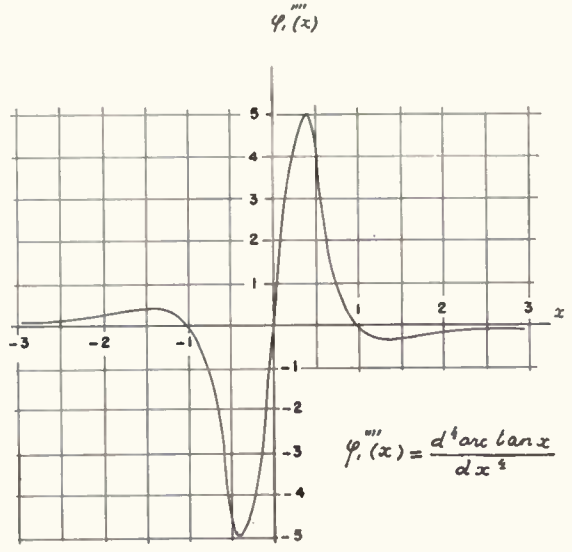

Fig. 7—Normalized contribution of a simple pole to the fourth derivative of the phase-frequency function.

single zero, to the normalized phase-frequency derivative of the order n for the examined network.

The respective equations of these curves are:

$$\phi_1(x) = \text{arc tan } x \tag{11}$$

$$\phi_1'(x) = d(\text{arc tan } x)/dx = (1 + x^2)^{-1} \tag{12}$$

$$\phi_1''(x) = d^2(\text{arc tan } x)/dx^2 = -2x(1 + x^2)^{-2} \tag{13}$$

$$\phi_1'''(x) = d^3(\text{arc tan } x)/dx^3 = 8x^2(1 + x^2)^{-3} - 2(1 + x^2)^{-2} \tag{14}$$

$$\phi_1''''(x) = d^4(\text{arc tan } x)/dx^4 = -48x^3(1 + x^2)^{-4} + 24x(1 + x^2)^{-3}. \tag{15}$$

In normalized phase-derivative computations the independent variable x should of course be substituted by the independent variable applicable to the pole (or zero) considered as indicated in (9).

In terms of the normalized data defined above we obtain for the second and third order distortions:

$$|D_2| \doteq |v_m\bar{v}\phi_0'' + \tfrac{1}{12}v_m(\bar{v})^3\phi_0''''| \tag{16}$$

and:

$$|D_3| \doteq \tfrac{1}{4}v_m(\bar{v})^2|\phi_0'''| \tag{17}$$

where

$$\phi_0'' = (d^2\phi/dv^2)_{v=v}$$

and

$$\phi_0''' = (d^3\phi/dv^3)_{v=v_0}$$

can be evaluated as indicated in (9).

The outlined process of normalization of circuit and modulation data proves helpful in both analysis and synthesis problems.[12,19] In certain conditions it may also provide good basis for determination of experimental method of circuit alignment.[14] Bearing in mind that curve (Fig. 4) represents normalized contribution of a pole to the delay of a network[8,9,19] numerical examples of pole location can easily be worked out for a given number of poles to realize "maximal delay flatness" condition in the vicinity of the pass-band center. Owing to space limitation presentation of these examples here is not feasible.

In designing networks on the basis of the quasi-stationary approach it is immaterial whether a network containing several identical stages in cascade connection is considered as a single network or its performance is evaluated on the "per stage" basis where phase shifts and phase shift derivative values are multiplied by the number of stages. If there are several identical interstages in the network and the network is considered as a single entity the poles will obviously appear as multiple poles. Sometimes it is convenient to subdivide the network into identical parts each part containing several interstages which are not identical. It should be remembered that the narrow-band approximation, if used, is based on the contributions of remote poles or zeros to the shape of the phase curve and its derivatives being insignificant, but that the method presented is not restricted to narrow-band approximation. If the narrow-band approximation is used we obtain one pole per single-tuned stage, if not, two conjugate poles per single-tuned stage. If the narrow-band approximation is not used, contribution of each pole in a conjugate pair should be added on an individual basis.

Distortion Terms Contributed by Non-Quasi-stationary Effects

We shall now try to evaluate the distortion contributed by the network as a result of the nonquasi-stationary condition neglected in previous considerations.

So far we have been treating the transfer function of the network as function of the frequency alone identifying the concept of *frequency* with that of *instantaneous frequency* and neglecting the resultant error. As already mentioned, there are two possible approaches to the problem of network response to frequency modulated signals. The first method [1,2,3] consists of computation of

[19] W. A. Lynch, "The role played by derivative adjustment in broadband amplifier design," *Proc. of the Symposium on Modern Network Synthesis*, April, 1952.

the signal in frequency domain, subsequent application of the network-transfer function to every spectral component and re-composition of the resulting terms. The disadvantage of this method of approach is the fact that it becomes very laborious when presentation of the outgoing modulation shape in the time domain is intended and even more so when numerical data concerning signal frequency distortion are required. The previously outlined method, applicable to quasi-stationary cases allows us to determine conveniently distortion data from the known positions of poles and zeros of the transfer function. Since the positions of these poles and zeros in the complex plane govern the character of the network response in the time domain it may be reasonably expected that the outlined method may be generalized to cover nonquasi-stationary applications. With this object in view we shall follow the second available method of evaluation of the response of networks to variable frequency signals—that outlined by Carson and Fry[9,20]

When the quasi-stationary case no longer applies the transfer function ceases to be considered as function of instantaneous frequency alone and a time-dependent correction term appears in the form:

$$Z(s, t) = Z(s) + \tfrac{1}{2}(d^2Z/ds^2)ds/dt. \tag{18}$$

In (18) above, $Z(s)$ represents transfer impedance expressed as function of $s=j\omega$ the concept of frequency being identified with that of the instantaneous frequency and considered as time-dependant variable.

Formula (18) is an approximation of an infinite series the leading term of which should be small compared to $|Z(s)|$:

$$|\tfrac{1}{2}(d^2Z/ds^2)ds/dt| \ll |Z(s)| \tag{19}$$

for the quasi-stationary condition to apply.

Function $Z(s)$ being complex has a modulus M and argument ϕ both being functions of s. Writing:

$$Z = M \exp j\phi \tag{20}$$

we can express the second derivative of Z with respect to s in terms of derivatives of functions M and ϕ with respect to s.

Performing necessary differentiations we arrive at the expression:

$$d^2Z/ds^2 = \{[(d^2M/ds^2) - M(d\phi/ds)^2] + j[2(dM/ds)(d\phi/ds) + M(d^2\phi/ds^2)]\} \exp j\phi. \tag{21}$$

As mentioned above we identify $s=j\omega$ with the concept of instantaneous frequency which in turn can be expressed as function of time

$$s = j\omega = j(\Omega + \Delta\Omega \cos \omega_m t). \tag{22}$$

Differentiating (22) with respect to time we have:

$$ds/dt = -j\omega_m \Delta\Omega \sin \omega_m t \tag{23}$$

and substituting $s=j\omega$ into (21) and then (21) and (23) into (18) we obtain:

$$Z(\omega t) = M \exp j\phi - (\omega_m\Delta\Omega/2)e^{j\phi} \sin \omega_m t\{[2(dM/d\omega)(d\phi/d\omega) + M(d^2\phi/d\omega^2)] + j[M(d\phi/d\omega)^2 - (d^2M/d\omega^2)]\}. \tag{24}$$

We notice that in (24) the term M ex $pj\phi$ represents $Z(\omega)$ while the remainder of the right hand side is the complex time dependant correction, while ω still remains identified with the instantaneous input frequency and, as such, is itself a function of time. In order to evaluate distortion terms in frequency modulation with the assumption of perfect amplitude limiting device following the examined network we are interested primarily in the argument of $Z(\omega, t)$ which after transformations involving conventional complex algebra can be expressed as:

$$\text{Arg } Z(\omega, t) = \phi + \arctan \frac{\omega_m\Delta\Omega[(1/M)(d^2M/d\omega^2) - (d\phi/d\omega)^2] \sin \omega_m t}{2 - 2\omega_m\Delta\Omega[(1/M)(dM/d\omega)(d\phi/d\omega) + \tfrac{1}{2}(d^2\phi/d\omega^2)] \sin \omega_m t}. \tag{25}$$

both M and ϕ being functions of ω. Similarly to the case of quasi-stationary condition we can now change the scale of the complex plane to the $\sigma+jv$ scale where v is defined as fractional frequency normalized to a selected pole or zero in accordance with the relation

$$v = \frac{\omega - \omega_0}{B/2} \tag{26}$$

B pertains to the singularity chosen as reference.

The process of normalization will enable us to change the variable ω to another one v with the object of evaluating in succession contributions of all poles and zeros to the distortion, similarly to the process used previously in connection with the quasi-stationary case. By the process of change of variable we obtain the formula:

$$\text{Arg } Z = \phi + \arctan \frac{v_m\bar{v}[(1/M)(d^2M/dv^2) - (d\phi/dv)^2] \sin \omega_m t}{1 - v_m\bar{v}[(2/M)(dM/dv)(d\phi/dv) + d^2\phi/dv^2] \sin \omega_m t}. \tag{27}$$

Equation (27) can also be presented in the form:

[20] J. R. Carson and T. C. Fry, "Variable frequency electric circuit theory with application to frequency modulation," *Bell Sys. Tech. Jour.*, vol. 16, pp. 513–540; October, 1937.

$$\operatorname{Arg} Z = \phi + \arctan \frac{v_m\bar{v}(\sin \omega_m t)\left[\dfrac{d^2 \ln M}{dv^2} + \left(\dfrac{d \ln M}{dv}\right)^2 - \left(\dfrac{d\phi}{dv}\right)^2\right]}{1 - v_m\bar{v}\left[2\,\dfrac{d \ln M}{dv}\cdot\dfrac{d\phi}{dv} + \dfrac{d^2\phi}{dv^2}\right]\sin \omega_m t}. \tag{28}$$

If the second term of (28) is called $\Phi(v, t)$ then nonlinear terms will arise from:

$$d(\operatorname{Arg} Z)/dt = (d\phi/dt) + (d\Phi/dt)$$

which for a single modulating signal becomes:

$$d(\operatorname{Arg} z)/dt = \text{constant term} + \text{fundamental and distortion terms arising from the quasi-stationary analysis} + d\Phi/dt \tag{29}$$

from which it is seen that contribution due to the fact of the process being nonquasi-stationary arises from $d\Phi/dt$. For small $v_m\bar{v}$ and the expression in the square brackets in the denominator of the second term of (28) small compared with $1/v_m\bar{v}$:

$$\operatorname{Arg} Z(v, t) = \phi + \arctan \{v_m\bar{v}[A'' + (A')^2 - (\phi')^2] \sin \omega_m t\}. \tag{30}$$

Symbols ϕ and $A = Ln M$ denote functions of v while superscripts "prime" and "double prime" symbolize differentiations with respect to that variable.

Approximating two terms of Taylor's expansion for every function:

$$[\phi'(v)]^2 = [\phi'(v_0)]^2 + \left\{\frac{d[\phi'(v)]^2}{dv}\right\}_{v=v_0} \bar{v} \cos \omega_m t + \left\{\frac{d^2[\phi'(v)]^2}{dv^2}\right\}_{v=v_0} (\bar{v})^2 \cos^2 \omega_m t + \cdots$$
$$\doteq (\phi_0')^2 + 2\phi_0'\phi_0''\bar{v} \cos \omega_m t + [(\phi_0'')^2 + \phi_0'\phi_0'''](\bar{v} \cos \omega_m t)^2. \tag{31}$$

Similarly:

$$(A')^2 = (A_0')^2 + 2A_0'A_0''\bar{v} \cos \omega_m t + [(A_0'')^2 + A_0'A_0'''](\bar{v} \cos \omega_m t)^2 \tag{32}$$

$$A'' = A_0'' + A_0'''\bar{v} \cos \omega_m t + \frac{A_0''''}{2}(\bar{v} \cos \omega t)^2. \tag{33}$$

Subscript "0" denotes values of derivatives for $v = v_0$. Utilizing Taylor expansions (31), (32) and (33) in (30):

$$\operatorname{Arg} Z(t) = \phi + \Phi(t)$$
$$\Phi(t) = v_m\bar{v}F_I \sin \omega_m t + v_m(\bar{v})^2F_{II} \sin \omega_m t \cos \omega_m t + v_m(\bar{v})^3F_{III} \cos^2 \omega_m t \sin \omega_m t \tag{34}$$

where F_I, F_{II} and F_{III} are functions of v_0 only:

$$F_I = A_0'' + (A_0')^2 - (\phi_0')^2 \tag{35}$$

$$F_{II} = A_0''' + 2A_0'A_0'' - 2\phi_0'\phi_0'' \tag{36}$$

$$F_{III} = \frac{A_0''''}{2} + (A_0'')^2 + A_0'A_0''' - (\phi_0'')^2 - \phi_0'\phi_0'''. \tag{37}$$

As explained before the contribution to harmonic output after demodulation will result from expression:

$$d\Phi/dt = [F_I - \tfrac{3}{4}F_{III}(\bar{v})^2]\omega_m v_m\bar{v} \cos \omega_m t + \omega_m v_m(\bar{v})^2F_{II} \cos 2\omega_m t - \tfrac{3}{4}\omega_m v_m(\bar{v})^3F_{III} \cos 3\omega_m t + \text{other terms.} \tag{38}$$

First term of the last equation represents an insignificant contribution to the recovered amplitude of the fundamental. To evaluate the second and third harmonic distortion terms due to nonquasi-stationary behavior the corresponding amplitudes in (38) should be divided by $\Delta\Omega$. Denoting these distortions by D_{2T} and D_{3T} we obtain:

$$|D_{2T}| = 2v_m^2\bar{v}|F_{II}| \tag{39}$$

$$|D_{3T}| = -v_m^2(\bar{v})^2|F_{III}|. \tag{40}$$

It can be demonstrated[12] that if the frequency modulation at the input to the network follows the law: $\omega = \Omega + \Delta\Omega \cos \omega_m t$, the second harmonic term in the output due to phase-frequency curve nonlinearity (quasi-stationary condition) is: $-D_2 \sin 2\omega_m t$ while by (38) above the additional term is $D_{2T} \cos 2\omega_m t$. The two contributions should therefore be added in quadrature forming the resultant D_{2R}:

$$|D_{2R}| = \sqrt{D_{2T}^2 + D_2^2}. \tag{41}$$

D_2 is composed predominantly of the contribution of second derivative $\phi''(v)$ for $v = v_0$:

$$D_{22} = v_m\bar{v}\phi_0'' \tag{42}$$

and that of the fourth derivative:

$$D_{24} = \tfrac{1}{12}v_m(\bar{v})^3\phi_0''''. \tag{43}$$

Thus:

$$|D_2| = |D_{22} + D_{24}|.$$

Similarly

$$|D_{3R}| = \sqrt{D_{3T}^2 + D_3^2} \tag{44}$$

$$D_3 = \tfrac{1}{4}v_m(\bar{v})^2\phi_0'''. \tag{45}$$

Taking the phasor $\exp j2\omega_m t$ as reference we can write in complex form:

$$D_{2R} = D_{2T} - jD_2. \tag{46}$$

After substitutions:

$$D_{2R} = 2v_m^2\bar{v}(A_0''' + 2A_0'A_0'' - 2\phi_0'\phi_0'') + j[v_m\bar{v}\phi_0'' + \tfrac{1}{12}v_m(\bar{v})^3\phi_0'''']. \tag{47}$$

Similarly taking $\exp j3\omega_m t$ as reference:

$$D_{3R} = -(D_{3T} + jD_3) \tag{48}$$

or

$$D_{3R} = \tfrac{3}{2}v_m^2(\bar{v})^2\left[(\phi_0'')^2 + \phi_0'\phi_0''' - A_0'A_0''' - (A_0'')^2 - \frac{A_0''''}{2}\right] - j\tfrac{1}{4}v_m(\bar{v})^2\phi_0'''. \tag{49}$$

Expressions (47) and (49) prove that second and third harmonic distortion can be evaluated entirely from the normalized data of modulation and values of normalized functions of gain and phase derivatives at $v = v_0$.

Closer inspection of the manner in which functions F_{II} and F_{III} are formed from normalized derivatives of gain and phase functions reveals that if the transfer function is symmetrical and the carrier is located at the center of the pass-band there is no first order contribution to the second harmonic distortion since ϕ_0'', A_0' and A_0''' vanish at that point, being odd functions.

Equation (49) reveals that the nonquasi-stationary term of third harmonic distortion vanishes only if, in addition to the above outlined set of conditions, the network is maximally flat both in amplitude and in delay (requirement for $A_0'' = 0$, $A_0'''' = 0$ and $\phi_0''' = 0$). A network of finite number of natural modes cannot satisfy such a requirement rigorously but the distortion value may be found negligible for all practical purposes in many applications and (49) may serve to evaluate it numerically.

For a given network both sets of curves $d^n\phi/dv^n$ and d^nA/dv^n can be evaluated from individual contributions of poles. Equation (9) and Figs. 3 to 7 describe adequately the process of evaluation of $d^n\phi/dv^n$ from $d^n\phi_1/dx^n$. Similarly each function d^nA/dv^n can be obtained from:

$$d^nA/dv^n = \sum_{i=1}^{i=s} \frac{1}{|\sigma_i|^n}\frac{d^n}{dv_i^n}\ln\left[1 + (v_i - v_{ii})^2\right]^{1/2} - \sum_{k=1}^{k=m} \frac{1}{|\sigma_k|^n}\frac{d^n}{dv_k^n}\ln\left[1 + (v_k - v_{kk})^2\right]^{1/2}. \tag{50}$$

Because of the logarithmic character of function $A_1(x)$ the contributions of all poles and all zeros are additive.

The symbols in (50) follow an analogous system to that outlined above for (9). Figs. 8, 9, 10, 11, 12 provide the graphs for the "individual contribution" function $A_1(x)$ and its derivatives up to the fourth order. The corresponding equations are:

$$A_1(x) = \ln(1 + x^2)^{1/2} \tag{51}$$

$$A_1'(x) = \frac{dA_1(x)}{dx} = x(1 + x^2)^{-1} \tag{52}$$

$$A_1''(x) = d^2A_1(x)/dx^2 = (1 + x^2)^{-1} - 2x^2(1 + x^2)^{-2} \tag{53}$$

$$A_1'''(x) = d^3A_1(x)/dx^3 = -6x(1 + x^2)^{-2} + 8x^3(1 + x^2)^{-3} \tag{54}$$

$$A_1''''(x) = d^4A_1(x)/dx^4 = -6(1 + x^2)^{-2} + 48x^2(1 + x^2)^{-3} - 48x^4(1 + x^2)^{-4}. \tag{55}$$

The reader will notice that $A_1(x)$ and its even order derivatives are even functions while its odd order derivatives are odd functions. The reverse is true in the case of function $\phi_1(x)$ and its derivatives (see Figs. 3 to 7).

Reverting to the discussion of obtained distortion terms it will be noticed that for small values of normalized swing $\bar{v}$ the term D_{24} can usually be neglected;

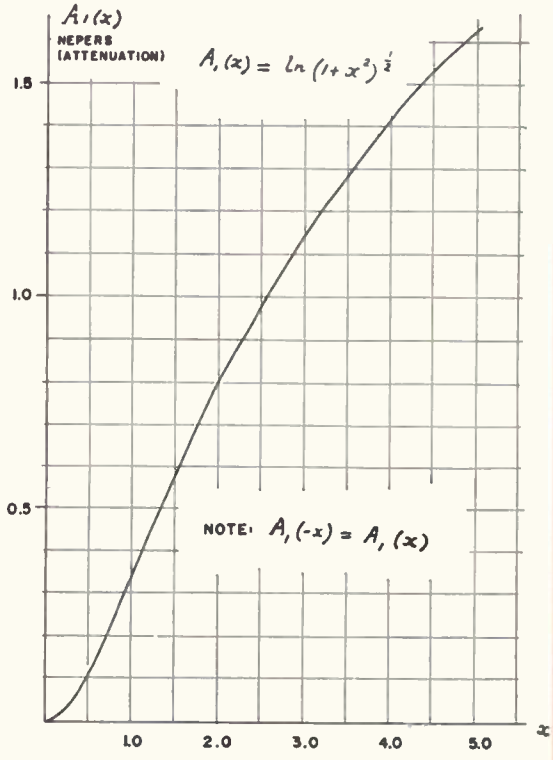

Fig. 8—Normalized contribution of a simple pole to the gain-frequency function.

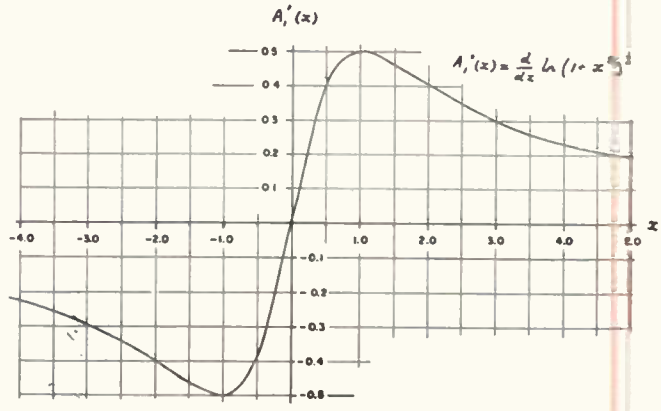

Fig. 9—Normalized contribution of a simple pole to the first derivative of the gain-frequency function.

if therefore the network is so designed that ϕ_0'' and F_{II} are of the same order, D_{2T} is smaller than D_2 by the factor v_m (usually a small fraction). Similar reasoning applies to D_3 and D_{3T} since D_{3T} is proportional to v_m^2 while D_3 is proportional to v_m. It should be pointed out that functions $\phi(v)$ and $A(v)$ of a network and their derivatives can be computed "per stage" (if the cascaded stages are identical, of course), but the applicable values of functions F_{II} and F_{III} must be computed for the network as a whole. Functions F_{II} and F_{III} are important because their magnitude indicates a measure of susceptibility of the network to nonquasi-stationary distortion for a given relative position of the carrier in the pass-band.

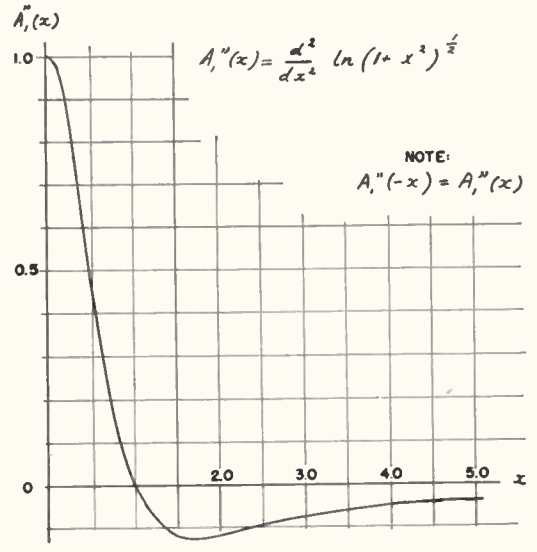

Fig. 10—Normalized contribution of a simple pole to the second derivative of gain-frequency function.

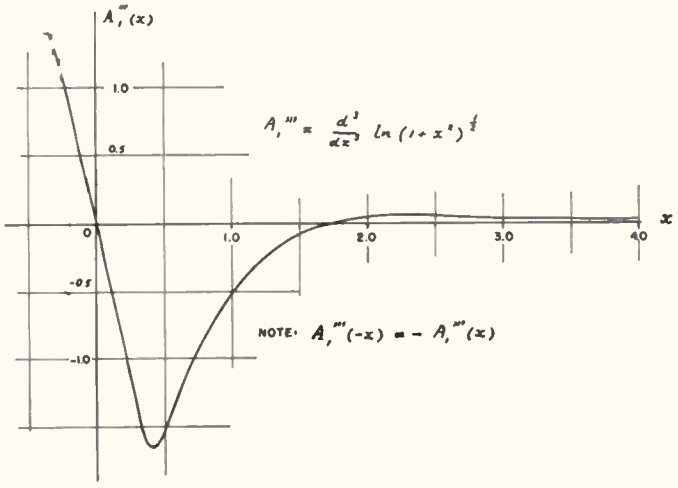

Fig. 11—Normalized contribution of a simple pole to the third derivative of gain-frequency function.

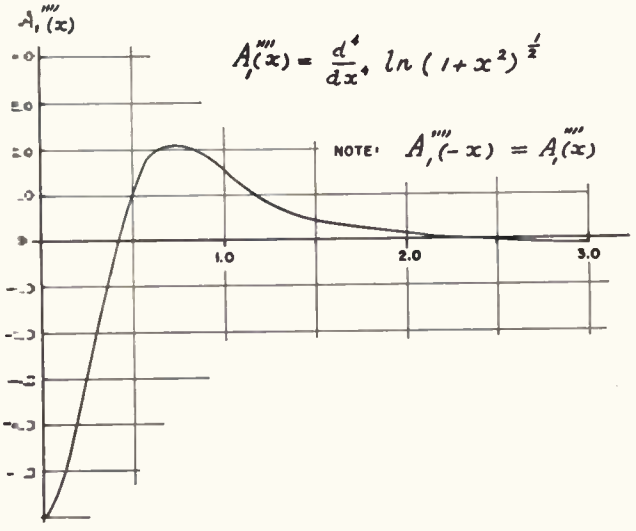

Fig. 12—Normalized contribution of a simple pole to the fourth derivative of gain-frequency function.

Numerical example

A numerical example of practical application of the above-outlined theoretical results follows:

Let us assume a two-pole network with the disposition of poles as shown in Fig. 13. The reader will recognize that the example corresponds to the "narrow-band" case of a flat, staggered double or, with further approximation, to that of a double-tuned, critically coupled stage.[21]

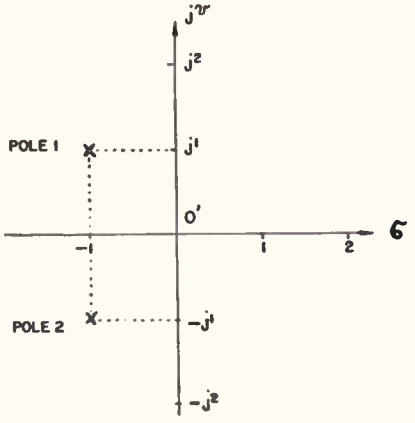

Fig. 13—Disposition of poles of the circuit considered in the numerical example.

Let us assume the following set of conditions: *Frequency of the center of the band:* 4,000 kc. *Magnification factor of each tuned circuit separately:* $Q=40$ which corresponds to the *"reference bandwidth" B as defined in this paper:* $B=100$ kc. Unit of the variable v is then equal to 50 kc. The value of B should not be confused with the total bandwidth of the stage—it is associated only with the reference pole. Since in this case $|\sigma_1| = |\sigma_2| = 1$ (see Fig. 13) it is immaterial which pole is considered as reference. We choose to place the origin of the plane O' in the center of the pass-band. We assume the modulation data as follows:

1. Carrier departure from the pass-band center: 25 kc
2. Modulation frequency: $\omega_m/2\pi = 5$ kc
3. Frequency swing: $2\Delta\Omega/2\pi = 40$ kc.

In normalized terms by previous definitions the above data correspond to: $v_0=0.5$; $\bar{v}=0.4$; $v_m=0.05$. Now we

[21] G. E. Valley and H. W. Wallman, "Vacuum Tube Amplifiers," M.I.T. Radiation Laboratory Series, vol. 18, chap. 4, McGraw-Hill Book Co. Inc., New York, N. Y.; 1948.

can proceed on the basis of (9) and (50) and calculate the derivatives of gain and phase of the network at $v=v_0$ from the contributions of individual poles. Remembering that: $v_{11}=+1$; $v_{22}=-1$; $\sigma_1=\sigma_2=-1$; we obtain from (9):

$$\phi'(v) = d\phi/dv = -\left[\phi_1'(v-1.0) + \phi_1'(v+1.0)\right]$$

and substituting $v=v_0=0.5$ we obtain from the curve Fig. 3: $\phi_0' - [\phi_1'(-0.5)+\phi_1'(1.5)] = -1.26$

By similar process we calculate from (9) and (50) and Figs. 3 to 7 and 9 to 12:

$$\phi_0'' = -0.32 \qquad A_0' = -0.12$$
$$\phi_0'''' = +4.35 \qquad A_0'' = -0.33$$
$$A_0''' = -1.4$$

We shall now calculate contributions to the second harmonic distortion introduced by the considered network:

a. For a single stage:

Quasi-stationary contribution of second order phase-frequency derivative is:

$$D_{22} = v_m\bar{v}\phi_0'' = -0.64(10^{-2}) \text{ per stage.}$$

Quasi-stationary correction for the effect of the fourth order phase derivative:

$$D_{24} = \tfrac{1}{12}v_m(\bar{v})^2\phi_0'''' = +0.106(10^{-2}) \text{ per stage}$$

(small correction at low deviations).
Total quasi-stationary distortion:

$$|D_2| = |D_{22} + D_{24}| = 0.534(10^{-2}) \text{ per stage.}$$

Nonquasi-stationary term:

$$|D_{2T}| = 2v_m^2\bar{v}|F_{II}| = (0.084)(10^{-2}).$$

A single stage of considered circuit contributes negligible nonquasi-stationary distortion at this point of operation.

b. For five identical stages:

Quasi-stationary distortion, five times larger than that of a single stage:

$$|D_2| = 2.66(10^{-2}).$$

Applicable value of $|F_{II}|$ should be re-computed assuming five times larger values of ϕ_0', ϕ_0'', A_0', A_0'' and A_0'''. This calculation gives:

$$|F_{II}| \doteq 27.0$$

and

$$|D_{2T}| = 1.28(10^{-2})$$

resulting in:

$$|D_{2R}| = (10^{-2})\sqrt{(2.66)^2 + (1.28)^2} = (2.94)10^{-2}.$$

Conclusions

In foregoing paragraphs it was demonstrated that both quasi-stationary and first order nonquasi-stationary terms of the second and third harmonic distortion can be evaluated from the network transfer function and modulation data with the use of normalized gain and phase derivatives as a convenient computational tool.

All formulas derived contain only factors depending on the relative positions of poles and/or zeros of the transfer function and on the modulation data normalized to the "reference bandwidth" B. Quasi-stationary and nonquasi-stationary terms add in quadrature.

As an aid in network design much information about the properties of the network can be obtained by plotting its phase and gain derivatives as functions of normalized frequency variable v from individual contributions of transfer function singularities. Similar plots of functions $F_{II}(v)$ and $F_{III}(v)$ will indicate the carrier frequency region where nonquasi-stationary contribution could be neglected with justification.

The nonquasi-stationary components of distortion are proportional to the square of modulation frequency while quasi-stationary contributions are proportional to its first power.

It is thought that this last property may be helpful in recognizing the prevailing contribution in the course of experimental work, provided the examined network does not contain nonlinear elements and assuming adequate limiting of amplitude in the stages by which the network is followed.

Correction

Eduard A. Gerber, author of the paper, "A Review of Methods for Measuring the Constants of Piezoelectric Vibrators," which appeared on pages 1103–1112 of the September, 1953 issue of the PROCEEDINGS OF THE I.R.E., has brought the following correction to the attention of the editors:

The right-hand term of the first line, on page 1109, equation (24) reads:

$$(\omega^2 c_T^2 R_c)^{-1}.$$

It should read:

$$(\omega^2 c_T^2 R_e)^{-1}.$$

A Comparison of Amplitude and Angle Modulation for Narrow-Band Communication of Binary-Coded Messages in Fluctuation Noise*

G. FRANKLIN MONTGOMERY†, MEMBER, I.R.E.

Summary—**Frequency-shift, phase-shift, and on-off amplitude modulation are examined as mechanisms. Curves of fractional error as a function of average carrier-to-noise ratio are derived for fading and non-fading carriers. The results are applicable to telegraph, teletype, and other binary-coded signals.**

Introduction

AN IMPORTANT engineering problem is the selection of a modulation method for communicating over a particular radio channel. Several forms of amplitude and angle modulation are available, and the problem is to choose the method that will transmit messages over the channel with less than a specified error and with the lowest cost of terminal equipment. Performance of each method is affected by the received carrier-to-noise ratio and the properties of the noise and interference. The purpose of this paper is to examine the effects of carrier fading and carrier-to-noise ratio on several types of modulation when the transmitted message is coded in binary form. It is assumed that the interference is band-limited white noise and that carrier amplitude, when fading occurs, is similarly random.

Binary Coding

Comparisons of amplitude and angle modulation are already available,[1-5] but the general case of an arbitrary modulation wave form and small carrier-to-noise ratios requires a difficult mathematical analysis. When the problem is limited to binary coding, the mathematics describing modulation performance is considerably simpler, and most of the argument can be given in terms of probability, phasor diagrams, and words. The limitation to binary-coded messages is not as restrictive as it may appear to be. On the basis of information theory[6,7] binary coding is in fact fundamental, for once the channel has been evaluated for binary coding, its performance with more complex coding or wave forms can be calculated. Telegraph, teletype, several forms of pulse modulation, and infinitely-clipped speech[8,9] are particular types of binary coding. In these practical cases binary-coded channel performance is at once applicable.

Binary coding is the process whereby the carrier amplitude, frequency, or phase is modulated in a time sequence of two mutually exclusive states. Reception of the message depends upon the successful determination of the state of the variable during each time interval in the sequence. In all binary-coded systems, a detecting device of some sort is used to make this determination, or choice, of one of the two possible states of the variable. In some systems the "device" is a human operator; more generally, the device is, for amplitude modulation, an amplitude detector, for frequency modulation, a frequency discriminator, and so on. It is most logical to compare modulation methods by assuming that information concerning the state of the carrier is derived only from the characteristic that is varied intentionally. A device that renders its decision only on the basis of this information will be called ideal. It is certain, of course, that most practical devices will lack the judgment of an ideal device. The system performance of interest, however, is the performance that an ideal device cannot improve, or the performance that can be approached by improvement in practical devices.

Error

Noise, added to the received carrier, makes the operation of the detecting device uncertain. An error occurs whenever the device, in a particular time interval, determines the state of the variable incorrectly. The average error, p, is defined as the fraction of the total number of intervals for which an incorrect decision is made. In terms of system operation, p may also be considered the probability of error. For binary coding, the average error cannot exceed one half, or 50 per cent. This maximum is the same error that would exist if the device were to operate by pure chance, making its decisions regularly but without influence of the received signal. A maximum error of one half applies only to binary coding and only when the two states are equally probable in the transmitted message.

* Decimal classification: R148. Original manuscript received by the Institute, March 12, 1953; revised manuscript received July 31, 1953.

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[1] E. H. Armstrong, "A method of reducing disturbances in radio signaling by a system of frequency modulation," PROC. I.R.E., vol. 24, pp. 689–740; May, 1936.

[2] M. G. Crosby, "Frequency modulation noise characteristics," PROC. I.R.E., vol. 25, pp. 472–514; April, 1937.

[3] J. R. Carson and T. C. Fry, "Variable frequency electric circuit theory with application to the theory of frequency-modulation," *Bell Sys. Tech. Jour.*, vol. 26, pp. 513–540; Oct., 1937.

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[5] D. Middleton, "On theoretical signal-to-noise ratios in f-m receivers: a comparison with amplitude modulation," *Jour. Appl. Phys.*, vol. 20, pp. 334–351; April, 1949.

[6] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. Jour.*, vol. 27, pp. 379–423, 623–656; July and Oct., 1948.

[7] A. E. Laemmel, "General Theory of Communication," Microwave Research Inst., Polytechnic Inst. of Brooklyn, Report R-208-49; July, 1949.

[8] K. D. Kryter, J. C. R. Licklider, and S. S. Stevens, "Premodulation clipping in AM voice communication," *Jour. Acous. Soc. Amer.*, vol. 19, pp. 125–131; Jan., 1947.

[9] J. C. R. Licklider, D. Bindra, and I. Pollack, "The intelligibility of rectangular speech-waves," *Amer. Jour. of Psych.*, vol. 61, pp. 1–20; Jan., 1948.

Modulation

The problem is to calculate the performance of amplitude, frequency, and phase modulation in terms of the received carrier-to-noise ratio. It is assumed that this ratio is measured at the input to the detecting device, whatever the device may be. In a radio receiver, the point of measurement is usually at the output of an intermediate-frequency amplifier.

The modulated carrier and the noise are presumed to be band-limited. This limitation allows either to be described mathematically as a sinusoidal time function for which the time derivatives of percentage amplitude and, generally, reference phase are small compared with the average frequency. The statistical fluctuation of the noise amplitude is given by the Rayleigh formula[10]

$$p_{N,E} = e^{-E^2/N_0^2},$$

where $p_{N,E}$ is the probability, or fraction of the time, that the noise amplitude (envelope) exceeds E, and N_0 is the long-term rms value of the noise envelope. When the carrier is subject to fading, the fading fluctuation is assumed to be specified by

$$p_{S,E} = e^{-E^2/S_0^2},$$

where $p_{S,E}$ is the probability that the carrier amplitude exceeds E, and S_0 is the long-term rms value of the carrier envelope.

In several cases, we shall need the fact that the noise, N, may be resolved into two components, N_p and N_q, the first of which is colinear with an arbitrary reference phase, while the second is in quadrature with the reference phase.[11] The two components are uncorrelated, and each accounts for half of the long-term average power of the noise. The statistical amplitude fluctuation of each component is given by

$$p_{N_i,E} = 1 - I\left(\frac{E}{N_0}\right),$$

where $p_{N_i,E}$ is the probability that the amplitude of the component exceeds E, N_0 is the long-term rms value of the total noise envelope, and

$$I\left(\frac{E}{N_0}\right) = \frac{2}{\sqrt{\pi}}\int_0^{E/N_0} e^{-x^2}dx,$$

which is the probability integral, a function whose values are conveniently tabulated.[12]

For each type of modulation, it is assumed that the ideal detecting device uses a reference threshold upon which to base its decision concerning the state of the variable. For amplitude modulation, it is convenient to designate the amplitude of the threshold, T, as a fraction, aS, of the amplitude of the carrier. Values of the variable that exceed this threshold force a decision for one state, and values below the threshold force a decision for the other. In addition, it is assumed the two states are equally probable in the transmitted message.

We shall make use of the following symbols:

S = carrier amplitude
N = noise amplitude
T = threshold amplitude
a = threshold fraction
S_0 = long-term rms carrier envelope, the square being averaged over a period much greater than the reciprocal of the device-input bandwidth
N_0 = long-term rms noise envelope
$R = S_0^2/N_0^2$, the average power ratio of carrier and noise
p = the probability of error
$p_{X,Y}$ = the probability that X exceeds Y
$\exp(X) = e^X$, where $e = 2.718 \cdots$
$I(X)$ = the probability integral.

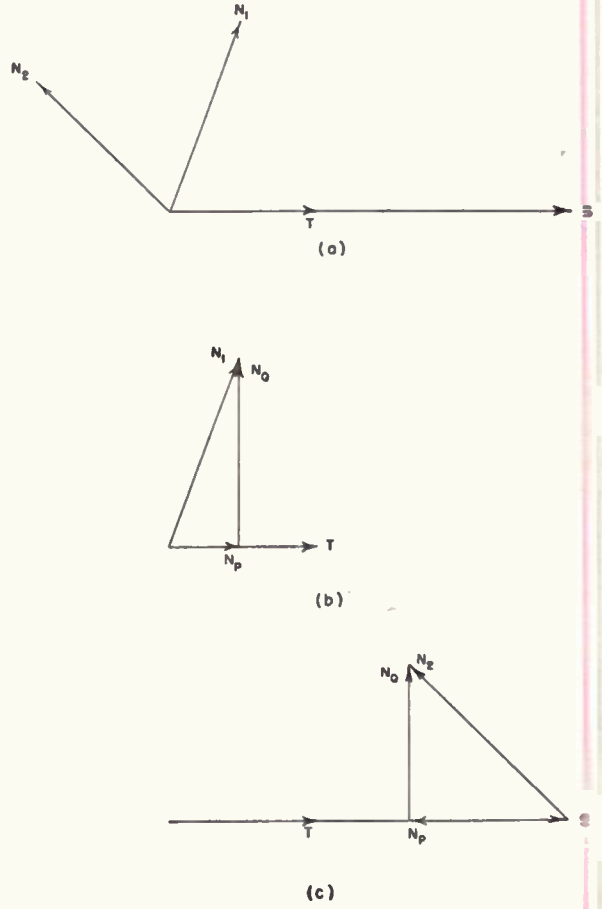

Fig. 1—Carrier and noise phase relations in narrow-band synchronous-detection amplitude modulation.

Amplitude Modulation

In amplitude modulation, transmission of the carrier represents one state and omission of the carrier the other. The ideal device must restrict its judgment, insofar as

[10] V. D. Landon, "The distribution of amplitude with time in fluctuation noise," Proc. I.R.E., vol. 29, pp. 50–55; February, 1941. (See also V. D. Landon and K. A. Norton, "Discussion," Proc. I.R.E., vol. 30, pp. 425–429; September, 1942.)

[11] S. Goldman, "Frequency Analysis, Modulation and Noise," McGraw-Hill Book Co., Inc., New York, N. Y., p. 336; 1948.

[12] B. O. Peirce, "A Short Table of Integrals," Ginn and Co., Boston, Mass., pp. 116–120; 1929.

possible, to amplitude in the mathematical sense, that is, to the values that the sinusoidal carrier function attains at the peak of each cycle. An envelope detector lacks this discrimination, because it is unable to distinguish between amplitudes at the peaks of the carrier function and amplitudes that may be attained by interference in the intervals between peaks. The first system to be examined, then, is one using synchronous detection.[13] For the purpose of this analysis, a synchronous detector is a device that measures amplitudes only during small time intervals that are synchronized with the peaks of the carrier wave.

In Fig. 1, phasor S represents the carrier amplitude and phase. T is a phasor representing the synchronous amplitude threshold; synchronous inputs to the detector that exceed T force a decision for one state, while inputs less than T force a decision for the opposite state. Phasor N_1 represents noise received during a carrier omission, and N_2 represents noise received during a carrier transmission. We resolve each noise into two components, one of which, N_p, is colinear with S while the other, N_q, is in quadrature with S. By virtue of the synchronous action of the detector, only the colinear component of the noise is effective as interference.

Case 1—Synchronous detection, no carrier fading ($S=S_0$), fixed threshold ($T=aS_0$).

The probability that N_p will be in phase with T during a carrier omission (Fig. 1b), as opposed to being out of phase with T, is 1/2. Then, in the synchronous sense,

$$p_{N_p,T} = \frac{1}{2}\left[1 - I\left(\frac{T}{N_0}\right)\right] = \frac{1}{2} - \frac{1}{2} I(a\sqrt{R}).$$

The probability that N_p will be out of phase with S during a carrier transmission (Fig. 1c) is likewise 1/2, and, in the synchronous sense,

$$p_{N_p,S-T} = \frac{1}{2}\left[1 - I\left(\frac{S-T}{N_0}\right)\right] = \frac{1}{2} - \frac{1}{2} I[(1-a)\sqrt{R}\,].$$

Since the two carrier states are equally probable,

$$p = \frac{1}{2}(p_{N_p,T} + p_{N_p,S-T}) = \frac{1}{2} - \frac{1}{4} I(a\sqrt{R}) - \frac{1}{4} I[(1-a)\sqrt{R}\,]$$

which is at minimum for $a=1/2$. For this optimum threshold,

$$p = \frac{1}{2}\left[1 - I\left(\frac{\sqrt{R}}{2}\right)\right]. \tag{1}$$

[13] D. B. Harris, "Selective demodulation," PROC. I.R.E., vol. 35, pp. 565–572; June, 1947.

For large X,

$$I(X) = 1 - \frac{1}{X\sqrt{\pi}} \exp(-X^2),$$

so that for large R,

$$p = \frac{1}{\sqrt{\pi R}} \exp\left(-\frac{R}{4}\right). \tag{2}$$

Case 2—Synchronous detection, carrier fading, *AGC*.

It is assumed in this case that automatic gain control is used to maintain the threshold at its optimum value, $T=S/2$. This regulation will fail if the maximum fading rate is greater than the binary symbol frequency.

For a given carrier level, S, the error is, from (1),

$$p' = \frac{1}{2}\left[1 - I\left(\frac{S}{2N_0}\right)\right] = \frac{1}{2}\left[1 - I\left(\frac{\sqrt{R}S}{2S_0}\right)\right].$$

The relative probability of carrier level S is

$$\frac{dp_S}{dS} = \frac{2S}{S_0^2} \exp\left(-\frac{S^2}{S_0^2}\right).$$

Then,

$$dp = p'dp_S = \left[1 - I\left(\frac{\sqrt{R}S}{2S_0}\right)\right]\frac{S}{S_0^2} \exp\left(-\frac{S^2}{S_0^2}\right)dS,$$

and

$$p = \int_{S=0}^{S=\infty} dp = \frac{1}{2} - \int_0^\infty I\left(\frac{\sqrt{R}S}{2S_0}\right)\frac{S}{S_0^2} \exp\left(-\frac{S^2}{S_0^2}\right)dS.$$

Integration by parts gives

$$p = \frac{1}{2}\left[1 - \sqrt{\frac{R}{R+4}}\right], \tag{3}$$

and, for large R,

$$p = \frac{1}{R}. \tag{4}$$

The second system to be examined is one using envelope detection. This system is admittedly not ideal but is useful for comparison purposes because of its close identification with many practical systems. In Fig. 2, phasor S again represents the carrier amplitude and phase. The envelope detector disregards the phase of its input, however, so that the threshold becomes a circle of radius T. Again, N_1 represents noise received during a carrier omission, and N_2 represents noise received during a carrier transmission.

Case 3—Envelope detection, no carrier fading ($S=S_0$), fixed threshold ($T=aS_0$).

The probability that the noise exceeds T during a carrier omission is

$$p_{N,T} = \exp\left(-\frac{T^2}{N_0^2}\right) = \exp\left(-a^2R\right).$$

Determination of the cumulative probability that noise will cause a detector error during a carrier transmission is a difficult mathematical problem,[11] and the exact solution will not be attempted. We should determine the probability that the resultant of S and N lies within the circular area of radius T in Fig. 2. We shall determine, instead, the probability that the resultant lies within the circumscribed square $ABCD$ in Fig. 2b. The calculation of error based on this probability is somewhat

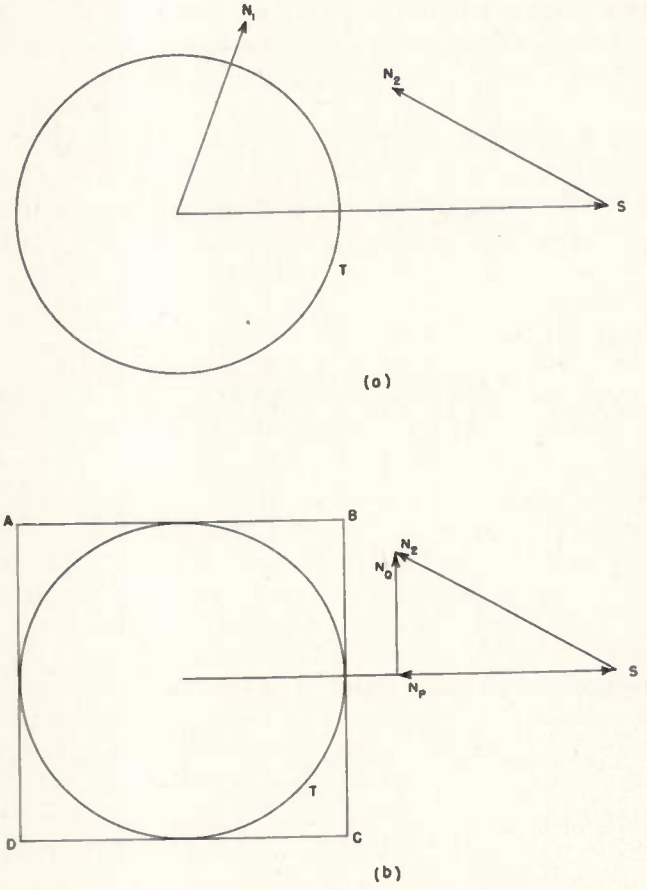

Fig. 2—Carrier and noise phase relations in narrow-band envelope-detection amplitude modulation.

larger than the actual ideal error but serves as a useful approximation. The probability that the resultant lies within $ABCD$ is the product of the probability that N_p is greater than $(S-T)$ but less than $(S+T)$ in the synchronous sense and the probability that N_q is less than T absolutely:

$$\frac{1}{2}\left\{\left[1 - I\left(\frac{S-T}{N_0}\right)\right] - \left[1 - I\left(\frac{S+T}{N_0}\right)\right]\right\} I\left(\frac{T}{N_0}\right)$$

$$= \frac{1}{2}\left\{I[(1+a)\sqrt{R}\,] - I[(1-a)\sqrt{R}\,]\right\} I(a\sqrt{R}).$$

Then, since the two carrier states are equally probable,

$$p \approx \frac{1}{2}\exp\left(-a^2R\right) + \frac{1}{4}\left\{I[(1+a)\sqrt{R}\,] - I[(1-a)\sqrt{R}\,]\right\} I(a\sqrt{R}). \quad (5)$$

Unfortunately, no generally optimum value exists for a, because for minimum p, a is a function of R. For large R, however, p becomes

$$p \approx \frac{1}{2}\left\{1 + \frac{1}{2(1-a)\sqrt{\pi R}}\exp\left[(2a-1)R\right]\right\}\exp\left(-a^2R\right), \quad (6)$$

which is minimum if, as $R \to \infty$, $a \to 1/2$ from above. We conclude, therefore, that for small R, minimum p occurs for $a > 1/2$. If we arbitrarily set $a = 1/2$, (5) and (6) become

$$p \approx \frac{1}{2}\exp\left(-\frac{R}{4}\right) + \frac{1}{4}\left[I\left(\frac{3\sqrt{R}}{2}\right) - I\left(\frac{\sqrt{R}}{2}\right)\right] I\left(\frac{\sqrt{R}}{2}\right), \quad (7)$$

and, for large R,

$$p \approx \frac{1}{2}\left(1 + \frac{1}{\sqrt{\pi R}}\right)\exp\left(-\frac{R}{4}\right). \quad (8)$$

On the basis of (7), the error of an automatic-gain-control envelope-detection system with $T = S/2$ could be calculated in the manner of Case 2. This result would be approximate also. We proceed instead to a system that allows an exact solution.

Case 4—Envelope detection, carrier fading, fixed threshold ($T = aS_0$).

With random carrier fading, the statistics of the carrier amplitude are the same as those of the noise and remain so regardless of the fading rate. Then,

$$p_{N,T} = \exp\left(-\frac{T^2}{N_0^2}\right) = \exp\left(-a^2R\right),$$

and

$$p_{T,N+S} = 1 - \exp\left(-\frac{T^2}{N_0^2 + S_0^2}\right) = 1 - \exp\left(-\frac{a^2R}{1+R}\right).$$

Since the two carrier states are equally probable,

$$p = \frac{1}{2}\exp\left(-a^2R\right) + \frac{1}{2} - \frac{1}{2}\exp\left(-\frac{a^2R}{1+R}\right),$$

which is at minimum for

$$a^2 = \frac{1+R}{R^2}\ln(1+R), \quad (9)$$

where ln indicates the logarithm to the base e. Equation (9) defines the optimum threshold, for which

$$p = \frac{1}{2}\left[1 - \frac{R}{(1+R)^{(1+R)/R}}\right]. \tag{10}$$

For large R,

$$p = \frac{1}{2}\left[1 - \left(\frac{1}{R}\right)^{1/R}\right] \approx \frac{1}{2R} \ln R. \tag{11}$$

Frequency Modulation

Assume that the modulated frequency is measured using the undeviated carrier frequency as a reference. In Fig. 3, the undeviated carrier is represented by a stationary phasor C, and the deviated carrier is represented by S. If the deviated frequency is greater than the center frequency, S rotates counterclockwise, completing one revolution in a period equal to the reciprocal of the deviation. The resultant of S and an added noise, N, is the phasor D. The phase of this resultant, measured with respect to C, is represented by θ.

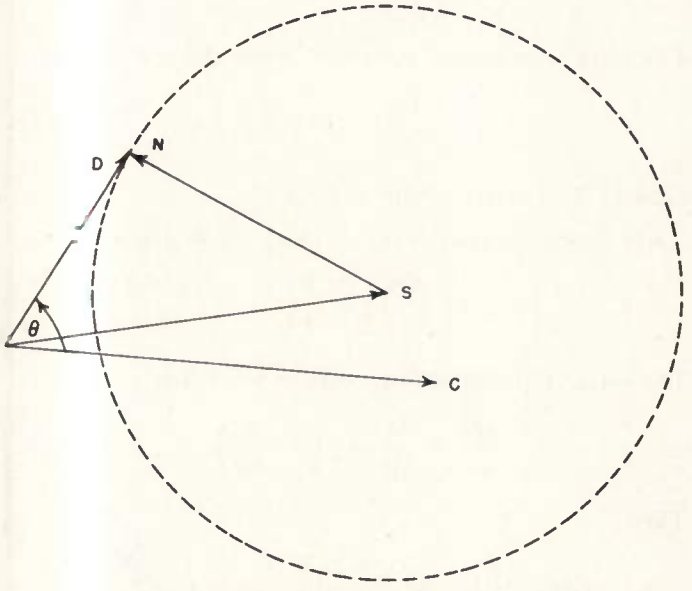

Fig. 3—Carrier and noise phase relations in narrow-band frequency modulation.

The instantaneous frequency of the sum of carrier and noise is defined as the center frequency plus $(1/2\pi)(d\theta/dt)$. For the usual system in which the bandwidth is a fraction of the center frequency, no direct importance is attached to measurements of instantaneous frequency that are made in a period of time that is less than one period of the transmitted carrier (although the average of such measurements is important).[14,15] In fact, for most systems, it is the average frequency of several cycles of the transmitted carrier that forms the decision of the device. In any case, it is useful to consider that the center frequency has been subtracted from the frequency to be measured and to confine attention only to the difference, $(1/2\pi)(d\theta/dt)$. The diagram of Fig. 3 represents this situation. If the phase θ advances during one binary interval, the device will decide that the carrier is deviated positively. If θ recedes, the device will decide that the carrier is deviated negatively.

In general, N rotates with respect to S, and the locus of D, with S as a reference, is a curve of radius N. As long as N is less than S, however, the maximum change in θ that N can effect is less than π radians. Consequently, if the phase change of S alone is at least π radians, the net change in θ will be greater than zero and will have the same sign as the phase change of S itself, and no error will occur. The requirement that the phase change of S exceed π radians implies a minimum value of the deviation, for if the deviation is insufficient, the required change of phase will not be completed within the time allotted to one binary interval. The minimum is equal to one-half of the binary symbol frequency, that is, to one-half the reciprocal of the binary interval.

Conversely, if N exceeds S, it is obvious from interchange of labels on the N and S phasors that the net change in θ will have the same sign as the phase change of N, provided that the deviation of N is greater than one-half the binary symbol frequency. In this case, since the sign of the deviation of N agrees, on the average, with the sign of the intended deviation for only half the time, the error is one half.

If N exceeds S, but the deviation of N is less than one-half the binary signaling frequency, the decision of the device will be generally indeterminate and will depend upon the phase difference and amplitude ratio of S and N. In this case, however, positive and negative deviations are still equally probable, although not necessarily determined by the deviation of N alone (it is interesting to note that this ambiguity disappears if the signal deviation is a multiple of the binary symbol frequency), and the error is one half.

The conclusion is that no error can occur as long as S exceeds N, and that if N exceeds S, errors occur for half the time.

Case 5—FM, no carrier fading ($S = S_0$).

$$p_{N,S} = \exp\left(-\frac{S_0^2}{N_0^2}\right) = \exp(-R),$$

and

$$p = \frac{1}{2}\exp(-R). \tag{12}$$

Case 6—FM, carrier fading.

For a given carrier level, S, the error is, from (12),

$$p' = \frac{1}{2}\exp\left(-\frac{S^2}{N_0^2}\right) = \frac{1}{2}\exp\left(-R\frac{S^2}{S_0^2}\right).$$

[14] L. B. Arguimbau and J. Granlund, "The possibility of transatlantic communication by means of frequency modulation;" *Proc. NEC*, vol. III, pp. 644–653; November, 1947.

[15] L. B. Smith and W. E. Bradley, "The theory of impulse noise in ideal frequency-modulation receivers," Proc. I.R.E., vol. 34, pp. 743–751; October, 1946.

The relative probability of carrier level S is

$$\frac{dp_S}{dS} = \frac{2S}{S_0^2}\exp\left(-\frac{S^2}{S_0^2}\right).$$

Then,

$$dp = p'dp_S = \frac{S}{S_0^2}\exp\left[-(1+R)\frac{S^2}{S_0^2}\right]dS$$

and

$$p = \int_{S=0}^{S=\infty} dp = \frac{1}{2(1+R)}. \tag{13}$$

Phase Modulation

Assume that the phase of the carrier is modulated with respect to a fixed reference phase. In Fig. 4, the reference carrier phase is that of the stationary phasor C, and the deviated phase, ϕ, is that of phasor S. The resultant of S and an added noise, N, is the phasor D;

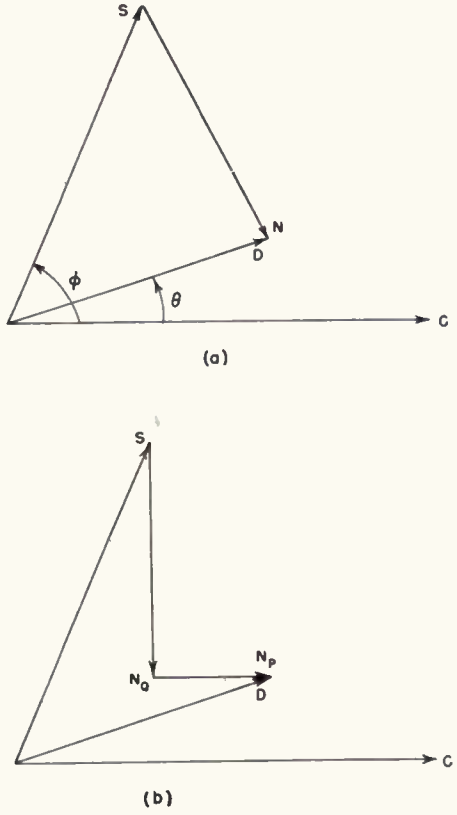

Fig. 4—Carrier and noise phase relations in narrow-band phase modulation.

the phase, θ, of this resultant determines the action of the phase-sensitive device. The signs of ϕ and θ are positive for counterclockwise rotations of their respective phasors. If θ is positive, but less than π radians, the device will judge that the phase of S is positive; if θ is negative, but greater than $-\pi$ radians, the device will judge that the phase of S is negative.

We resolve the noise into two components, N_p and N_q, such that N_p is colinear with C while N_q is in quadrature with C. An error occurs whenever the amplitude and phase of the noise are such as to change θ from positive to negative, or vice versa. It is evident from Fig. 4 that only the quadrature component, N_q, is responsible for such error, and an error is made whenever N_q exceeds $S \sin \phi$ in the direction shown. On the average, N_q will be in the direction shown for only half the time, for it is equally probable that N_q will be directed oppositely to its position in the figure. Errors occur, then, for half the time that N_q exceeds $S \sin \phi$.

Case 7—PM, no carrier fading ($S = S_0$).

$$p_{N_q, S\sin\phi} = 1 - I\left(\frac{S\sin\phi}{N_0}\right) = 1 - I(\sqrt{R}\sin\phi),$$

and

$$p = \frac{1}{2}\left[1 - I(\sqrt{R}\sin\phi)\right]. \tag{14}$$

The error is evidently minimum when $\phi = \pm\pi/2$, or

$$p = \frac{1}{2}\left[1 - I(\sqrt{R})\right]. \tag{15}$$

Case 8—PM, carrier fading, $\phi = \pi/2$.

For a given carrier level, S, the error is, from (15),

$$p' = \frac{1}{2}\left[1 - I\left(\frac{S}{N_0}\right)\right] = \frac{1}{2}\left[1 - I\left(\frac{\sqrt{R}S}{S_0}\right)\right].$$

The relative probability of carrier level S is

$$\frac{dp_S}{dS} = \frac{2S}{S_0^2}\exp\left(-\frac{S^2}{S_0^2}\right).$$

Then,

$$dp = p'dp_S = \left[1 - I\left(\frac{\sqrt{R}S}{S_0}\right)\right]\frac{S}{S_0^2}\exp\left(-\frac{S^2}{S_0^2}\right)dS$$

and

$$p = \int_{S=0}^{S=\infty} dp$$

$$= \frac{1}{2} - \int_0^\infty I\left(\frac{\sqrt{R}S}{S_0}\right)\frac{S}{S_0^2}\exp\left(-\frac{S^2}{S_0^2}\right)dS.$$

Integration by parts gives

$$p = \frac{1}{2}\left[1 - \sqrt{\frac{R}{R+1}}\right] \tag{16}$$

and, for large R,

$$p = \frac{1}{4R}. \tag{17}$$

Nothing has been said concerning phase deviations greater than π radians, for the ideal binary phase-de-

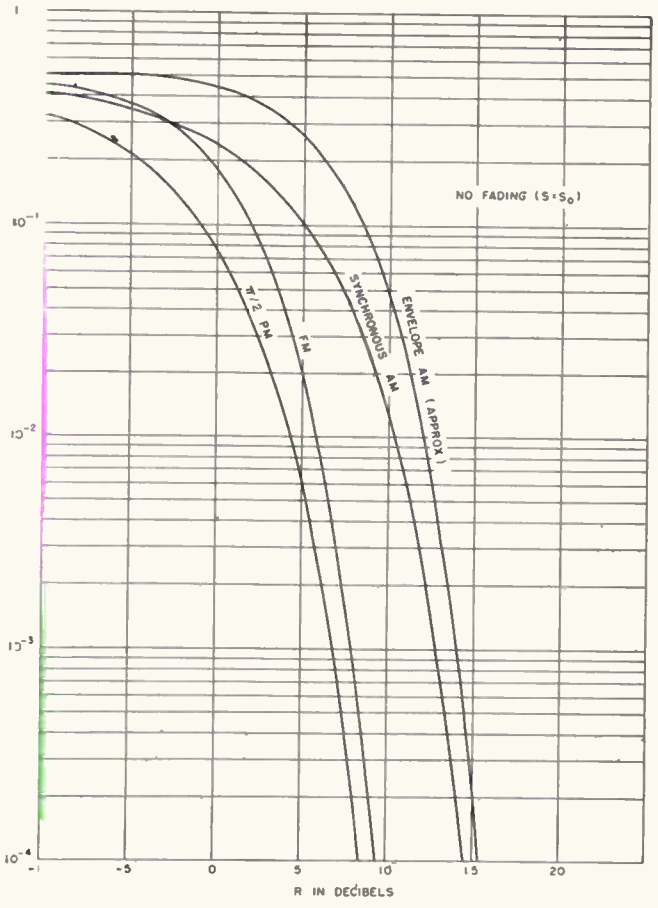

Fig. 5—Narrow-band error vs. average carrier-to-noise ratio, no carrier fading.

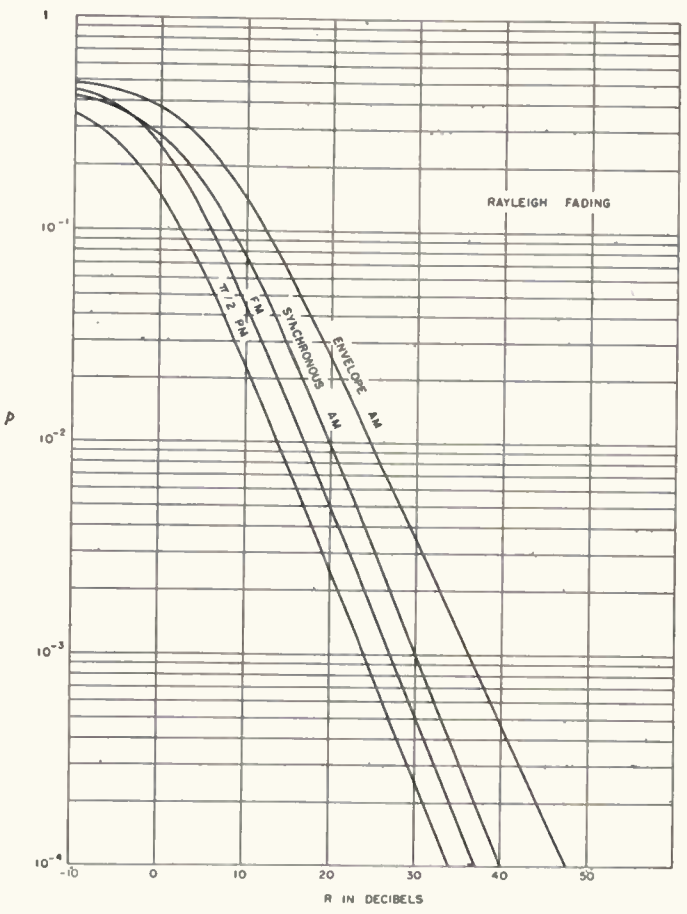

Fig. 6—Narrow-band error vs. average carrier-to-noise ratio, Rayleigh carrier fading.

tecting device is incapable of distinguishing between deviations whose sines have identical signs. For example, as far as the device is concerned, there is no difference between a deviation of $\pi/2$ radians and one of $5\pi/2$ radians. This does not mean that an apparatus cannot be devised that will distinguish between the two, but such an apparatus must examine the behavior of the phase during the transition between the two phase states and thus becomes a frequency-measuring device rather than a phase detector. In practice, a frequency discriminator followed by an integrating network is often used as a phase discriminator. This device is not an ideal device in the sense of this discussion, however, for its action depends on the rate of phase change during the transition between states rather than upon the phase states as such.

Discussion

The various error functions, (1), (3), (7), (10), (12), (13), (15), and (16), are plotted in Figs. 5 and 6. In the practical system sense, these functions apply only to systems wherein the received bandwidth is either equal to, or less than, twice the bandwidth of the equipment following the detecting device. For systems in which the received bandwidth is greater than twice the post-detection bandwidth, the error will be less, for a given carrier-to-noise ratio, than that shown in the curves. The reason for this difference is that some fraction of the noise fluctuations in a wide-band system will produce detecting-device errors that are filtered out by the restricted bandwidth of the equipment following the device.[1,16]

Several interesting facts are revealed in a comparison of the different modulation types. Note that the AM error functions are derived on the basis of peak power; the average AM power is 3 db less than that shown in the curves. In Fig. 5, on the basis of equal average powers and equal bandwidths, FM is superior to synchronous AM for all carrier-to-noise ratios that produce an error less than about one tenth. On the same basis, $\pi/2$ PM is superior to both FM and AM for all carrier-to-noise ratios. At low carrier-to-noise ratios, the envelope-detection error curve rises much more rapidly than the others. This difference is due in part to approximations in the analysis for this case and in part to the assumption of a constant threshold fraction of one half. In the low carrier-to-noise region, a larger threshold fraction would yield lower values for p.

In Fig. 6, for equal average powers and equal bandwidths, FM and synchronous AM are equal in perform-

[16] S. Goldman, "Some fundamental considerations concerning noise reduction and range in radar and communication," Proc. I.R.E., vol. 36, pp. 584–594; May, 1948.

ance for high carrier-to-noise ratios, and $\pi/2$ PM has a 3-db advantage over both FM and AM. For this fading case, the envelope-detection error curve is exact.

Application

For a simple illustration of the effect of binary error on more complex coding, consider the standard five-unit teletype system. Communication of each teletype symbol requires the recognition of a sequence of five binary symbols. Since these may occur in any combination, there are 2^5, or 32, possible sequences. As a practical matter, teletype operation requires also the reception of start and stop pulses which precede and follow each sequence, but it will be assumed for simplicity that the transmitter and receiver remain synchronous when these pulses are omitted or that the only interference of importance occurs during the symbol sequences. A teletype symbol will be received incorrectly whenever at least one of its five binary symbols is in error. If p is the probability of error for a binary symbol, then $(1-p)$ is the probability that a binary symbol will be received correctly. The probability that all five binary symbols of a sequence will be correct is $(1-p)^5$, and the probability of a teletype symbol error is

$$P = 1 - (1 - p)^5$$

or approximately $5p$ as $p \to 0$. Note that the maximum average error rate for teletype, on the basis of perfect synchronization, is 96.875 per cent.

The error, p, is not a direct measure of the loss of communicated information. Information, as defined by Shannon,[6] is given in terms of the binary error by

$$H = 1 + p \log_2 p + (1 - p) \log_2 (1 - p)$$

where H is the received information in bits per binary symbol. The concept of information permits the comparison of systems having different numbers of symbols and different transmitted symbol probabilities, whereas the comparison of such systems on the basis of average system symbol error is meaningless. In the teletype example given above, for instance, the teletype symbol error and the binary symbol error are nowhere equal (except at zero). In terms of received information, however, the two errors are equivalent, provided that all teletype symbols are equally probable.

A General Tchebycheff Rational Function*

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Summary—**The principal result of this paper is the derivation of a general rational function having arbitrary poles which approximates zero with a Tchebycheff characteristic. It is shown how this function can be used to obtain transfer functions which possess a Tchebycheff, low-pass filter characteristic and for which either the poles or zeros are arbitrary. Two applications for this function are given: (1) low-pass RC filters having a specified ripple in the pass band, and (2) low-pass filters having a specified ripple in the pass band and points of infinite attenuation at arbitrary frequencies outside the pass band.**

Introduction

THE TRANSFER FUNCTION of a linear and discrete two-terminal pair network is, in general, a rational function of the form:

$$H(p) = K \frac{p^m + a_1 p^{m-1} + \cdots + a_m}{p^n + b_1 p^{n-1} + \cdots + b_n}, \qquad (1)$$

where p is the complex frequency variable $p = \sigma + j\omega$ and the coefficients a_i, b_i, and K are real constants. The symbol $H(p)$ represents any output-to-input ratio such as $(E_2/E_1)(p)$, $(E_2/I_1)(p)$, etc. The corresponding frequency-response function of the network is given by

$$|H(j\omega)|^2 = [H(p)H(-p)]_{p=j\omega} \qquad (2)$$

$$= K^2 \frac{\omega^{2m} + A_2\omega^{2m-2} + \cdots + A_{2m}}{\omega^{2n} + B_2\omega^{2n-2} + \cdots + B_{2n}}, \qquad (3)$$

where the coefficients are again real constants.

The approximation problem in filter design involves the determination of a rational function of the form (3) which approximates a specified frequency response characteristic over a given band of frequencies and which is realizable as a passive network. The approximation problem for the case of the low-pass filter has a number of well-known solutions. For example, Tchebycheff low-pass response functions having a monotonic cut-off characteristic are easily obtained through the use of Tchebycheff polynomials.[1] The response function

$$|H(j\omega)|^2 = \frac{M}{\gamma + T_{2n}(\omega^2)}, \qquad 1 < \gamma, \qquad (4)$$

where $T_{2n}(\omega^2)$ is the Tchebycheff polynomial of degree $2n$ defined by

$$T_{2n}(\omega^2) = \cos 2n\phi, \qquad \phi = \cos^{-1}\omega, \qquad (5)$$

and M and γ are positive constants, has the low-pass frequency characteristic illustrated in Fig. 1. This follows from the fact that $T_{2n}(\omega^2)$ oscillates between ± 1 in the interval $-1 \leqq \omega \leqq 1$ and approaches $2^{2n-1}\omega^{2n}$ for large ω. The poles of the corresponding transfer function $H(p)$ lie on an ellipse whose major axis lies along the $j\omega$-axis.

There are many applications of low-pass filters requiring a more general solution to the approximation

* Decimal classification: R143.2. Original manuscript received by the Institute, June 16, 1953. This paper represents part of a doctoral thesis entitled "Tchebycheff RC Filters," submitted in the Department of Electrical Engineering, University of Michigan, 1953.

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[1] B. Van der Pol and T. J. Wayers, "Tchebycheff polynomials and their relation to circular functions, Bessel functions, and Lissajous figures," *Physica*, vol. 1, p. 78; 1933.

problem than that represented by the above example. For instance, one may wish to synthesize a Tchebycheff low-pass filter using only RC elements; or he may require a filter with points of infinite attenuation at specified frequencies outside the pass band. Extensions of this sort are impossible with the above solution since one has no control over either the poles or zeros of the

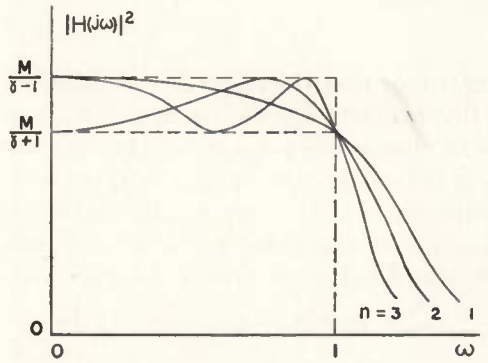

Fig. 1—Tchebycheff low-pass filter response functions.

transfer function. However, both these problems can be solved by means of a general rational function which, like the Tchebycheff polynomial, approximates zero with an equal-ripple characteristic. This general Tchebycheff function, which will now be derived, undoubtedly has many other uses, although only applications to low-pass filters will be considered here. In particular, it will be shown how this function can be used to

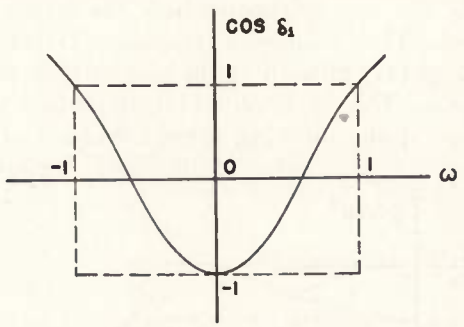

Fig. 2(a)—Plot of cos δ_i versus frequency.

obtain transfer functions which possess a Tchebycheff low-pass filter characteristic and for which either the poles or zeros are arbitrary. The method presented gives an exact Tchebycheff characteristic and has the advantage over other approximate solutions[2,3] of being entirely algebraic.

Derivation of a Tchebycheff Rational Function[4]

Consider the function illustrated in Fig. 2(a),

$$\cos \delta_i = \frac{(2a_i^2 + 1)\omega^2 - a_i^2}{\omega^2 + a_i^2}, \tag{6}$$

where it is assumed for the moment that a_i is a real constant. Corresponding to (6) there is

$$\sin \delta_i = \frac{2a_i\omega\sqrt{(1-\omega^2)(a_i^2+1)}}{\omega^2 + a_i^2}. \tag{7}$$

Equation (6) is analogous to the Tchebycheff polynomial given in (5) for the case of $n=1$. That is, referring to Fig. 2(a), both $\cos \delta_i$ and $\cos 2\phi = 2\omega^2 - 1$ pass through -1 at $\omega = 0$ and through $+1$ at $\omega = \pm 1$. The angle δ_i varies in the same manner as 2ϕ in the interval $-1 \leqq \omega \leqq 1$ if the convention is made that the sign of $\sqrt{1-\omega^2}$ is positive and the sign of $\sqrt{a_i^2+1}$ is the same

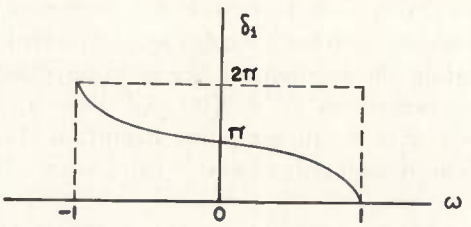

Fig. 2(b)—Plot of δ_i versus frequency.

as the sign of a_i. That is, both δ_i and 2ϕ are symmetric about π and have the same end points as illustrated in Fig. 2(b). The above convention with regard to signs insures that the derivative

$$\frac{d\delta_i}{d\omega} = -\frac{2a_i\sqrt{a_i^2+1}}{(\omega^2 + a_i^2)\sqrt{1-\omega^2}} \tag{8}$$

is always negative and makes δ_i a monotonically decreasing function of ω in the interval $-1 \leqq \omega \leqq 1$.

Now form the function

$$F(\omega^2) = \cos (2m\phi + \delta_1 + \delta_2 + \cdots + \delta_n). \tag{9}$$

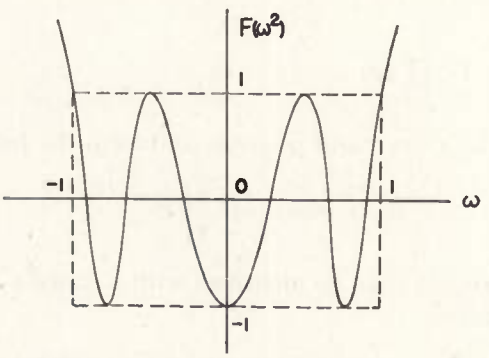

Fig. 3—Plot of $F(\omega^2)$ versus ω for $m+n=3$.

It follows that $F(\omega^2)$ oscillates between ± 1 with $2(m+n)+1$ points of maximum deviation in the interval $-1 \leqq \omega \leqq 1$ since the argument passes through $2(m+n)\pi$ radians in the same interval. The variation of $F(\omega^2)$ with ω is illustrated in Fig. 3 for the particular case of $m+n=3$. $F(\omega^2)$ can be expanded in the form

$$F(\omega^2) = \frac{C_0\omega^{2(m+n)} + C_2\omega^{2(m+n)-2} + \cdots + C_{2(m+n)}}{(\omega^2 + a_1^2)\cdots(\omega^2 + a_n^2)} \tag{10}$$

[2] E. A. Guillemin, "Synthesis of RC-networks," *Jour. Math. Phys.*, vol. 28, pp. 22–42; April, 1949.

[3] G. L. Mattaei, "Filter transfer function synthesis," Proc. I.R.E., vol. 41, pp. 377–382; March, 1953.

[4] This derivation is suggested by a somewhat different treatment given in S. Bernstein, "*Leçons sur les Propriétés Extrémales et La Meilleure Approximation des Fonctions Analytiques d'une Variable Reelle*," Gauthier-Villars, Paris, pp. 1–15; 1926.

by noting that

$$F(\omega^2) = \tfrac{1}{2}[(\cos\phi + j\sin\phi)^{2m}(\cos\delta_1 + j\sin\delta_1)\cdots(\cos\delta_n + j\sin\delta_n) + (\cos\phi - j\sin\phi)^{2m}(\cos\delta_1 - j\sin\delta_1)\cdots(\cos\delta_n - j\sin\delta_n)] \quad (11)$$

$$= \frac{(\omega + \sqrt{\omega^2-1})^{2m}[(2a_1^2+1)\omega^2 - a_1^2 + 2a_1\omega\sqrt{(\omega^2-1)(a_1^2+1)}]\cdots + (\omega - \sqrt{\omega^2-1})^{2m}[(2a_1^2+1)\omega^2 - a_1^2 - 2a_1\omega\sqrt{(\omega^2-1)(a_1^2+1)}]\cdots}{2(\omega^2+a_1^2)\cdots(\omega^2+a_n^2)}. \quad (12)$$

Functions which oscillate in the manner illustrated in Fig. 3 with a sufficient number of points of maximum deviation are known as Tchebycheff functions of best approximation to zero in the interval $-1 \leqq \omega \leqq 1$. It can be shown $F(\omega^2)$ is a Tchebycheff function of best approximation to zero for complex as well as real values of a_i, excluding the segment of the imaginary axis $-j \leqq p \leqq j$. The coefficients C_{2i} will be real if the a_i are taken either real or in conjugate pairs. Equation (12) reduces to the Tchebycheff polynomial in (5) when $n=0$. That is,

$$T_{2m}(\omega^2) = \tfrac{1}{2}[(\omega + \sqrt{\omega^2-1})^{2m} + (\omega - \sqrt{\omega^2-1})^{2m}]. \quad (13)$$

Application to RC Filters

The general function $F(\omega^2)$ can be used to obtain Tchebycheff, low-pass RC filters having a specified ripple in the pass band. Set $m=0$ in (12). Then the coefficients C_{2i} in (10) are given as follows:

$$C_0 = \frac{1}{2}\left[\prod_{i=1}^{n}(a_i+\sqrt{a_i^2+1})^2 + \prod_{i=1}^{n}(a_i-\sqrt{a_i^2+1})^2\right] \quad (14)$$

$$C_2 = -\frac{1}{2}\left[\prod_{i=1}^{n}(a_i+\sqrt{a_i^2+1})^2 \sum_{j=1}^{n}\frac{a_j}{a_j+\sqrt{a_j^2+1}} + \prod_{i=1}^{n}(a_i-\sqrt{a_i^2+1})^2 \sum_{j=1}^{n}\frac{a_j}{a_j-\sqrt{a_j^2+1}}\right] \quad (15)$$

$$\vdots$$

$$C_{2n} = (-1)^n \prod_{i=1}^{n} a_i^2. \quad (16)$$

Take the a's real and positive, and form the function

$$|H(j\omega)|^2 = \frac{C_0}{C_0 - 1}\left[1 - \frac{F(\omega^2)}{C_0}\right]. \quad (17)$$

Equation (17) can be identified with a transfer function of the form

$$H(p) = K\frac{c_0 p^{n-1} + c_1 p^{n-2} + \cdots + c_{n-1}}{(p+a_1)\cdots(p+a_n)}, \quad (18)$$

where the degree of the denominator exceeds the degree of the numerator by one. Since all the poles lie on the negative real axis. $H(p)$ can be synthesized as an RC or RL network. Several typical response curves for this function are sketched in Fig. 4. $|H(j\omega)|^2$ approaches zero monotonically for large ω since, for the case of $m=0$ and a_i real, $F(\omega^2)$ approaches monotonically the constant C_0.

It is apparent that the ripple or deviation in the pass band is determined once the coefficient C_0 is chosen. It remains to choose the poles a_i in (14) so that the desired C_0 is obtained. The solution to this problem is far from unique since $n-1$ of the n poles are arbitrary. The determination of the poles for a given ripple is considerably simplified if one orders the poles according to

$$a_k = \sinh k\alpha, \quad k = 1, 2, \cdots, n. \quad (19)$$

It follows from (14) that

$$C_0 = \cosh n(n+1)\alpha. \quad (20)$$

The new variable α can be determined from

$$\tanh^2\left[n(n+1)\frac{\alpha}{2}\right] = \frac{C_0 - 1}{C_0 + 1}, \quad (21)$$

where n is the number of poles.

Although RC filters can be designed by this method to give a low-pass characteristic with any pre-assigned tolerance, one has no control over the response in the stop-band. The frequency response $|H(j\omega)|$ corresponding to (17) falls off at the rate of 6 db per octave for large ω. The question naturally arises as to the advantage, if any, of using a large number of poles. If

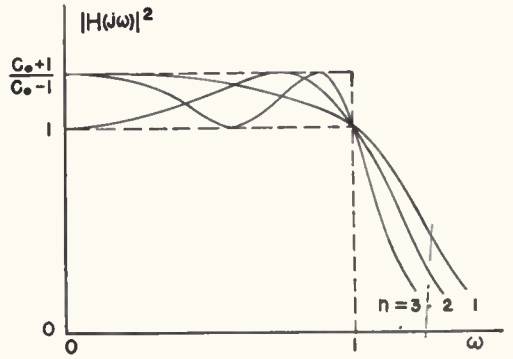

Fig. 4—Typical RC filter response functions.

one defines the theoretical gain of the filter as the maximum value of $|H(j\omega)|$ when the constant multiplier K in (3) is unity, then it can be shown that increasing n increases the gain for a given ripple. The relative improvement in gain with increasing n becomes greater as the required ripple decreases. However, there is little advantage to increasing n beyond 4 or 5 unless a ripple of 1 db or less is required. The principal use for this type filter seems to be in low-frequency applications where tolerance in the pass band is a critical factor.

Application to Filters with Arbitrary Zeros

The Tchebycheff rational function $F(\omega^2)$ can be used to obtain low-pass response functions with arbitrary zeros outside the pass band. Consider the function

$$|H(j\omega)|^2 = \frac{M}{\gamma + F(\omega^2)}, \quad 1 < \gamma, \tag{22}$$

where M and γ are positive constants. This function oscillates between $M/(\gamma-1)$ and $M/(\gamma+1)$ in the interval $-1 \leqq \omega \leqq 1$ and has arbitrary zeros at $\omega = \pm ja_i$. The response $|H(j\omega)|$ falls off at the rate of 6m db per octave for large ω since $|H(j\omega)|^2$ has the form:

$$|H(j\omega)|^2 = M \frac{(\omega^2 + a_1^2) \cdots (\omega^2 + a_n^2)}{B_0\omega^{2(m+n)} + \cdots + B_{2(m+n)}}. \tag{23}$$

Equation (23) satisfies the necessary condition that $0 \leqq |H(j\omega)|^2$ for all ω since, in general, $0 < F(\omega^2)$ for $1 < \omega^2$. This follows from the fact that all $2(m+n)$ zeros of $F(\omega^2)$ lie on the segment $-1 \leqq \omega \leqq 1$.

As an example set $m=1$ in (12), and assume $a_1 = jb$ and $a_2 = -jb$, where b is any real number greater than unity. The resulting response function has a zero at $\omega = b$ as shown in Fig. 5. The corresponding transfer function will not, in general, be realizable as an RC filter since one has no control over the location of the poles.

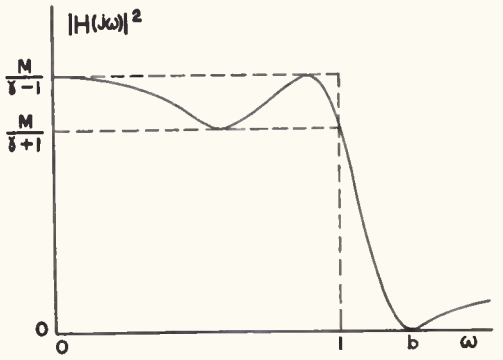

Fig. 5—A Tchebycheff filter response having a zero at $\omega = b$.

However, this method should be useful in designing Tchebycheff LC filters with points of infinite attenuation at real or complex frequencies outside the pass band.

Acknowledgment

The author wishes to thank Professor A. B. Macnee for his helpful suggestions concerning this research.

Frequency Analysis of Digital Computers Operating in Real Time*

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Summary—When used in a control system, the digital computer operates in real time and interacts dynamically with other parts of the system. It is desirable to examine its effect by the use of conventional techniques of servo-mechanism design, namely, frequency analysis.

An important feature of most digital computers is that the data passing through them are in sampled form. Frequency analysis has already been applied to the process of sampling and sampled-data systems.[1] The present paper extends this approach to operations performed on the data in sampled form, such as by a digital computer. A real-time linear digital-computer program can be represented by a transfer function. Thus, the computer with its program is a filter whose analysis and design can proceed along lines familiar in network theory and servo design.

The error analysis of familiar numerical formulas is illustrated. Various properties of programs are studied in both the s-domain (complex frequency) and the z-domain (complex delay). Tests for stability (or convergence) are described and illustrated. Programming is discussed from a new point of view, and a new programming method is developed. It is indicated how such studies could conceivably affect the logical design or instruction code of a computer.

I. Introduction

IN A RECENT ARTICLE[1] an outline was given for the analysis of control systems in which information is sampled at a certain point. It was noted that a digital computer of adequate speed could operate on the data in sampled form and that frequency analysis could be applied to these computer operations in the linear case. An example of simple numerical differentiation was also given.

The present article is intended to enlarge upon the methods of frequency analysis as applicable to digital computer programs. Except for papers[2] presented, this article is the first published report summarizing the work done some time ago.[3] This work was motivated by the need for analyzing control systems consisting of both digital and analog units. The cardinal theme of this investigation is that a real-time linear program can be described by a difference equation whose transform

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[1] W. K. Linvill and J. M. Salzer, "Analysis of control systems involving digital computers," PROC. I.R.E., vol. 41, pp. 901–906; July, 1953.

[2] Among papers presented: W. K. Linvill and J. M. Salzer, "Analysis of digital computers in control systems," delivered at the Wayne Computer Conference, Detroit, Mich.; March 28, 1951; and J. M. Salzer, "Frequency analysis of digital computers used in control systems," delivered at 1952 I.R.E. National Convention.

[3] J. M. Salzer, "Treatment of Digital Control Systems and Numerical Processes in the Frequency Domain," thesis for Sc.D. degree, Massachusetts Institute of Technology; 1951.

leads to a transfer function characterizing the action of the digital computer. The digital computer with its program is a filter (a discrete or sampled-data filter) which can be analyzed by frequency techniques as conventional filters (analog or continuous-data filters).

II. Example

The example of the previous article[1] can serve as the background to the present discussion. The block diagram of Fig. 1(a) may represent one of many systems controlled by the same digital computer, one channel of a complex system, or a simple system for which a special-purpose digital computer may be justified.

At the input and output of the digital computer the data must be converted. The input device performs two functions: (a) it changes the analog representation to digital one (e.g., voltage to binary number); (b) it changes the continuous function into a sequence of samples. The first process introduces static errors, mostly due to quantization, which can be analyzed as other noise effects. The second function, the sampling process, affects the dynamics of the system and is often a primary consideration in the study of closed-loop systems. It is this second aspect of a digital-analog system which is emphasized in this paper.

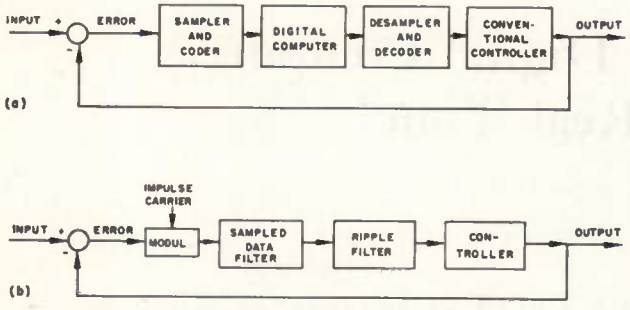


Fig. 1—Equivalent representation of a simple control system incorporating a digital computer.

In the equivalent diagram in Fig. 1(b) the quantizing effect is disregarded, and the samples are replaced by impulses having corresponding areas. This equivalent representation, previously discussed,[1] means that the digital computer is pictured as operating on sequences of impulses having areas equal to the actual numerical samples. This one-to-one correspondence between impulses and samples is maintained throughout the computations as well as in the result. The chosen representation permits the use of frequency analysis and the adaptation of this analysis to methods already established for sampled-data systems.[4,5]

III. Definition of Objective

In what follows attention is paid only to that block of the above system which corresponds to the digital computer and its program. The methods described have a wide applicability, even though the following assumptions are made in analysis: (*a*) Data are sampled at a constant rate; i.e., the time T between consecutive samples is constant. (*b*) The time delay due to computation is zero or negligible. (This assumption is often realistic, or the delay can be included in the analysis.) (*c*) The computer relates a single output quantity to a single input quantity. (*d*) The computer performs linear operations, defined below. (*e*) The computer operates in real time; so, it can use only past and present information.

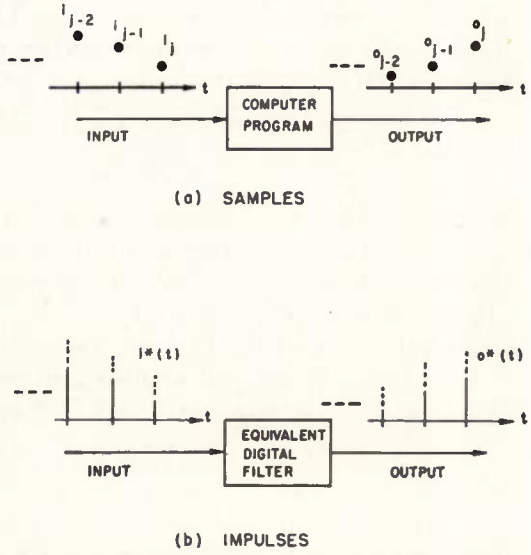


Fig. 2—Representation of input and output sequences.

With these limitations imposed, our problem is to deal with the situation illustrated in Fig. 2(a). The computer input consists of a sequence of samples and so does the output. To obtain the present output sample o_j, the computer can use the present input sample i_j, and a finite number of past input and output samples. Thus,

$$o_j = a_0 i_j + a_1 i_{j-1} + \cdots + a_m i_{j-m} - b_1 o_{j-1} - \cdots - b_n o_{j-n} = \sum_{k=0}^{m} a_k i_{j-k} - \sum_{k=1}^{n} b_k o_{j-k}, \quad (1)$$

where a_k and b_k are constants,[6] and the notation is the usual one of numerical work.

Replacing each sample by an impulse of corresponding area, and calling the resulting impulse sequences $i^*(t)$ and $o^*(t)$, respectively, changes (1) to

$$o^*(t) = \sum_{k=0}^{m} a_k i^*(t - kT) - \sum_{k=1}^{n} b_k o^*(t - kT). \quad (2)$$

The operations indicated in (1) or (2) are well suited to a digital computer. Its memory permits introduction of arbitrary *ideal* delays and storage of constants, and its arithmetic unit performs multiplications and additions.

IV. Transfer Function or Program

The transfer function of a filter can be defined as the ratio of its output transform to its input transform. If

[4] W. K. Linvill, "Sampled-data control systems studied through comparison of sampling with amplitude modulation," *Trans. AIEE*, vol. 70, pp. 1779–1788; 1951.

[5] J. R. Ragazzini and L. A. Zadeh, "The analysis of sampled-data systems," *Trans. AIEE*, part II, pp. 225–234; 1952.

[6] They could be time-variant functions and still permit linear analysis, but the resulting complication would be beyond the aims of this article.

$I^*(s)$ and $O^*(s)$ denote the Laplace transforms of $i^*(t)$ and $o^*(t)$ respectively, the transform of (2) becomes

$$O^*(s) = \sum_{k=0}^{m} a_k e^{-kTs} I^*(s) - \sum_{k=1}^{n} b_k e^{-kTs} O^*(s) \qquad (3)$$

and yields the transfer function

$$W^*(s) = \frac{O^*(s)}{I^*(s)} = \frac{\sum_{k=0}^{m} a_k e^{-kTs}}{1 + \sum_{k=1}^{n} b_k e^{-kTs}}$$

$$= \frac{a_0 + a_1 e^{-Ts} + \cdots + a_m e^{-mTs}}{1 + b_1 e^{-Ts} + \cdots + b_n e^{-nTs}}. \qquad (4)$$

Within the limits of our assumptions, (1) was the most general program possible, and it led to the transfer function (4). Furthermore, the steps of derivation can be retrotraced from (4) to (1). It follows, therefore, that a *linear, real-time program can be realized if, and only if, its transfer function can be written in the form* (4). (Realizability theorem.)

While the transfer function of an analog (continuous-data) filter is rational in s, that of a digital (sampled-data) filter is seen to be rational in e^{-Ts}. The extension of conventional frequency techniques to the latter fulfills an important need in dealing with mixed systems.

V. Bandwidth Limitation

When a continuous function is sampled, information is generally lost, because more than one continuous curve can be drawn through these samples. However, *a unique correspondence between a continuous curve and its sampled form is maintained if the band of the former is limited to one half of the sampling frequency or 1/2T cps.* (Sampling theorem.)[7,8]

Actually, a sampling rate just higher than this limit permits the reconstruction of the signal only at the expense of considerable time delay. In an open-loop system, such as audio transmission, time delays of several cycles of the highest signal frequency may be inconsequential; however, in a closed-loop system, delays of even a fraction of a cycle (10 or 20 degrees) can be critical. In a control system a sampling frequency of, at least, five times the highest signal frequency is likely to be required. While the sampling theorem establishes an ultimate limit, it must not be construed as a working rule for sampled-data control systems.

The bandwidth limitation of sampling is maintained throughout the computer, since the output of the computer is as sampled as its input. To prove this directly, we recognize two properties of $W^*(s)$: conjugate symmetry and periodicity.

The transfer function $W^*(s)$ possesses conjugate symmetry because any of the terms $a_k e^{-kTs}$ in $W^*(s)$ possesses conjugate symmetry, for

$$e^{-kT(\sigma \pm j\omega)} = e^{-kT\sigma}(\cos kT\omega \mp j \sin kT\omega). \qquad (5)$$

Consequently, if $W^*(s)$ is known at positive real frequencies, it is also known at negative real frequencies.

Furthermore, e^{-kTs} is periodic in the frequency domain, having a repetition interval of $j\Omega = j2\pi/T$, because (with m an integer)

$$e^{-kT(s+jm\Omega)} = e^{-jkmT\Omega} \cdot e^{-kTs} = e^{-jkm2\pi} \cdot e^{-kTs} = e^{-kTs}. \qquad (6)$$

It follows that $W^*(s)$ has the same periodicity. Thus, $W^*(s)$ is defined completely if it is defined in the strip of the s-plane bounded by the lines $\pm j\Omega/2$. However, because of its conjugate symmetry, $W^*(s)$ *needs to be defined only in the strip 0 to* $+j\Omega/2$, which is the same *limitation on bandwidth* as that established by the sampling theorem. For example, a program cannot have low-pass characteristics in the true sense. A smoothing program has low-pass characteristics in the range 0 to $\Omega/2$ radians/sec; i.e., it has a pass band near zero frequency. Consequently it will have pass bands also at all multiples of Ω. These higher frequencies can be removed ony by continuous-data filters.

If a digital computer is used in a system, the highest effective[9] sampling rate possible is the rate at which the required computation is carried out. Thus, a faster computer program can be said to have a greater bandwidth. The speed of a program depends on several factors: the speed of the individual computing components (pulse repetition rate), the logical design, the instruction code, the computational method and the coding efficiency. The computing task can often be simplified by programming with the lowest permissible accuracy. The balance between static (roundoff) errors and dynamic (truncation) errors[10] deserves the same emphasis as in the design of analog units. All these considerations influence the applicability of the computer to a particular control problem.

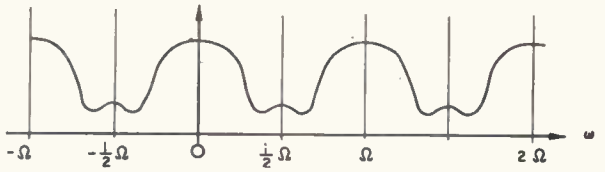
VI. Amplitude, Phase, Locus

If in (4) s is replaced by $j\omega$, the transfer characteristics $W^*(j\omega)$ of the program is obtained at real frequencies. The physical interpretation of $W^*(j\omega)$ is similar to that of the analog case. A sinusoidal input sequence of angular frequency ω_1 (where ω_1 is less than $\Omega/2$), when operated upon by the program $W^*(j\omega)$, results in a sinusoidal output sequence of the same frequency having $|W^*(j\omega_1)|$ times the amplitude of the input and a phase shifted by the angle $\angle W^*(j\omega_1)$. In view of (4) the proof is trivial. The thing to note is that the input and output functions are sampled (impulsed) and contain, in addition to the frequency $\pm\omega_1$, also $\pm\omega_1 \pm \Omega$, $\pm\omega_1 \pm 2\Omega$, etc., in equal strength.

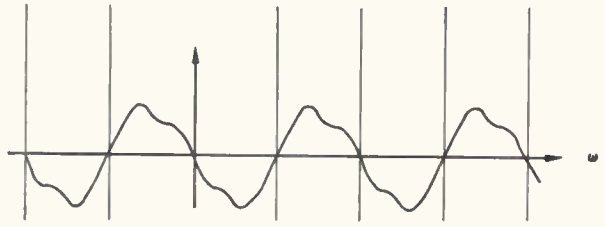
[7] B. M. Oliver, J. R. Pierce, and C. E. Shannon, "The philosophy of pcm," Proc. I.R.E., vol. 36, pp. 1330–1331; November, 1948.
[8] C. E. Shannon, "Communication in the presence of noise," Proc. I.R.E., vol. 37, pp. 11–12; January, 1949.

[9] The input conversion device might have its own sampling rate, which is usually higher than the computing rate. The *effective* sampling rate is the rate at which the program makes use of the input information.
[10] J. M. Salzer, "Fundamental characteristics of digital and analog units," *Proc. NEC* (Chicago), vol. VIII, pp. 621–628; 1952; also *Radio-Electronics Eng.*, pp. 13–15 and 30; February, 1953.

Fig. 3 shows possible characteristics of a program. The amplitude is an even, the phase an odd function of frequency because of conjugate symmetry, and both patterns repeat every Ω radians/sec because of periodicity.



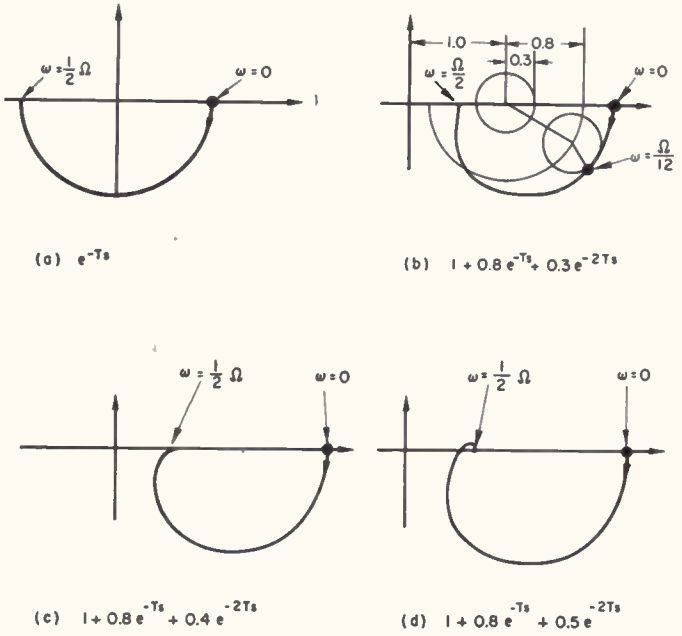
(a) AMPLITUDE



(b) PHASE

Fig. 3—Possible frequency characteristics of a program.

The amplitude and phase characteristics may be combined into a polar plot, called the locus, which can often be sketched rapidly, as illustrated in Fig. 4. In part (a) the locus of e^{-Ts} is shown as the unit circle, as the magnitude of $e^{-jT\omega}$ is unity and its phase is $-T\omega$. Only the range $O \leqq \omega \leqq \Omega/2$ is plotted; the range $\Omega/2 \leqq \omega \leqq \Omega$ would complete the circle, which would repeat itself thereafter. The other three illustrations of Fig. 4 are loci of second-degree polynomials in e^{-Ts}. The construction, illustrated for frequency $\omega = \Omega/12$ (or $\omega T = 30$ degrees) in (b), is based on the principle of rotating vectors.



(a) e^{-Ts}

(b) $1 + 0.8e^{-Ts} + 0.3e^{-2Ts}$

(c) $1 + 0.8e^{-Ts} + 0.4e^{-2Ts}$

(d) $1 + 0.8e^{-Ts} + 0.5e^{-2Ts}$

Fig. 4—Simple examples of program loci.

The construction of loci is facilitated by the following rules, stated without proof:[11]

1. The locus repeats for each range of Ω.
2. The locus possesses conjugate symmetry (the real axis acts as a mirror).
3. At $\omega = 0$ and $\omega = \Omega/2$, the locus crosses the real axis either perpendicularly (Fig. 4(a), (b), and (d)) or tangentially (Fig. 4(c)).
4. With $W^*(s)$ defined by (4), its locus can cross the real axis at most $m-1$ or $n-1$ times (whichever is greater) in the range $0 < \omega < \Omega/2$ (excluding the ends of this range).

VII. Prediction Programs

Classical prediction formulas will serve to illustrate the preceding sections. Suppose the digital computer is to predict the value of the next input sample; ideally it should solve the difference equation

$$o^*(t) = i^*(t + T). \tag{7}$$

Transformation of (7) yields the desired program transfer function,

$$W^*(s) = \frac{O^*(s)}{I^*(s)} = e^{+Ts}, \tag{8}$$

which is not of the realizable form (4).

Classical extrapolation formulas based on polynomial approximations perform approximate prediction. The transforms of the four simplest ones are

$$\begin{aligned} W_0^*(s) &= 1, \\ W_1^*(s) &= 2 - e^{-Ts}, \\ W_2^*(s) &= 3 - 3e^{-Ts} + e^{-2Ts}, \\ W_3^*(s) &= 4 - 6e^{-Ts} + 4e^{-2Ts} - e^{-3Ts}. \end{aligned} \tag{9}$$

To interpret these, replace $W_m^*(s)$ by $O_m^*(s)/I_m^*(s)$, multiply $I_m^*(s)$, and perform the inverse transformation to get the equations

$$\begin{aligned} o_0^*(t) &= i^*(t) \\ o_1^*(t) &= 2i^*(t) - i^*(t-T) \\ o_2^*(t) &= 3i^*(t) - 3i^*(t-T) + i^*(t-2T), \\ o_3^*(t) &= 4i^*(t) - 6i^*(t-T) + 4i^*(t-2T) - i^*(t-3T). \end{aligned} \tag{10}$$

No use is made of previous output samples in any of the above equations; that is, previous predictions are disregarded in making the new prediction. This is why each $W_m^*(s)$ of (9) has a unity denominator.

The loci of the transfer functions (9) are all shown in Fig. 5. The approximation to the ideal characteristics seems to improve as the order of the formula increases. Actually, the comparison is made only at low frequencies; indeed, at $\omega = 0$ all are ideal. As the frequency increases, each locus departs from the ideal one. It turns

[11] For proof see ref. 3, pp. 75, 76 and 265–267, of a Hughes Res. and Devel. Labs reprint (available on request) from which the present article was reduced.

out that the loci shown stay outside of the unit circle, the ideal locus, for all frequencies, which indicates that they can be improved for specified signal spectra.[12]

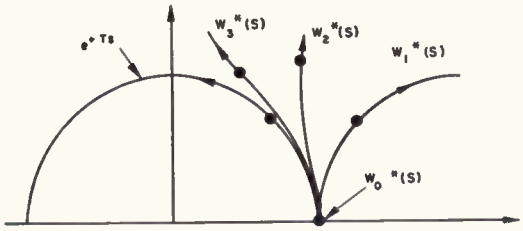


Fig. 5—Loci of extrapolation formulas compared with ideal prediction.

To compare these formulas at higher frequencies consider the frequency $\omega = \Omega/2$ radians/sec. It is readily shown that $W_m^*(j\Omega/2)$ is $2^{m+1}-1$, so that at this frequency the gain goes up rapidly with the order m of prediction. A more complex prediction formula, therefore, accentuates the noise problem, including noise due to quantization or roundoff. As in the analog case, one is faced with a compromise between good prediction and good smoothing.

VIII. The Complex Delay Variable, Infinite Weighting Sequences

It is expedient to introduce a new symbol z, defined:

$$z = e^{-Ts}. \tag{11}$$

This is more than notational convenience. First, z is a complex variable related to the complex frequency variable s, and much can be learned about computer programs (and sampled functions) from conformal studies in the two planes. Second, z can be given a definite physical meaning.

Fig. 4(a) has already illustrated that the imaginary axis of the s-plane is conformal to the unit circle of the z-plane, with the right half s-plane corresponding to the inside of the unit circle. Note that the z-plane circle is traversed once for each interval $j\Omega$ traveled in the s-plane. To find the region conformal to the *finite* right

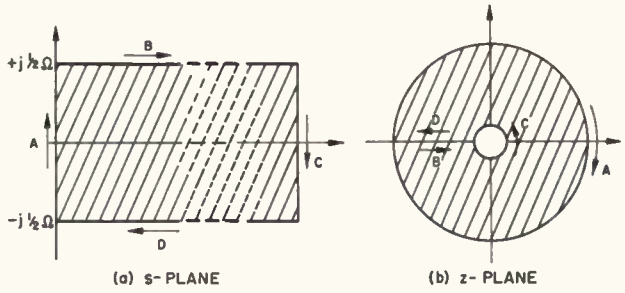


Fig. 6—Conformality with right-half plane.

half s-plane, it is sufficient to surround the semi-infinite strip of Fig. 6(a). The conformal region in Fig. 6(b) excludes the origin of the z-plane.

For another example of conformality consider a possible zero-pole configuration of a digital filter (or a sampled function). The s-plane pattern, Fig. 7(a), repeats at intervals $j\Omega$, but there is only a single conformal pattern in the z-plane, Fig. 7(b). This indicates why for sampled functions and discrete filters, the use of the variable z is often preferable. For mixed systems, described by a mixture of rational functions of both s and z, only the variable s has a unique meaning. In such cases the variable z has been used,[13,5] but the approximation must be justified.

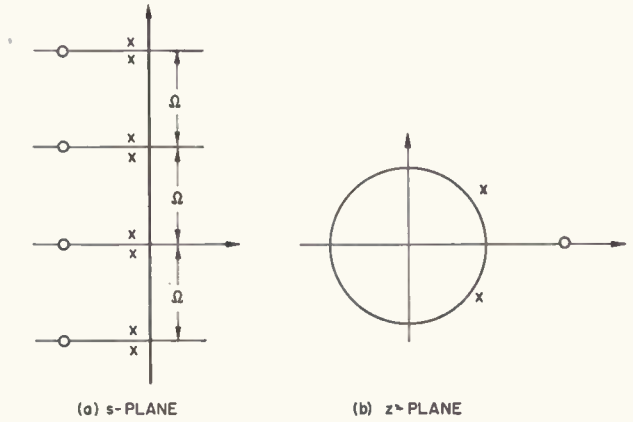


Fig. 7—Conformality of zero-pole configurations.

The variable z can also be considered an operator, since multiplication by z corresponds to a delay T in the time domain.[14] This difference operator is analogous to the differential operator, p or D. Considering these operators as complex variables opens the way to greater generality.

In terms of z the transfer function (4) of a program is written as

$$W^*(z) = \frac{O^*(z)}{I^*(z)} = \frac{a_0 + a_1 z + a_2 z^2 + \cdots + a_m z^m}{1 + b_1 z + b_2 z^2 + \cdots + b_n z^n}. \tag{12}$$

Since cross-multiplication yields

$$(1 + b_1 z + \cdots + b_n z^n) O^*(z) = (a_0 + a_1 z + \cdots + a_m z^m) I^*(z), \tag{13}$$

the numerator of (12) is seen to be a weighting sequence applied to the input, while the denominator refers to output samples.

A finite weighting sequence referring to previous outputs can always be replaced by an infinite weighting sequence referring only to input samples. For example, if

$$W^*(z) = \frac{1}{1 - bz}; \tag{14}$$

then

$$O^*(z) = I^*(z) + bzO^*(z), \tag{15}$$

which can be written as

[12] It can be shown that these formulas are optimum in the mean-square sense *only* for a signal band approaching zero width at zero frequency, which is a trivial case.

[13] A Tustin, "A method of analyzing the behavior of linear systems in terms of time series," *Jour. IEE* (London), vol. 94, part 2A, pp. 130–142; May, 1947.

[14] In the literature the operator (E, ρ, or z) is usually defined as a time advance operator and corresponds to our $1/z$. In a real-time application time advance is an unrealizable operation; hence our choice. This choice leads also to other simplifications in the analysis.

$$O^*(z) = I^*(z) + bzI^*(z) + b^2z^2O^*(z). \quad (16)$$

Continuance of this processs leads to

$$O^*(z) = (1 + bz + b^2z^2 + b^3z^3 + \cdots)I^*(z), \quad (17)$$

which makes

$$W^*(z) = \frac{O^*(z)}{I^*(z)} = 1 + bz + b^2z^2 + b^3z^3 + \cdots . \quad (18)$$

The latter result could also have been obtained directly from (14) by long division.

Levinson[15] has dealt with the optimization of discrete filters in terms of infinite weighting sequences applied to input samples only. In a practical case such a sequence is truncated and corresponds to a polynomial transfer function, which is a special case of the rational function (12).

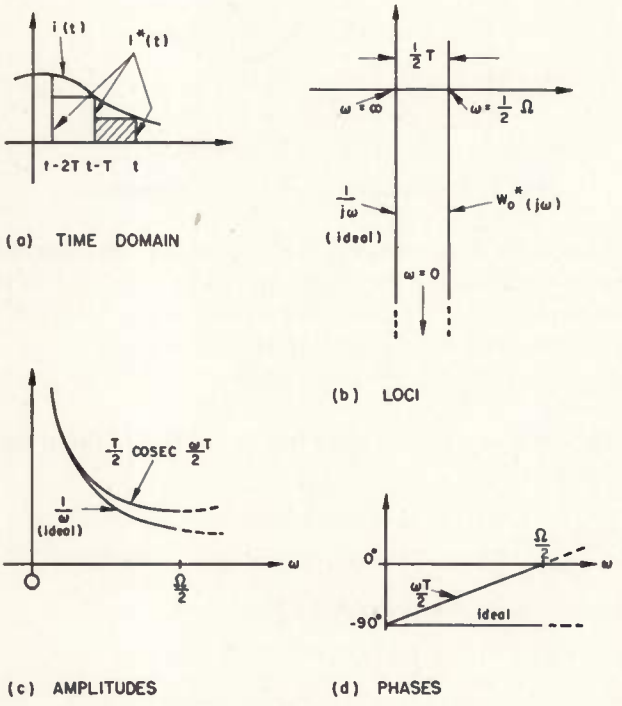


Fig. 8—Characteristics of the rectangular rule.

IX. Integration

If the transfer function (12) has a denominator, long division may change it into an infinite weighting sequence. Thus, the denominator signifies an infinite (though usually failing) memory. Integration formulas should be good examples, because the memory of an ideal integrator is unfailing. Since the transfer function of the ideal integrator is $1/s$, which is evidently not of the form of (4), a program can only approximate this.

The integral of, or the area under, a continuous curve can be approximated by rectangles, as shown in Fig. 8(a). The area is brought up-to-date by adding the latest rectangle to the area already accumulated. Thus, the integral $o^*(t)$ is computed by

$$o^*(t) = T \cdot i^*(t) + o^*(t - T), \quad (19)$$

which, after transformation, leads to the program transfer function,

$$W_0^*(z) = \frac{O^*(z)}{I^*(z)} = \frac{T}{1 - z}. \quad (20)$$

As expected, $W_0^*(z)$ has a nonunity denominator.

Recalling the definition $z = e^{-Ts}$ and letting $s = j\omega$ lead to

$$W_0^*(j\omega) = \frac{T}{1 - e^{-jT\omega}} = \frac{T}{2}(1 - j \cot T\omega/2), \quad (21)$$

from which the locus (Fig. 8(b)) is readily obtained and compared with the ideal one. In parts (c) and (d) of Fig. 8 the corresponding amplitude and phase comparisons are made. If the spectrum of the input function is known, the rms error in the integral can be determined.

A bit more sophistication than above yields the so-called trapezoidal formula, as illustrated in Fig. 9(a). The corresponding difference equation is

$$o^*(t) = (T/2)[i^*(t) + i^*(t - T)] + o^*(t - T), \quad (22)$$

and the transfer function is

$$W_1^*(z) = \frac{T}{2}\frac{1 + z}{1 - z}. \quad (23)$$

The effect of the trapezoidal rule at various frequencies is shown by

$$W_1^*(j\omega) = \frac{T}{2}\frac{1 + e^{-jT\omega}}{1 - e^{-jT\omega}} = \frac{T}{2}(-j \cot T\omega/2). \quad (24)$$

Comparison of (21) and (24) shows that the latter locus is shifted to the left by the amount $T/2$, which will therefore coincide with the locus of the ideal integration. Thus, the trapezoidal rule has ideal phase, and its shortcoming is in its amplitude characteristics, as seen in Fig. 9(b).

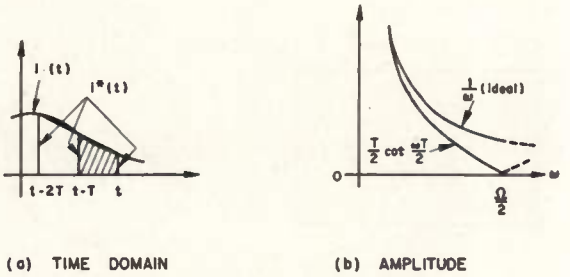


Fig. 9—Characteristics of the trapezoidal rule.

It is surprising that a quadrature formula derived by a crude approximation in the time domain has ideal phase. Other classical integration rules also have ideal phase characteristics in their useful ranges of frequency.[16] Since this is so, a complete comparison of the various formulas is afforded in their amplitude characteristics, as done in Fig. 10.

[15] N. Wiener, "Extrapolation, Interpolation and Smoothing of Stationary Time Series," Technology Press, Mass. Inst. Tech., Cambridge, Mass., & John Wiley and Sons, Inc., New York, N. Y., pp. 129–148; 1949.

[16] Cf. ref. 3, pp. 233–234 or extended version of this paper.

Fig. 10 confirms, at a glance, several known properties of these formulas: the superiority of Simpson's one-third rule over his more complex three-eighth rule; the high precision of Weddle's rule (up to a certain frequency). The figure also shows effects not generally emphasized in numerical calculus but often important in system design. For example, the trapezoidal rule is inferior to all the others at low frequencies, but only to Simpson's one-third rule at the high end of the range.

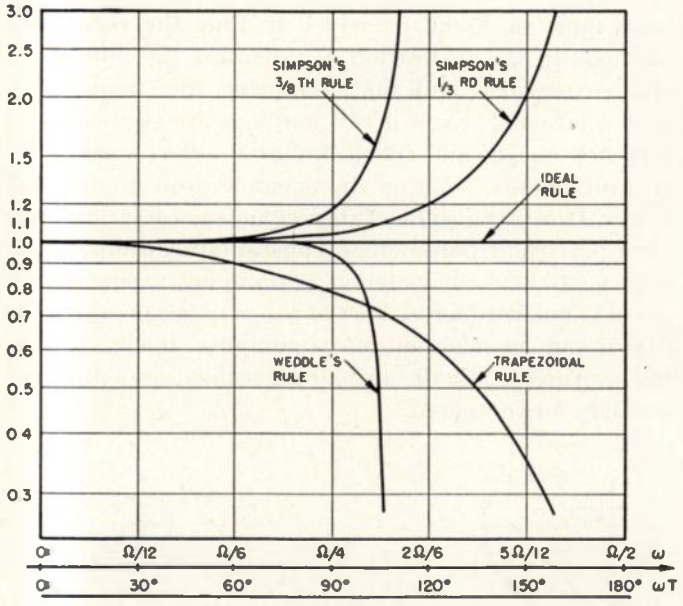


Fig. 10—Comparison of quadrature formulas by ratios of their amplitudes to that of ideal integration.

Moreover, one may actually prefer the trapezoidal rule to Simpson's rule, not only because it is simpler, but also because it attenuates at higher frequencies, thereby counteracting the effect of noise.

Let us make use of this error study in a specific example. Suppose a sine wave of ω_0 radians/sec is to be integrated by using six samples in each cycle. What would be the amplitude error in the result? At $\omega_0 = \Omega/6$, or $\omega_0 T = 60$ degrees, Fig. 10 shows that the errors are: trapezoidal -9.3 per cent; Simpson's one-third $+0.77$ per cent; Simpson's three-eighth $+1.12$ per cent; Weddle's -0.25 per cent.

For reference, the various relevant transfer functions are listed herewith.

Simpson's one-third rule:

$$W_2^*(z) = \frac{T}{3} \cdot \frac{1 + 4z + z^2}{1 - z^2}. \tag{25}$$

Simpson's three-eighths rule:

$$W_3^*(z) = \frac{3T}{8} \cdot \frac{1 + 3z + 3z^2 + z^3}{1 - z^3}. \tag{26}$$

Weddle's Rule:

$$W_3^*(z) = \frac{3T}{10} \cdot \frac{1 + 5z + z^2 + 6z^3 + z^4 + 5z^5 + z^6}{1 - z^6}$$

$$= \frac{3T}{10} \cdot \frac{1 + 6z + 6z^2 + 6z^3 + z^4}{1 + z - z^3 - z^4}. \tag{27}$$

X. Stability

One must bear in mind that a digital computer with its program acts as an active, rather than a passive filter. The question of stability or convergence has always been important in numerical analysis, but in servo applications it becomes crucial.

The stability of a program may be studied in terms of either the variable s or z. In terms of s, stability is predicated upon the absence of poles in the right-half s-plane and of multiple poles on the imaginary axis. The numerator of $W^*(s)$ cannot go to infinity for any finite value of s because it is an entire function. To prove this note that the magnitude of a typical term,

$$\left| a_k e^{-kTs} \right| = \left| a_k e^{-kT(\sigma + j\omega)} \right| = \left| a_k e^{-kT\sigma} \right|, \tag{28}$$

is finite for all finite σ. Consequently, only the zeros of the denominator need to be examined.

The denominator of $W^*(s)$ of (4) is

$$Q^*(s) = 1 + b_1 e^{-Ts} + \cdots + b_n e^{-nTs}. \tag{29}$$

If $Q^*(s)$ has a zero in the right half s-plane, it has one also in the cross-hatched strip of Fig. 6(a) because of periodicity. If such a zero exists, the Q^*-plane contour conformal to $ABCD$ of Fig. 6(a) encloses the origin. However, only the A-contour (the locus of $Q^*(s)$) needs to be mapped, because B and D will always coincide in the Q^*-plane (periodicity!) and C will be a vanishingly small circle around 1(not affecting the encirclement of the origin). Therefore, if the numerator and denominator of $W^*(s)$ have no common zeros, *a sufficient test of stability of a realizable program consists of determining if the denominator locus excludes the origin.*

Very often a rough sketch of the Q^*-locus is all that is required to determine stability. The amount by which the locus avoids the origin is an indication of the relative stability. The frequency of closest approach to the origin is often very nearly the resonant frequency of the program. Fig. 11 gives a few simple examples of Q^*-loci and the conclusions they lead to.

Stability can also be determined in terms of the variable z. Again only the zeros of the denominator, $Q^*(z)$, need to be examined. Since the z-plane region conformal to the right half s-plane is the inside of the unit circle, *for a stable program the magnitudes of the zeros of $Q^*(z)$ must be greater than unity, or equal to unity and simple.*

Using the variable z to determine stability calls for finding the roots of $Q^*(z) = 0$. Although bilinear transformation[17] and the subsequent application of Routh's or Hurwitz's criteria define absolute stability, they give no indication of relative stability.

[17] A. Madwed, "Number Series Method of Solving Linear and Non-Linear Differential Equations," Sc.D. thesis in Mechanical Engineering, Mass. Inst. Tech.; 1950.

The interpretation of stability in the time domain is helpful. If we let $W^*(z)=1/Q^*(z)$, then for parts (a), (b), and (c) of Fig. 11

$$W^*(z) = \frac{O^*(z)}{I^*(z)} = \frac{1}{1-bz}. \tag{30}$$

The corresponding difference equation is

$$o^*(t) = i^*(t) + bo^*(t-T). \tag{31}$$

If $i^*(t)$ is allowed to be a single small impulse, δ, at one sampling instant, $o^*(t)$ will also be δ. But the next output sample will be $b\delta$, the second one $b^2\delta$, and so on. It is clear that if $b>1$, the output becomes ever greater (unstable). Since the zero of $Q^*(z)$ is $1/b$, the stability criteria are verified.

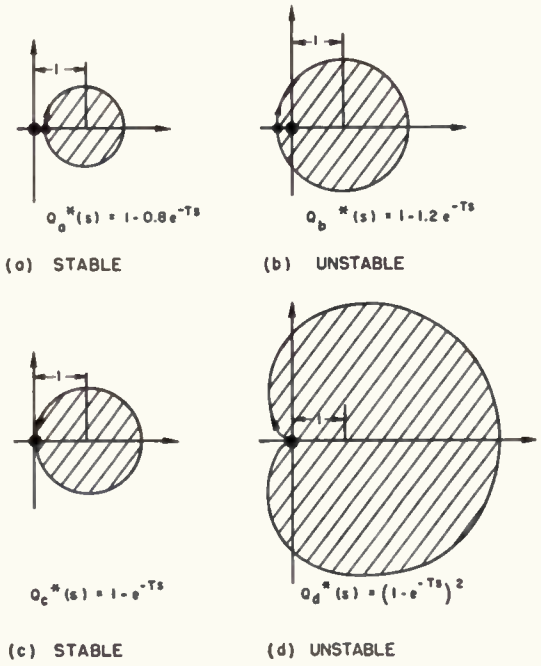


Fig. 11—Illustration of stability studies by conformal mapping.

XI. Differentiation

Numerical differentiation can be used as an illustration of the stability conditions. Since the transfer function of an ideal differentiator, s, is the reciprocal of that of an ideal integrator, it appears that the reciprocals of the integration formulas, (20), (23), (25), (26), and (27) ought to be suitable differentiation programs.

It turns out that only the first two of the five reciprocated programs are stable. They were derived on a different basis in a preceding article.[1] Both are stable, because the first has a constant denominator, while the second has a simple pole at $z=-1$.

The stability tests of the other three differentiations are illustrated in Fig. 12 both by conformal mapping and by the location of the poles in the z-plane. We discuss the first one only, which is illustrated in part (a) of Fig. 12. Actually, it is not necessary to sketch the whole locus, for its crossings of the real axis are easily calculated. At the frequencies $\omega=0$ and $\Omega/2$, $z=\pm 1$, and $1+4z+z^2$ (the denominator) is respectively $+6$ and -2 which are of opposite sign so that the origin is clearly enclosed. The z-plane plot shows that one of the two poles is, indeed, a violator of the stability condition.

XII. Realization of Programs (Programming)

In arriving at the realizability theorem in section IV, we have also illustrated a method of realizing a program, given its transfer function. Are there other methods of realizing a program? If so, what advantage might they offer? Just as in network realization or synthesis, some basis must be found on which to gage the value of a method. In the realization of programs the number of storage registers and the computing time required is such a measure. Even if two methods are equivalent in terms of storage and time, they may differ in other important details, such as the accumulation of roundoff errors. It might also be of particular advantage to associate particular parameters (constants) of the program with particular effects, such as damping, resonant frequency, poles, or zeros. Part of the realization is also the design and construction of the computer itself. In what follows three methods of program realization will be described and compared.

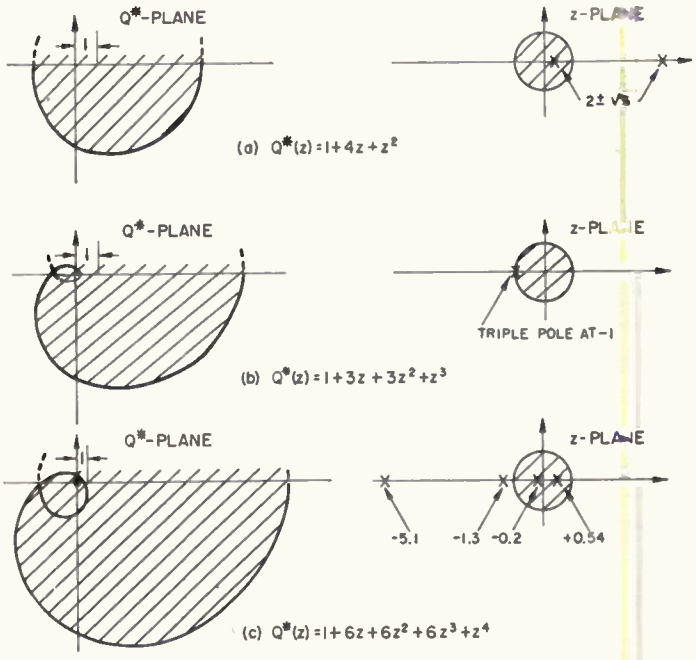


Fig. 12—Proof of instability of "precise" differentiation programs.

Direct Programming

Starting with the transfer function (4) we can retrogress through (3) to (2) and (1), which are two forms of the same difference equation. The program that solves the difference equation (2) has two distinct tasks: an arithmetic task and a manipulative task. The arithmetic task is the performance of the multiplications and additions indicated explicitly in (2). Since there are $m+n+1$ terms on the right side of (2), $m+n+1$ multiplications and $m+n$ additions are required.

This computation is repeated in every sampling interval but not with the same data; the second (manipu-

lative) task of the program is to bring the data up-to-date. The sample $i^*(t)$ of the present calculation becomes $i^*(t-T)$ in the next one, and therefore, the present $i^*(t)$ must be manipulated into the storage location associated with $i^*(t-T)$. Similar manipulations take account of the aging of other input and output data and are carried out by means of storage transfer operations.

The operations required to perform these tasks are within the capabilities of most conventional computers. A scoreboard showing the required storage is a helpful yardstick, for the number of instructions stored gives an indication also of the required computing time. For the programming method just discussed this scoreboard is shown below, if all constants in (4) are different and non-zero.

Storage for direct programming (nominal):

Constants		$m+\ n+1$
Data		$m+\ n$
Multiplications	$m+n+1$	
Additions	$m+n$	
Transfers	$m+n$	
Total Operations		$3m+3n+1$
		Total $5m+5n+2$

It is emphasized that this score is entirely nominal and does not necessarily relate to a particular computer, as the logic and instruction code radically influence the number of instructions.[18]

Cascade Programming

Taking a clue from network synthesis, we should explore other means of realizing programs on the basis of the transfer function (4) or (12). We assume that $m \leqq n-1$ and that both the numerator and the denominator have only real and simple zeros. These restrictions do not encroach on the generality of our arguments, and they allow us to write (12) as

$$W^*(z) = \frac{a_0}{1+d_1z}\cdot\frac{1+c_2z}{1+d_2z}\cdot\ \cdots\ \cdot\frac{1+c_nz}{1+d_nz}, \qquad (32)$$

where $-1/c_k$ and $-1/d_k$ are zeros of the numerator and denominator, respectively. (Some of the c's may be zero.)

Each factor in (32) considered by itself is the transfer function of a realizable program, so that the original program (Fig. 13(a)) is replaced by a cascade of simpler ones (Fig. 13(b)), the output of one block being the input to the next one. Thus,

$$\frac{O^*(z)}{I^*(z)} = \frac{O_1^*(z)}{I^*(z)}\cdot\frac{O_2^*(z)}{O_1^*(z)}\cdot\ \cdots\ \cdot\frac{O^*(z)}{O_{n-1}^*(z)}. \qquad (33)$$

If we now identify corresponding factors of (32) and (33), perform cross multiplication and inverse transformation, we obtain the following set of difference equations

$$\begin{aligned}
o_1^*(t) &= a_0 i^*(t) && - d_1 o_1^*(t-T)\\
o_2^*(t) &= o_1^*(t) + c_2 o_1^*(t-T) && - d_2 o_2^*(t-T)\\
&\ \ \vdots\\
o^*(t) &= o_{n-1}^*(t) + c_n o_{n-1}^*(t-T) - d_n o^*(t-T).
\end{aligned} \qquad (34)$$

The mere counting of appropriate items in (34) gives the score below.

Storage for cascade programming (nominal):

Constants		$m+\ n+1$
Data		n
Multiplications	$m+n+1$	
Additions	$m+n$	
Transfers	n	
Total operations		$2m+3n+1$
		Total $3m+5n+2$

This nominal figure of merit definitely favors cacade programming over the direct method. Actually, the cascade method is often used in numerical analysis but not called so. For example, in double integration a single formula can be used to find the double integral *directly*, or a single-quadrature formula may be applied twice in a row amounting to two *cascaded* operations.

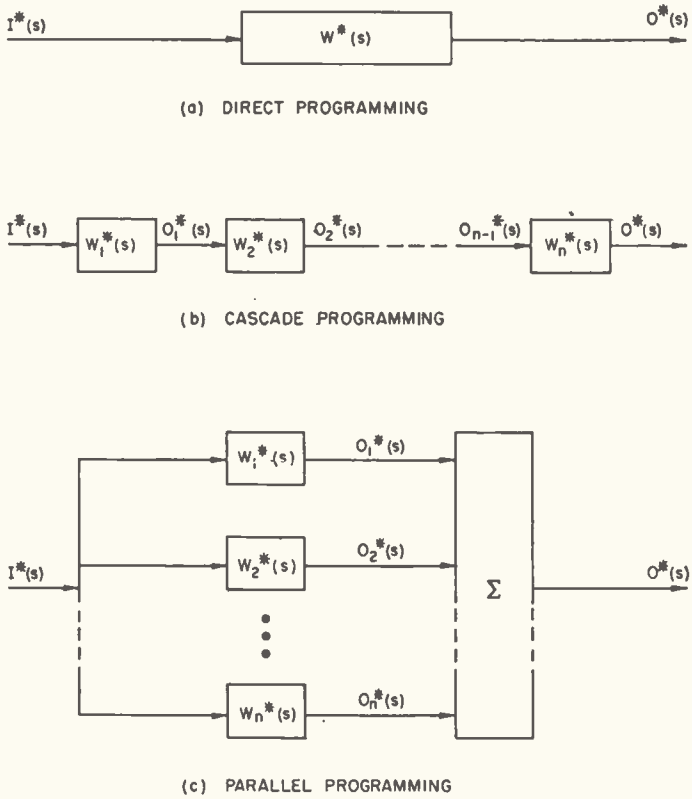


Fig. 13—Comparison of methods of program realization.

Parallel Programming

The obvious thing to examine next is the partial-fraction expansion of $W^*(z)$. With the same restrictions as above, $W^*(z)$ is now written as

$$W^*(z) = \frac{f_1}{1+d_1z} + \frac{f_2}{1+d_2z} + \cdots + \frac{f_n}{1+d_nz}, \qquad (35)$$

[18] Details of this program in the Whirlwind code were worked out in ref 3, pp. 273–292.

where f_k are the residues of $W^*(z)$ at the poles $z_k = -1/d_k$. As illustrated in Fig. 13(c), the sum of the transfer functions is realized by applying the same input to each block and adding their outputs. Mathematically this corresponds to[19]

$$\frac{O^*(z)}{I^*(z)} = \frac{O_1^*(z)}{I^*(z)} + \frac{O_2^*(z)}{I^*(z)} + \cdots + \frac{O_n^*(z)}{I^*(z)}. \quad (36)$$

Identification of corresponding terms of (35) and (36), cross multiplication and inverse transformation lead to the set

$$\begin{aligned} o_1^*(t) &= f_1 i^*(t) - d_1 o_1^*(t - T), \\ o_2^*(t) &= f_2 i^*(t) - d_2 o_2^*(t - T), \\ &\vdots \\ o_n^*(t) &= f_n i^*(t) - d_n o_n^*(t - T), \\ o^*(t) &= o_1^*(t) + o_2^*(t) + \cdots + o_n^*(t), \end{aligned} \quad (37)$$

which score as shown below.

Storage for parallel programming (nominal):

Constants		$2n$
Data		n
Multiplications	$2n$	
Additions	$2n-1$	
Transfers	n	
Total Operations		$5n-1$
	Total	$8n-1$

Parallel programming seems to occupy a position in between the other two methods discussed as far as efficiency is concerned. It is not a method generally used in numerical work. Indeed, it is not at all obvious how the correspondence between (2) and (37), just established with such natural ease, could be obtained by more conventional methods in the time domain.

Comparison of Programming Methods

Although the number of storage registers and the computing time required can serve as a figure of merit of the programming methods, much depends on the particular computer and instruction code. Indeed, studies of programming methods do, or should influence, the designs of both the computer and the code.

Inclusion into the code of the Whirlwind computer of a so-called "exchange" instruction (which exchanges the contents of the arithmetic unit and the selected storage register) made the direct programming method more efficient than the cascade method. The direct programming may also be preferred when several of the constants in $W^*(z)$ are zero, because factorization (for cascade programming) "recreates" all the constants, thereby increasing the number of multiplications. Moreover, direct programming can minimize the delay due to computation because all terms of (2), except $a_0 i^*(t)$, can be calculated before the input $i^*(t)$ is sampled. In a control application such a consideration might be important.

Beside its good score, cascade programming has the advantage of producing intermediate results, which may also be useful. A double integration (used "directly") produces a position from a measured acceleration, but two simple integrations ("cascaded") make the velocity also available.

Outside of its academic interest, parallel programming could affect the basic design of a computer. Since each equation of the set (37) has the same input $i^*(t)$, they can be solved simultaneously if n suitable arithmetic units are built. This seems to be an extravagance of equipment; nevertheless, the speed advantage is nearly n-fold. In particular cases a mixture of these programming methods may be the solution.

When a program is to be designed or improved by experimental methods, the cascade or parallel schemes may offer particular advantages. If the design is based on the zero-pole configuration, the cascade method would be used, for each constant of (34) is directly related to a particular design parameter. Similarly, in parallel programming each constant has to do with either a pole or a residue. For a desired effect, a particular constant needs to be changed.

XIII. Conclusion

The discussion was aimed at establishing a method applicable to the analysis of digital computers used in control systems. The emphasis was on the approach to this problem rather than the use of specific techniques. The methods can also be applied to usual problems of numerical analysis, such as the convergence and resonance of numerical solutions. Although the treatment presented is not complete, a groundwork has been laid which permits ready extension.

One difficulty is that the field of digital controls is young, and the problems and needs are not yet clearly defined. For example, the general problem of approximating desired characteristics by a realizable program transfer function is a difficult one, but specific cases arise only with specific application. New realization techniques are also bound to be created by new developments in computer components, circuitry, and logic.

One could question the usefulness of linear analysis of a digital computer, which is capable of a variety of nonlinear operations. However, even in nonlinear systems, linear analysis can be applied and used as an aid. The extension of frequency analysis to the digital computer permits, rather than commits, one to extend existing methods of servo analysis to mixed systems; and this might be useful in many cases.

XIV. Acknowledgment

This investigation was done under the auspices of the Digital Computer Laboratory at M.I.T. The author is indebted to Professor W. K. Linvill for his guidance and active help. The aid and advice of Professors Z. Kopal and M. F. Gardner are also gratefully acknowledged.

[19] The outputs $O_1^*(z), \cdots, O_n^*(z)$ are not the same as the ones so named in (33).

Unbalanced RLC Networks Containing Only One Resistance and One Real Transformer*

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Summary—**Any transfer function, in order to be physically realizable as a lossless coupling network terminated in a single resistance, must have an even or odd polynomial as its numerator. Such a function, representing a transfer admittance, a transfer voltage ratio, or a transfer impedance, can be realized by the method of this paper as a lossless lattice network terminated in resistance and possessing no mutual inductance. If the numerator of the transfer admittance or voltage ratio is even and the numerator of the transfer impedance is odd, these lattices can *always* be reduced to unbalanced networks. No ideal transformers are needed in the final network; only one *real* transformer is required, where a real transformer is defined as one that has a coupling coefficient smaller than one and finite magnetizing inductance.**

Introduction

IN A PREVIOUS paper[1] a solution was given to the problem of realizing a given transfer admittance by a Darlington network, such a network being defined as one possessing lossless elements plus only one resistance. We demonstrate in this paper an alternate solution to the Darlington problem by realizing a transfer function as an unbalanced lossless network terminated in a resistance. This procedure is more general in that the transfer function is not restricted to the dimensions of an admittance but may also be a transfer voltage ratio or a transfer impedance. The final network is achieved through the intermediate step of designing a lattice whose arms are canonic Foster two-element kind networks containing the same poles. No mutual inductance is needed for realizing the lattice. In many cases the lattice can be reduced to an unbalanced network without the addition of mutual inductance; it can *always* be reduced by the use of *one* real transformer, that is, a transformer with finite magnetizing inductance and a coupling coefficient smaller than one.

It is well known that it is impossible to realize a minimum phase transfer function with zeros off the j axis by means of a lossless coupling network terminated in a single resistance.[2] Therefore for realization as a lattice by the method of this paper it is necessary that the given function have an even or odd numerator. However, to guarantee reduction to an unbalanced form, the numerator must be an even polynomial for a transfer admittance or transfer voltage ratio and an odd polynomial for a transfer impedance. Attention is focused on realization of the unbalanced network since it includes the steps for lattice realization. Furthermore, only the procedures for the transfer admittance and the transfer impedance need be demonstrated; it is clear that because of the resistance termination the problem of achieving a transfer voltage ratio is almost identical with that of the transfer admittance. Some of the general formulas upon which we base our discussion are given in a previous paper[3] on another method of RLC synthesis; they are also given in the Bower-Ordung method of RC synthesis.[4] In both cases they are applied to different problems. A detailed discussion is therefore not required, but for purposes of clarity and definition of symbols we shall restate the necessary development and formulas in this paper.

The method for lattice synthesis will be given first and then the general reduction procedure will be demonstrated.

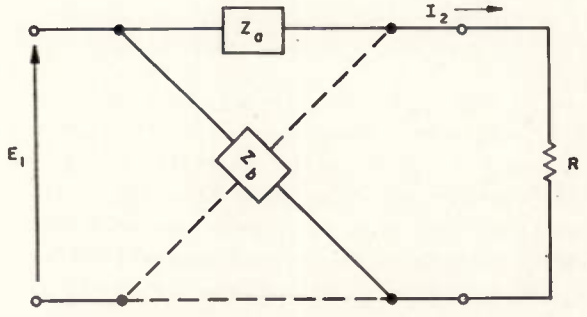

Fig. 1—Lossless lattice terminated in resistance.

I. The Synthesis Procedure

A. Realization of a Transfer Admittance

We are given a quotient of polynomials $p(s)/q(s)$, where p is even and of degree equal to or less than that of q, and q is completely general, i.e., is restricted only in that it must be a Hurwitz polynomial. We wish to realize the given function within a multiplicative constant as the transfer admittance Y_{12} of the resistance-terminated lattice shown in Fig. 1. First we realize the

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† Hughes Research and Development Laboratories, Culver City, Calif.

[1] L. Weinberg, "The Darlington problem," *Jour. Appl. Phys.*, vol. 24, pp. 776–779; June, 1953.

[2] E. A. Guillemin, "A Summary of Modern Methods of Network Synthesis," Advances in Electronics, vol. III, Academic Press, Inc., New York, N.Y.; 1951.

[3] L. Weinberg, "A General RLC Synthesis Procedure," *IRE National Convention*, 1953, part 5.

[4] J. L. Bower and P. F. Ordung, "The synthesis of resistor-capacitor networks," PROC. I.R.E., vol. 38, pp. 263–269; March, 1950.

short-circuited lattice shown in Fig. 2; then by lattice transformations[3] we obtain the resistance-terminated lattice characterized by $Y_{12} \equiv I_2/E_1$.

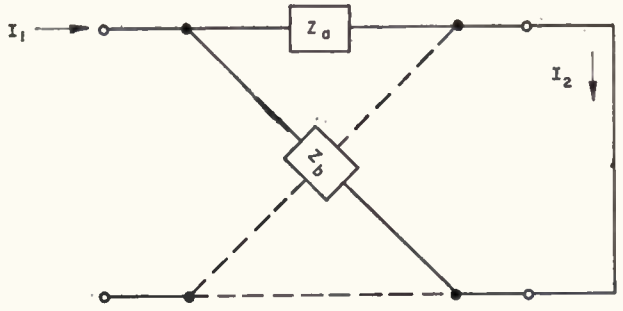

Fig. 2—Short-circuited lattice for which $K = (I_2/I_1) = [(Y_a - Y_b)/(Y_a + Y_b)]$.

We let the ratio of the currents in Fig. 2 be expressed

$$K \equiv \frac{I_2}{I_1} = \frac{p}{Hq} = \frac{m_1}{H(m_2 + n_2)}, \tag{1}$$

where H is a positive constant, m_1 is the even numerator, and m_2 and n_2 are respectively the even and odd parts of the denominator. Since for a short-circuited lattice we know that

$$K = \frac{I_2}{I_1} = \frac{y_{21}}{y_{11}} = \frac{Y_a - Y_b}{Y_a + Y_b} \tag{2}$$

where y_{21} and y_{11} are the familiar short-circuit transfer and driving-point admittances respectively, and $Y_a = 1/Z_a$ and $Y_b = 1/Z_b$ are the respective admittances of the lattice series and cross arms. Since the short-circuit transfer admittance y_{21} of a lossless network cannot be given by the ratio of two even or two odd polynomials, we divide numerator and denominator of (1) by n_2 to obtain

$$K = \frac{m_1/n_2}{H(1 + m_2/n_2)}, \tag{3}$$

which, when equated with (2), yields

$$K = \frac{m_1/n_2}{H(1 + m_2/n_2)} = \frac{Y_a - Y_b}{Y_a + Y_b}. \tag{4}$$

It is now necessary to show how the numerator and denominator that have been derived from the given rational function may be identified with the lattice arms so that (4) is satisfied.

We consider Y_a and Y_b and also m_1/n_2 and m_2/n_2 as expanded in partial fractions. The residues of m_1/n_2, which we will designate by $k_\mu^{(n)}$, are all real but may be positive or negative; those of m_2/n_2, moreover, are real and positive, since the ratio of the even and odd parts of a Hurwitz polynomial forms a positive real lossless function.[5] Thus the partial fraction expansion of the denominator may be written as

$$\begin{aligned} H(1 + m_2/n_2) &= H\left(k_0^{(d)} + k_1^{(d)}s + \frac{k_2^{(d)}}{s} + \frac{k_3^{(d)}}{s - s_3} + \frac{k_4^{(d)}}{s - s_4} + \frac{k_5^{(d)}}{s - s_5} + \frac{k_6^{(d)}}{s - s_6} + \cdots\right) \\ &= H\left(k_0^{(d)} + k_1^{(d)}s + \frac{k_2^{(d)}}{s} + \frac{k_3^{(d)}}{s - s_3} + \frac{k_3^{(d)}}{s - \bar{s}_3} + \frac{k_5^{(d)}}{s - s_5} + \frac{k_5^{(d)}}{s - \bar{s}_5} + \cdots\right), \end{aligned} \tag{5}$$

where $k_0^{(d)} = 1$ and all the other k's are real and non-negative. It is clear that $k_1^{(d)} = 0$, if the degree of q, the denominator of the given quotient of polynomials, is odd. It is pointed out, furthermore, because of its importance in the reduction of the lattice to an unbalanced network, that $k_2^{(d)}$ is always greater than zero, that is, the pole at the origin is always present.

If we now equate the residues of like terms for the numerator and denominator of (4), we obtain

$$\left.\begin{aligned} k_\mu^{(a)} - k_\mu^{(b)} &= k_\mu^{(n)} \\ k_\mu^{(a)} + k_\mu^{(b)} &= Hk_\mu^{(d)} \end{aligned}\right\}. \quad (\mu = 0, 1, 2, \cdots, r, \text{ where } r \text{ is the degree of } q \text{ for } q \text{ of even degree, and } r \text{ equals one plus the degree of } q \text{ for } q \text{ of odd degree}) \tag{6}$$

This set of two simultaneous equations may be solved simply for the unknown residues of the lattice arms, $k_\mu^{(a)}$ and $k_\mu^{(b)}$. The solution is obtained at sight as

$$\left.\begin{aligned} k_\mu^{(a)} &= \tfrac{1}{2}(Hk_\mu^{(d)} + k_\mu^{(n)}) \\ k_\mu^{(b)} &= \tfrac{1}{2}(Hk_\mu^{(d)} - k_\mu^{(n)}) \end{aligned}\right\}. \tag{7}$$

For $\mu = 0$ we see from the form of m_1/n_2 that $k_0^{(n)} = 0$ so that it is always true that $k_0^{(a)} = k_0^{(b)} = H/2$. In order for $(Y_a - H/2)$ and $(Y_b - H/2)$ to be positive real susceptance functions, it is necessary and sufficient that their respective residues $k_\mu^{(a)}$ and $k_\mu^{(b)}$ for $\mu \neq 0$ be non-negative. Applying this condition to each of the solutions in (7) and also making use of the fact that $k_\mu^{(d)}$ is never negative, we obtain as the condition to be satisfied,

$$-1 \leqq \frac{k_\mu^{(n)}}{Hk_\mu^{(d)}} \leqq 1. \tag{8}$$

The positive constant H is within our control; by choosing it sufficiently large, the above condition can be met for all values of μ. The lattice arms are then given by an expansion of the form

$$Y = H/2 + k_1 s + k_2/s + k_3/(s - s_3) + k_3/(s - \bar{s}_3) + \cdots \tag{9}$$

[5] E. A. Guillemin, "The Mathematics of Circuit Analysis," John Wiley and Sons, Inc., New York, N.Y.; 1949.

and may be realized by inspection as a network of the form shown in Fig. 3. In Fig. 4 is shown the type of short-circuited lattice achieved, where Y_a' and Y_b' are lossless functions.

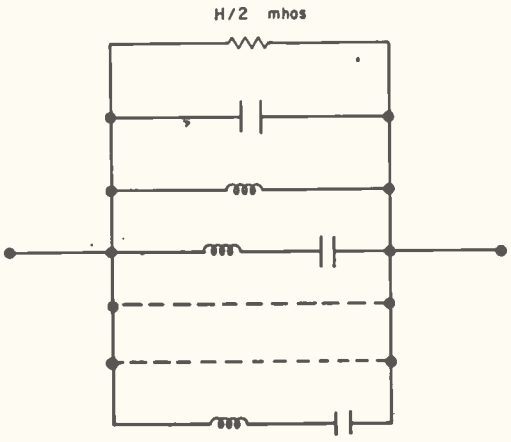

Fig. 3—General form of each of the lattice arms in the short-circuited lattice.

To convert to a resistance-terminated lattice, we first derive an equivalent lattice by removing the total conductance from each arm and inserting it as a shunt branch in parallel with each of the input and output terminal pairs; then we replace the current source and

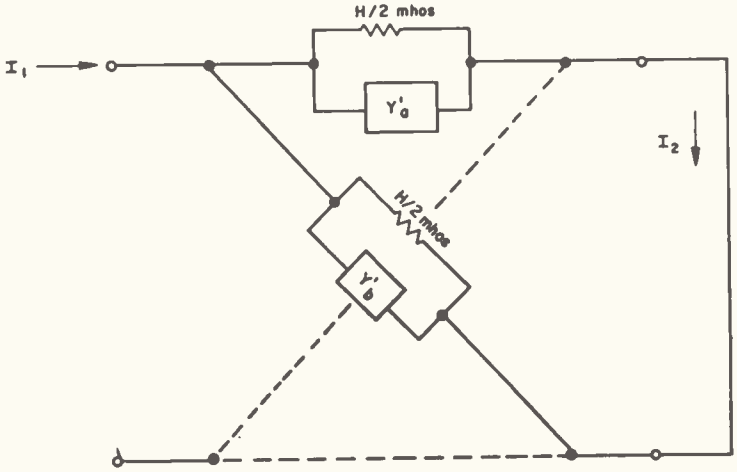

Fig. 4—Short-circuited lattice for which $(I_2/I_1) = (p/Hq)$ and Y_a' and Y_b' are lossless functions.

its shunt conductance by an equivalent voltage source plus a series resistance; finally the application of the reciprocity theorem gives a lattice terminated in a conductance of $H/2$ mhos for which[5,6]

$$Y_{12} = \frac{I_2}{E_1} = (1/2)p/q. \tag{10}$$

The sequence of steps is shown in Fig. 5, next page.

The lattice synthesis procedure is now complete and may be summarized as follows:

1. Divide the numerator and denominator of the given transfer admittance by the odd part of the denominator. (We are assuming an even numerator; of course, for p an odd polynomial we divide by the even part of the denominator.)
2. Expand m_1/n_2 and m_2/n_2 in partial fractions.
3. For each pole satisfy the appropriate part of (8) with the equals sign; that is, for $k_\mu^{(n)}$ negative choose a value of H that satisfies $k_\mu^{(n)}/Hk_\mu^{(d)} = -1$; and for $k_\mu^{(n)}$ positive we satisfy the condition on the $+1$ with the equals sign. Tabulate all the H values. For the conjugate poles on the j axis the residues are equal, i.e., $k_\mu = k_{\mu+1}$ for μ odd and $\mu \geqq 3$.
4. Choose a value of H greater than the largest value of H obtained in step 3. This guarantees the satisfaction of (8) with the inequality signs. (In some cases if we wish to save some elements we can let H equal the largest value; this makes the residue zero in one pole for one of the arms.)
5. Using (7) find all the residues of the lattice arms, and realize the arms in the Foster manner.
6. Convert the short-circuited lattice into one terminated in a conductance of $H/2$ mhos and possessing pure reactance arms.

B. Realization of Transfer Impedance

The procedure is analogous to the one for the transfer admittance with a few minor changes that are pointed out below.

In order for reduction of the lattice to be possible in all cases the numerator must be an odd polynomial. We first realize the transfer voltage ratio[7]

$$K' \equiv E_2/E_1 = z_{21}/z_{11} = \frac{Z_b - Z_a}{Z_b + Z_a}, \tag{11}$$

for the open-circuited lattice and then by lattice conversions obtain a lattice terminated in a resistance of $H/2$ ohms for which

$$Z_{12} \equiv E_2/I_1 = (1/2)p/q. \tag{12}$$

In addition, however, we now realize the lattice arms by a partial fraction expansion of the lossless *admittances;* that is, we have the extra step of combining the partial fraction expansions of $(Z_a - H/2)$ and of $(Z_b - H/2)$, inverting to the admittance forms, and then expanding these admittances in partial fractions. It is pointed out that the admittances will always have a pole at the origin; the resulting shunt inductance will be used in the reduction to an unbalanced network. Except for this pole at the origin and, in those cases pointed out below, a pole at infinity, like poles will not be present in the admittances of the lattice series and cross arms.

[6] L. Weinberg, "New Synthesis Procedures for Realizing Transfer Functions of RLC and RC Networks," Technical Report No. 201, R.L.E., Mass. Inst. Tech.; September, 1951.

[7] Because of the assumed positive directions of voltages and currents, the subscript b is interchanged with a in the transfer impedance case in (11). This therefore also interchanges these letters in (6) and (7).

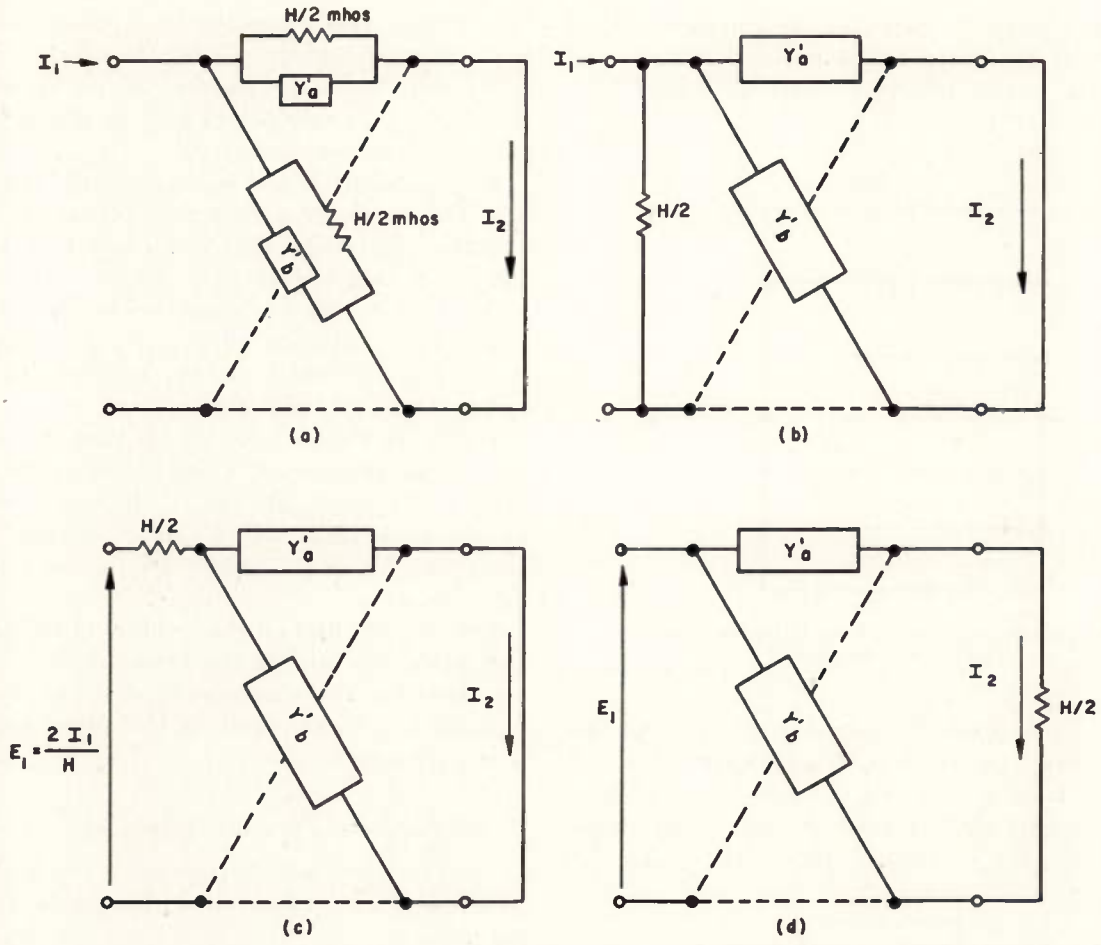

Fig. 5—Steps in the conversion of short-circuited lattice for which $K = (I_2/I_1) = (p/Hq)$ to a resistance-terminated lattice with $Y_{12} = (I_2/E_1) = (1/2)p/q$. (a) Short-circuited lattice for which $K = (p/Hq)$. (b) Lattice equivalent to that in (a). (c) Lattice after application of Thevenin's theorem. (d) Lattice given by application of reciprocity theorem, where $Y_{12} = (I_2/E_1) = (1/2)p/q$.

II. Reduction of the Lattices to Unbalanced Forms

To demonstrate that reduction of the lattice to an unbalanced form is always possible, we will consider the transfer admittance and the transfer impedance separately.

A. Transfer Admittance

Because of its usefulness in some problems, we first discuss a procedure that requires the use of more than one real transformer. Then we consider the more significant method that uses only one real transformer.

The general form of lattice obtained has arms composed of parallel branches of series LC structure, plus one pure inductance branch and one pure capacitance branch. As pointed out previously the arms always contain a shunt inductance branch because their admittances have been designed to have a pole at the origin. It has been noted also that the pure capacitance branch is present when the degree of q is even. As will be shown subsequently, the total capacitance in the cross arm can be removed by a lattice transformation; this capacitance is often useful in the practical instrumentation of the circuit, since it may provide for stray input or output capacitances or tube capacitances.

For convenience but without loss of generality (since the capacitance in the cross arm may always be removed and the remainder in the series arm becomes simply a bridging capacitance), we use as an illustration a lattice with three branches in each of its arms. The lattice is shown in Fig. 6, where like numerical subscripts de-

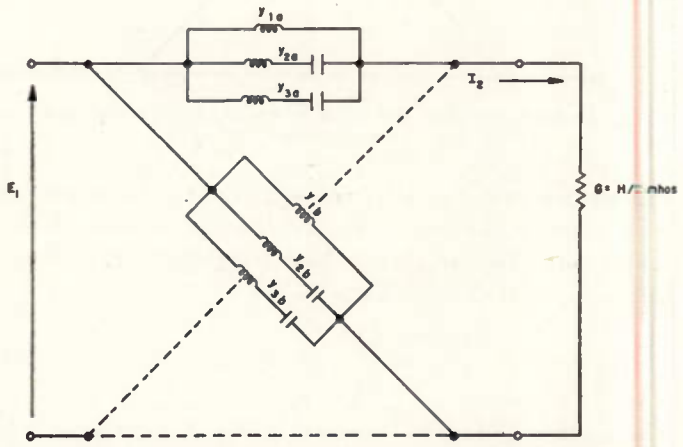

Fig. 6—Lossless lattice network terminated in a resistance.

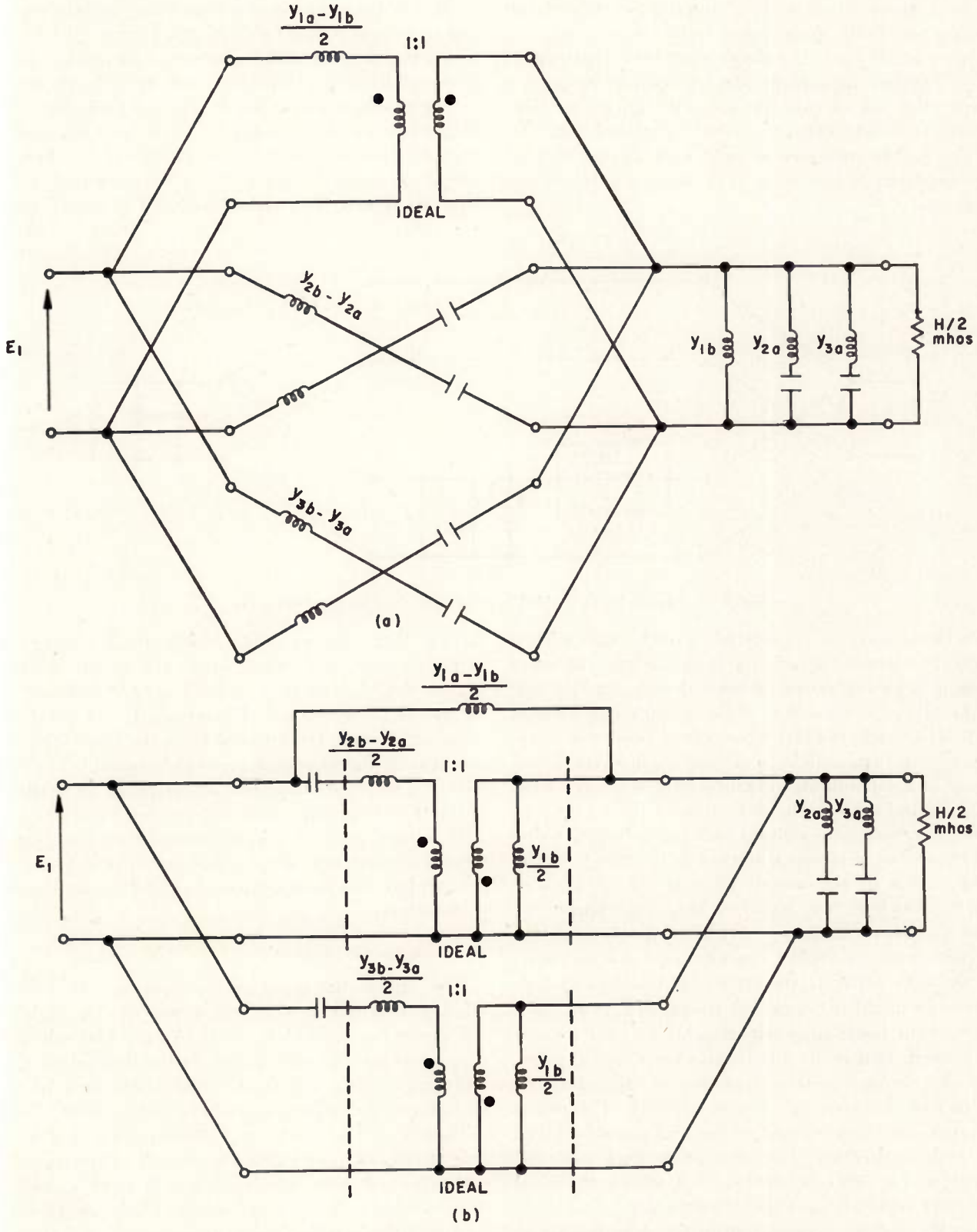

Fig. 7—Reduction of lattice to an unbalanced form with use of only real transformers.

nate like poles. Assume that the residues in y_{2b} and y_{3b} are greater than those in y_{2a} and y_{3a}, respectively, but the residue in y_{1a} is greater than that in y_{1b}. By the methods of transforming to equivalent lattices [3,6] we may remove y_{1b}, y_{2a}, and y_{3a} from each arm, and break the ladder into a group of parallel ladders, as shown in Fig. 7(a). Then by a rotation of output terminals of middle and bottom lattices, Fig. 7(b), we obtain an unbalanced form where each of the two sets of elements within the broken lines represents a real transformer,[8]

[8] E. E. Department, Mass. Inst. Tech., "Magnetic Circuits and Transformers," John Wiley and Sons, Inc., New York, N. Y.; 1943.

that is, a transformer with a coupling coefficient less than one and finite magnetizing inductance.

Inspection of Fig. 7(b) shows that two transformers will be required; for a higher degree transfer function it is clear that use of this method will require a larger number of transformers. It is most often desirable that only one real transformer be used, and we now demonstrate an alternate procedure that requires only one real transformer.

ance. The total shunt capacitance may be removed from the cross arm if $C_a \geqq C_b$, or in other terms, if $k_1^{(a)} \geqq k_1^{(b)}$, where k_1 represents the coefficient of the pole at infinity. This condition is always satisfied, as is shown below.

The numerator may be of lower degree than the denominator or of equal degree. If it is of lower degree, then $k_1^{(n)} = 0$ while $k_1^{(d)} > 0$ so that use of (7) shows that the shunt capacitances in the arms are equal and thus may be totally removed and placed in shunt with the

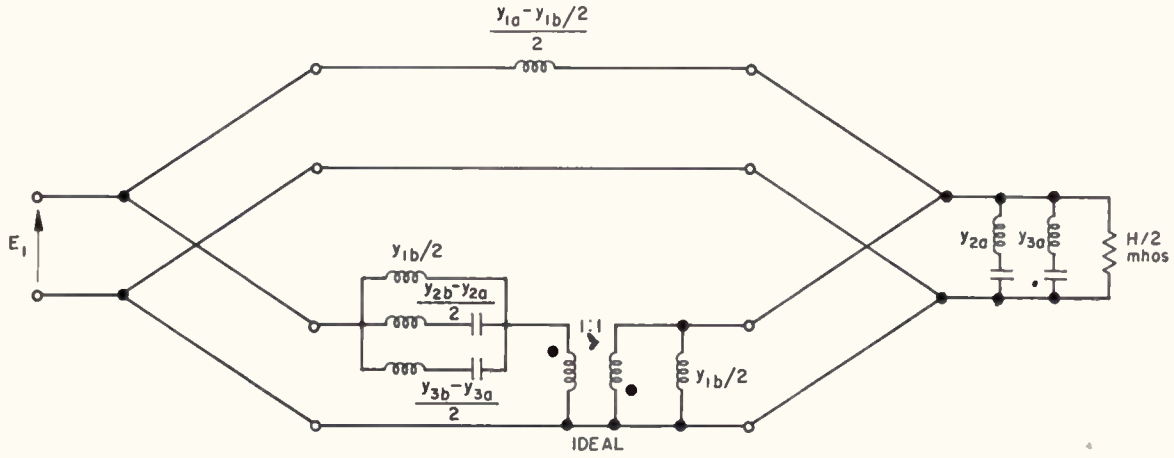

Fig. 8—Lattice of Fig. 6 divided into only two four-terminal networks.

Two cases need be considered, namely, one with a denominator whose highest degree is odd and the other possessing a denominator of even degree. In the first case the lattice arms will contain no pure capacitance branch, whereas in the latter the pole at infinity of m_2/n_2 places a pure capacitance branch in each of the arms. The case of a denominator of odd degree is treated first.

As an initial step in the reduction of the lattice, we remove from each arm a shunt inductance whose value is less than or at most equal to that in the arm with the smaller residue at the origin. Thus in the network of Fig. 6, for which $y_{1a} > y_{1b}$ has been assumed, suppose we remove $y_{1b}/2$ from each arm. We may also remove the other poles with the smaller residues, as was done previously. Now we separate the lattice into only *two* paralleled four-terminal networks, as shown in Fig. 8. Since the arm of the lower network contains no pure capacitance branch, that is, its admittance has a zero at infinity, we can obtain a *series* inductance by expanding the *impedance* of this arm as a Foster network. This series inductance can then be used as the leakage inductance of the real transformer. The form of network obtained is shown in Fig. 9(a), while Fig. 9(b) shows the ideal transformer replaced by a real transformer.

When the highest degree term in the denominator is even, a pole at infinity appears in the admittance of the lattice arms. This pole (which corresponds to the pure capacitance branch) must be completely removed from the cross arm before the rotation of the output terminals of the lattice; for if it is not removed, it is clear that the corresponding impedance of the arm will have a zero at infinity and thus not yield the required series inductlattice. If the degree of the numerator is equal to that of the denominator, then, since the coefficient of the highest degree term of m_1 is positive, $k_1^{(n)} > 0$, and again by use of (7) $k_1^{(a)} > k_1^{(b)}$. Consequently the total shunt capacitance may be removed from the cross arm. After removal of the capacitance it is then possible to proceed with the rotation of the output terminals as in the case without a shunt capacitance in the lattice arms.

It is thus clear that in the transfer admittance case we can always obtain an unbalanced form of network that employs at most one real transformer and no ideal transformers.

B. *Transfer Impedance*

The numerator is an odd polynomial, that is, has missing alternate powers and a zero at the origin, but the denominator, which has both even and odd parts, may be of even or odd degree. If the denominator is of odd degree, then a pole at infinity is present in n_2/m_2 and also in the impedances of the lattice arms. The admittances of the arms, as a consequence, will have a pole at the origin and a zero at infinity. Thus the Foster realization of these admittances will yield a shunt inductance but no shunt capacitance. There are, therefore, no difficulties in the subsequent procedure: it is the same as that for the case of the transfer admittance with no shunt capacitance in the lattice arms.

However, when the degree of the denominator is even, the shunt capacitance again appears in the arms. It must again be demonstrated that the total capacitance can be removed from the cross arm. The shunt capacitance, it should be clear, is present since the impedances

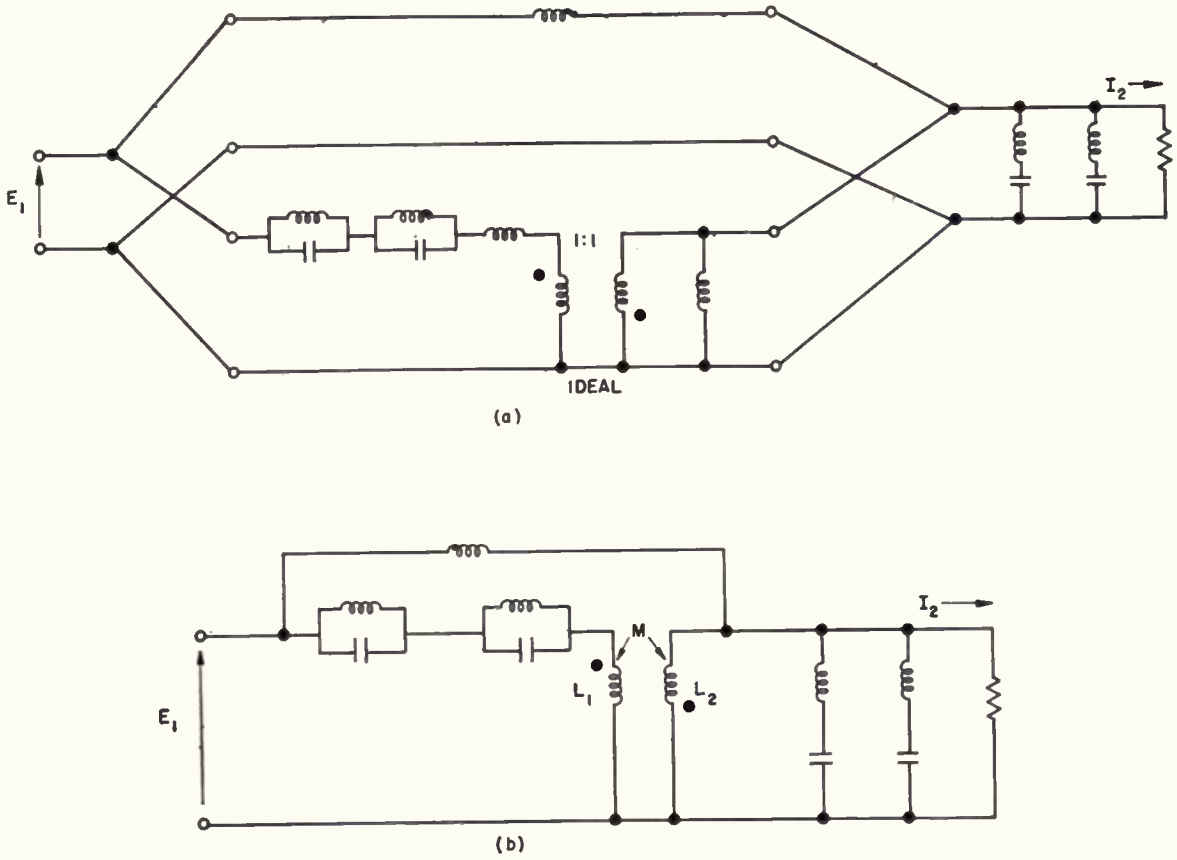

Fig. 9—Steps in obtaining unbalanced network with the use of only one real transformer. (a) Network before real transformer has been substituted. (b) Unbalanced network containing only one real transformer, i.e., a transformer with L_1 and L_2 finite and also $M/\sqrt{L_1L_2}<1$.

of the arms have no pole at infinity and therefore must have a zero there, so that the corresponding admittances have poles at infinity.

The impedances of the arms, possessing no pole at the origin or at infinity, will be of the form

$$(Z_a - H/2) \equiv Z_a' = \frac{2k_3^{(a)}s}{s^2+s_3^2} + \frac{2k_5^{(a)}s}{s^2+s_5^2} + \cdots + \frac{2k_r^{(a)}s}{s^2+s_r^2} = \sum_{\mu}^{r} \frac{2k_\mu^{(a)}s}{s^2+s_\mu^2} \quad (\mu = 3, 5, \cdots, r) \tag{13}$$

with a corresponding expression for Z_b'. The constants s_μ^2 are of course real and positive. If the terms of (13) are combined into one fraction, the coefficient of the highest degree term in the numerator will be $2\sum_\mu^r k_\mu^{(a)}$. Since the reciprocal of this coefficient is the value of the shunt capacitance of $Y_a'=1/Z_a'$, then in order for this value of capacitance not to be less than the capacitance of Y_b', it is only necessary that

$$2\sum_{\mu}^{r} k_\mu^{(b)} \geqq 2\sum_{\mu}^{r} k_\mu^{(a)}. \tag{14}$$

Substitution of the values given by (7) (with the b and a interchanged since this is the transfer impedance case) and some algebraic manipulation yield

$$\sum_{\mu}^{r} (Hk_\mu^{(d)} + k_\mu^{(n)}) \geqq \sum_{\mu}^{r} (Hk_\mu^{(d)} - k_\mu^{(n)})$$

$$\sum_{\mu}^{r} k_\mu^{(n)} \geqq -\sum_{\mu}^{r} k_\mu^{(n)} \tag{15}$$

$$\sum_{\mu}^{r} k_\mu^{(n)} \geqq 0.$$

The above condition on the residues is always satisfied. This may be seen by recalling that the sum of all the residues of a rational function is equal to the negative of the residue at infinity;[9] the residue at infinity of n_1/m_2 is given by

$$\text{Res. at infinity} = \lim_{s\to\infty}\left(\frac{-sn_1}{m_2}\right), \tag{16}$$

which is always negative or zero since the coefficient of the highest degree term of n_1 and m_2 is positive.

The total capacitance of the cross arm may therefore be removed, and the subsequent procedure will again yield an unbalanced network containing no ideal transformers and only one real transformer.

[9] T. M. Macrobert, "Functions of a Complex Variable," Macmillan and Co., Ltd., London, Eng.; 1947.

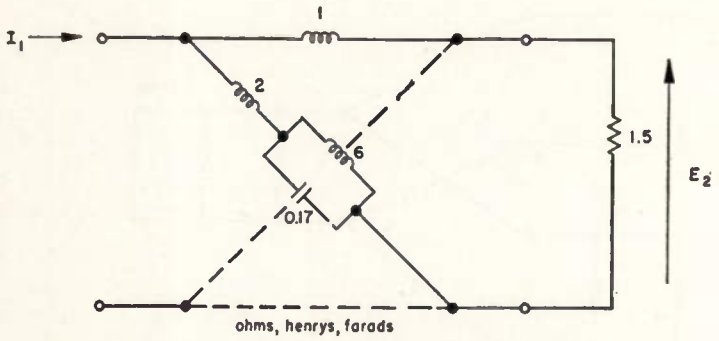

Fig. 10—Lattice achieved for illustrative example.

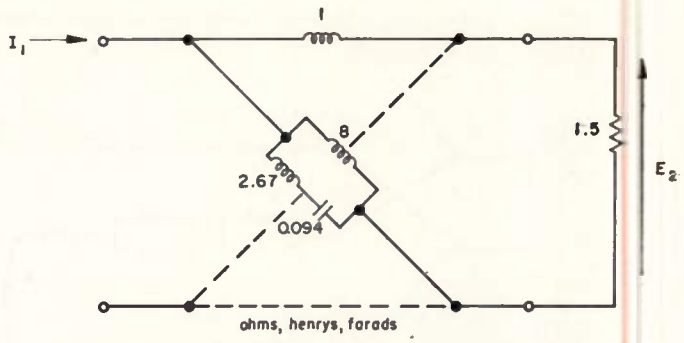

Fig. 11—Lattice equivalent to the one in Fig. 10.

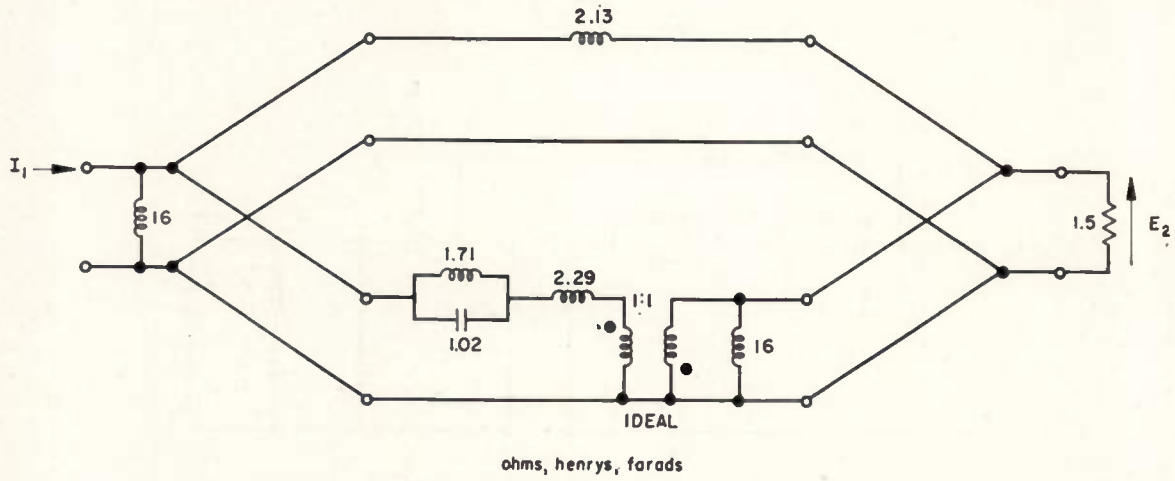

Fig. 12—Intermediate unbalanced network achieved for illustrative example.

It has thus been demonstrated that realization of each of the transfer functions yields the desired type of network.

III. Illustrative Example

We are given

$$Z_{12} = \frac{E_2}{I_1} = \frac{s^3 + 7s}{s^3 + s^2 + 3s + 1} = \frac{p}{q}$$

which we wish to realize within a multiplicative constant H as a lossless unbalanced network terminated in resistance and containing at most one real transformer.

We work with

$$K' = \frac{E_2}{E_1} = \frac{p}{Hq} = \frac{Z_b - Z_a}{Z_b + Z_a}.$$

Dividing the numerator and denominator by the even part of the denominator yields

$$K' = \frac{(s^3 + 7s)/(s^2 + 1)}{H[1 + (s^3 + 3s)/(s^2 + 1)]}.$$

We compute the residues and the values of H.

Pole at infinity:

$$k_1^{(n)} = 1$$
$$k_1^{(d)} = 1$$
$$\therefore \quad H = k_1^{(n)}/k_1^{(d)} = 1.$$

Pole at $s = \pm j1$

$$k_3^{(n)} = 3$$
$$k_3^{(d)} = 1$$
$$\therefore \quad H = k_3^{(n)}/k_3^{(d)} = 3.$$

Suppose we are interested in saving elements. Therefore we choose H equal to 3. Then

$$k_1^{(a)} = 1 \qquad k_3^{(a)} = 0$$
$$k_1^{(b)} = 2 \qquad k_3^{(b)} = 3$$

and the load resistance R for the Z_{12} is $R = H/2 = 1.5$ ohms. The lattice arms are

$$Z_a = s$$
$$Z_b = 2s + \frac{6s}{s^2 + 1} = \frac{2s(s^2 + 4)}{s^2 + 1}$$

and the lattice shown in Fig. 10 realizes

$$Z_{12} = (1/2)p/q.$$

We now proceed to convert the lattice to an unbalanced form. We find

$$Y_a = \frac{1}{s}$$
$$Y_b = \frac{1/8}{s} + \frac{3s/8}{s^2 + 4}$$

which yields the equivalent lattice shown in Fig. 11.

We first remove an inductance with an admittance value of $1/(16s)$ from each arm and then divide the lattice into two parallel networks. The $[Y_b - 1/(16s)]$ is now realized by a partial fraction expansion of its impedance:

$$Z = 1/[Y_b - 1/(16s)]$$
$$= \frac{16s(s^2+4)}{(s^2+4/7)}$$
$$= \frac{16s}{7} + \frac{48s}{49(s^2+4/7)}.$$

The network achieved thus far is shown in Fig. 12.

We now can replace the ideal transformer by a pair of coupled coils with a nonunity coupling coefficient. The final network is shown in Fig. 13.

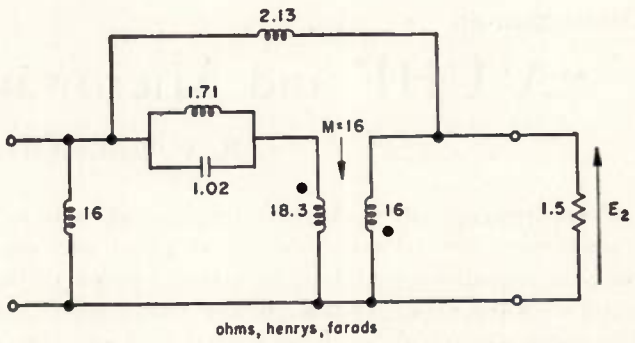

Fig. 13—Final network that realizes the given transfer function with the use of no ideal transformers and only one real transformer.

IV. Conclusion

In order that a transfer function be physically realizable by a lossless coupling network terminated in a single resistance, it must have an odd or even numerator. In this paper a method has been demonstrated for realizing any of these transfer functions by means of a lossless lattice network terminated in resistance. With the further restriction of an odd numerator for a transfer impedance and an even numerator for a transfer admittance or transfer voltage ratio, the achieved lattice can always be reduced to an unbalanced network with the use at most of one real transformer, where the real transformer is defined as one that has finite magnetizing inductance and a coupling coefficient smaller than one.

Correction

A list of corrections to "Elements of Boolean Algebra for the Study of Information-Handling Systems," by Robert Serrell, which appeared in the October, 1953 issue of the PROCEEDINGS OF THE I.R.E., is given below. Because of printing difficulties, a number of complementing bars (such as in $\overline{A}$) were missing. Suitable corrections cannot be listed for these missing bars, since their number and location depends upon the order in which copies came off the press. However, the author will be glad to supply error-free reprints on request.

Page 1366, Footnote 3, should read: [3] See B, p. 1368, for a detailed explanation of the ambiguity.
Page 1366, Footnote 6, line 2, should read: the use of parentheses within parentheses. See A, p. 1368.
Page 1367, Column 1, line 13, should read: nitions as are needed for an orderly development of the
Page 1367, Column 1, line 19, should read: work may safely be based—rather than to give an ex-
Page 1368, Column 2, line 4, should read: A. Any expression involving independent variables
Page 1368, Column 2, line 10, should read: An elementary function which is a union of inter-
Page 1370, Column 2, line 25, should read: $\{[F\cap(\overline{A}\cap\overline{B})]\cup[F\cap(\overline{A}\cap B)]\cup\cdots$
Page 1371, Column 1, line 15, should read: placed by the union of 2^γ terms, where $\gamma<n$ is the num-
Page 1371, Column 1, line 25, should read: We must now show that there is only one representation
Page 1371, Column 1, line 27, should read: by Theorem II, the minterms are all disjunct, distinct
Page 1371, Column 1, line 29, should read: Boolean functions. This completes the proof.
Page 1371, Column 1, line 37, should read: Theorem IV) we find successively:
Page 1371, Column 1, line 44, should read: $\cdots\cap(A\cup B\cup\overline{C})$.
Page 1371, Column 2, line 7 from bottom, should read: $F_M=\{(\overline{A}\cup\overline{B}\cup\overline{C})\cap(\overline{A}\cup B\cup C)\cap(A\cup\overline{B}\cup\overline{C})\cap\cdots$
Page 1371, Column 2, line 4 from bottom, should read: Note that, in view of Theorem Id, the order of the two steps
Page 1371, Footnote 20, line 6, should read: number N is smaller than 2^{264}. But with $n=9$, we find $2^{29}=2^{512}$. Thus,
Page 1372, Column 1, line 1, should read: and 2. Write down the intersection of all maxterms not
Page 1372, Column 2, line 29, should read: to F_{Rm} while Sm designates the set of minterms the
Page 1372, Column 2, line 37, should read: Theorem II, all distinct minterms are disjunct. In this
Page 1372, Column 2, line 38, should read: expansion, therefore, only idempotent intersections (of a minterm
Page 1372, Column 2, line 40, should read: Theorem XI. $(F_{Rm})\cap(F_{Sm})=F_{(Rm)\cap(Sm)}$.
Page 1374, Column 2, line 2, should read: $M_2\cap M_3=A$
Page 1375, Column 2, line 1, should read: $(\overline{A}\cup C)\cap(\overline{B}\cup C)\cap(\overline{A}\cup\overline{B})\cap(A\cup B\cup\overline{C})$
Page 1375, Column 2, line 7, should read: variable, and one containing three appearances of a variable.
Page 1375, Column 2, line 18, should read: Thus, by proceeding from the simpler elementary com-
Page 1375, Column 2, line 19, should read: ponents to the more complex, direct simplification results.
Page 1378, Column 2, line 14, should read: $\cdots\cap(\overline{A}_1\cup A_2\cup\overline{C}_i)\cap(A_1\cup\overline{A}_2\cup\overline{C}_i)\}$
Page 1378, Column 2, line 9 from bottom, should read: (physical) logical elements than those of examples I and II.
Page 1379, Column 2, line 27, should read: Using the superset representations of S and Co pre-

Discussion on

"A UHF and Microwave Matching Termination"*

R. C. ELLENWOOD AND W. E. RYAN

Prem Swarup:[1] At microwave frequencies, matching transmission line terminations are of great use, especially in impedance matching, or fixing a given voltage standing-wave ratio (VSWR) in the transmission line. The authors of the above paper have described a type of termination which makes use of a dual dielectric slug tuner and a lossy dielectric in contact with a short circuit.

The thickness of the tuner slugs of the less lossy dielectric (polystyrene) has been kept $\frac{3}{4}\lambda$ for X-band measurements and $5/4\lambda$ for K-band region. The attenuation of the lossy catalin dielectric is about 50 db/inches in the 3 cm region. The arrangement is shown in Fig. 1 of Ellenwood and Ryan.

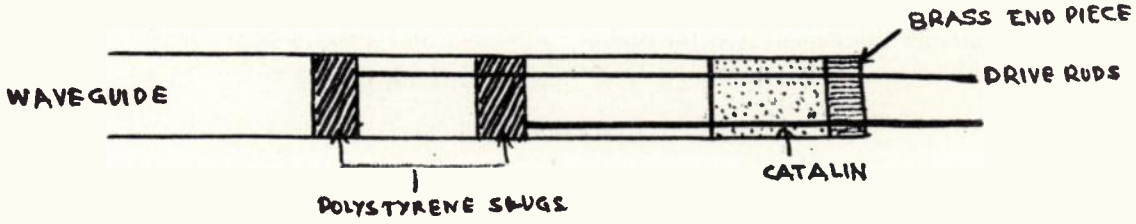

Fig. 1—Waveguide matching termination.

In such an arrangement, the tuner slugs (2.25 cm long), with crosssection just fitting the waveguide, are moved inside the waveguide till the tuning is perfect, by moving drive rods of polystyrene passing through the slugs. The authors say, that because of the presence of these drive rods and the loose fit of the slugs, accurate calculations of the electrical properties of the load are impractical.

In the case of the rectangular waveguide (X or K band), a thin slot along the length on the middle of the broad side may be made. The loss due to leakage through this slot is negligibly small (since its use is made in standing-wave detector; squeeze-section; flap attenuator etc.). The slugs may be placed inside the guide and their positions easily adjusted by moving them with a thin dielectric pin fixed in them and coming out of the slotted line. Such an arrangement will ensure better and easy adjustment of the termination.

The dielectric drive rods create unnecessary distortion in the radiation field, especially since their position near the short sides are likely to disturb the boundary conditions, change the electric path, and produce phase shift. The less lossy rods passing through lossy catalin will hinder in the uniform attenuation of the wave front, while rods passing through the brass endpiece will make it to some extent an imperfect short-circuit.

The present suggestion removes these unwanted disturbances and thus facilitates the calculation of the electrical properties of the load.

The author expresses his respectful thanks to Mr. Krishnaji, in charge of the Microwave Section.

R. C. Ellenwood and W. E. Ryan: Mr. P. Swarup in his comments on our paper suggests a design for a double-slug transformer for rectangular waveguide (similar to double-slug transformer models 109A and 146, manufactured by Sperry Gyroscope Co., Inc.) where the tuning slugs are positioned by dielectric pins extending through a slot in the waveguide. He states that such a design would be an improvement on ours, in that it would ensure easier adjustment, and remove disturbances within the termination, facilitating the calculation of its performance.

In reply, we would first like to point out that a detailed analysis of the theory and operation of the double-slug matching termination has already been made by Mr. Ellenwood and Dr. E. H. Hurlburt.[2] This analysis provided all the necessary design data for the construction and operation of our termination. The design parameters are not critical as is evident from a study of its experimental performance shown in Fig. 2 of our paper. The presence of the drive rods, the loose fit of the slugs, etc., have a negligible effect on the correlation between the design equations and performance.

Furthermore, the disturbances in the field referred to by Mr. Swarup occur, in our design, within the termination itself, where their presence is immaterial, and not, as in his suggested design, in the input guide, the region where a match is desired. If a slot is used (necessarily extending beyond the ends of the slugs in both directions), there are disturbances in this input region caused by reflections from the end of the slot[3] which, though small, will limit the minimum VSWR obtainable. These slot reflections, in addition, will vary as the slugs are moved back and forth. It should be emphasized again that the function of the termination is to remove reflec-

* R. C. ELLENWOOD and W. E. RYAN, "A UHF and microwave matching termination," PROC. I.R.E., vol. 41, pp. 104–107; January, 1953.

[1] Dept. of Physics, University of Allahabad, India.

[2] R. C. Ellenwood and E. H. Hurlburt, "The determination of impedance with a double-slug transformer," PROC. I.R.E., vol. 40, pp. 1690–1693; December, 1952.

[3] C. G. Montgomery, "Techniques of Microwave Measurements," Mass. Inst. Tech. Radiation Laboratory Series, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 11, pp. 480–483; 1947.

tions from the waveguide which it terminates, and as long as this purpose is accomplished, any reflections within the termination itself are immaterial.

Lastly, whereas our design is that of an adjustable sliding load, Mr. Swarup describes only a transformer. An important advantage of a sliding load is that a fixed probe detector (which can be made a part of the termination) can be used to determine the matched condition instead of a standing-wave machine. If he is proposing to adjust the lossy-dielectric slug also by a pin through the slot so that it operates as a sliding load, he neglects to state what arrangement he would use to fix the separation between the three slugs after they have been adjusted. Any carriage or superstructure on the guide to perform this function adds appreciably to its cost, as does the single requirement of a slot. In addition, it would not be easy to change sets of tuning slugs, and thin dielectric pins can easily be broken. A feature of our design for a matching termination is that it may be built simply, quickly, and inexpensively. For the above cited reasons, a slot was avoided in the design of this termination.

Determination of Equivalent Circuit Parameters for Dissipative Microwave Structures*

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Summary—Measurement procedures based upon Weissfloch's separation of a dissipative network into lossy and lossless portions are presented for the determination of the equivalent circuit parameters of dissipative four-terminal structures. The methods are developed around "distance invariant" equivalent networks which possess a number of desirable features in addition to permitting a simplification in the necessary calculations. They are particularly suited to such applications as the junction between surface wave or microstrip lines and their respective feed lines. Rapid and convenient semi-precision procedures are treated in some detail and sample calculations are presented.

Introduction

THE CONVENTIONAL three-point method for determining the equivalent circuit parameters of dissipative four-terminal structures is subject to the usual inaccuracies inherent in isolated point measurements so that a demand for increased precision requires some averaging scheme utilizing a series of input-output measurements. At the time of the development of the methods presented herein no satisfactory precision procedure had been noted in the literature. The highly satisfactory tangent relation method[1] for lossless structures cannot be adapted or extended to dissipative structures in any reasonably simple manner despite its formal applicability. The measurement methods proposed herein proceed from an analysis made by Weissfloch[2] where the dissipative network is separated into lossy and lossless portions; but the approach is modified so as to lead to a simplified equivalent circuit whose parameters may be determined precisely and whose over-all composition possesses a number of desirable properties. Since the completion of this work,[3] the authors have noticed a recently published approximate procedure[4] which closely resembles some of the material which has been presented,[3] but which is valid only for low-loss structures. A method[5] of obtaining by geometrical constructions the scattering matrix elements of a lossy (or lossless) four-terminal structure from a series of special input-output measurements has also been developed since that time.

We wish to distinguish between semi-precision and precision procedures, as the terms are employed here. In a semi-precision method a smooth curve is drawn through the measured set of data so that the curve may be taken to represent the average behavior of the data. The network parameters are then determined from selected points taken off the smooth curve. Although the evaluation of the equivalent circuit parameters thus depends on point values, it should be stressed that these values are obtained from an averaged curve so that no single datum point is given excessive weight. This is not true for a three-point method where an error in one of the directly measured values may appreciably affect the results. In a precision method the averaged data curve is compared with a known theoretical curve and the difference curve is minimized systematically, by either analytic or graphical means. A precision method, in addition, presents a way of estimating the error in the final results. When optimally precise values are not needed but accuracy greater than that afforded by the conventional three-point method is desired, the semi-precision methods are recommended; they maintain, to a large extent, the calculational simplicity of isolated point procedures. Since it is felt that greater interest will lie in a semi-precision method, and because of space limitations, the discussion of the precision method is omitted. Because of their usefulness and simplicity, the

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[1] N. Marcuvitz, "Waveguide Handbook," Radiation Laboratory Series, McGraw-Hill Book Co., Inc., New York, N. Y., vol. 10, sec 3-4; 1951.

[2] A. Weissfloch, *Elek. Nach. Tech.*, vol. 19; 1942.

[3] I.R.E. National Convention, New York, N. Y. Paper no. 78; March, 1951.

[4] D. D. King, "Measurements at Centimeter Wavelengths," D. VanNostrand and Co., Inc., New York, N. Y., p. 223; 1952.

[5] G. A. Deschamps, "Determination of reflection coefficients and insertion loss of a waveguide junction," *Jour. Appl. Phys.*, vol. 24, p. 1046; 1953.

the semi-precision methods are presented in a stepwise form which may be followed directly in the analysis of a set of data.[6]

The final equivalent network, termed the "canonical" network, resulting from the procedure described herein possesses a number of desirable properties. It contains the minimum possible number of elements and is strongly shunt or series in form (see Fig. 5; the normalized R_s or G_p is always <1). The network parameters are determined independently of any absolute distance measurements, a common source of error. This point is discussed later in some detail. The network is applicable directly to any asymmetric dissipative structure; it is particularly suited for such applications as the junction between surface wave or microstrip lines and their respective feed lines.

Weissfloch[2] shows that the parameters of the dissipative part of his divided network may be obtained directly from the input impedance circle which results when a series of reactive terminations is connected to the output terminals of the structure. He further proposes that the dissipative part be then subtracted off and the input-output impedance data for the over-all structure be reformulated as input-output reactance data for the lossless part alone, which is then to be analyzed by a graphical procedure. While the subtraction of the dissipative portion is necessary for a precision analysis, a much more rapid and elegant semi-precision procedure is presented herein which does not require this subtraction scheme. Weissfloch's method suffers from a number of additional drawbacks which are overcome by the modifications discussed in the next section.

Instead of restricting the analysis of the data, as Weissfloch does, to the particular pair of terminals located at the ends of the structure, a unique input reference plane is sought out. As a result of this step, one of the network elements of the lossy part is eliminated, the determination of the remaining elements is greatly simplified, and their values become independent of the absolute location of the input reference plane. This latter "distance invariant"[7] property is important from a precision point of view, since the location of the input reference plane is often determined by a mechanical measurement whose accuracy is almost always less than that of the electrical measurements employed. At a specially determined output reference plane, the lossless portion of the network contains only two (instead of the original three) network elements so that the over-all equivalent circuit (called the "canonical network") has been simplified still further to the strongly shunt or series form mentioned above.

In many cases, the structure to be measured is representable by a pure shunt or series network; examples of such structures are zero-thickness transverse resistive sheets (films), symmetrical thick radiating slots, etc. Such a representation is possible if and only if the structure is symmetrical and if the data circle is tangent to the unit circle in the input reflection coefficient plane. Since the semi-precision analysis becomes particularly simple for these networks, this analysis is included herein in stepwise form. Although a factor to be considered in all measurements, the effect of dissipation in the waveguide walls is particularly noticeable in the measurement of shunt or series structures since it prevents the exact tangency mentioned above. The effect of this dissipation may be rigorously taken into account, but the necessary computations are considered excessive. It is advisable in any case to compensate for the effect of the dissipation in at least an approximate manner if accurate results are desired.

Although the asymmetric canonical network may be employed to represent a symmetric structure, it is usually desirable to obtain for such a structure a symmetric network between reference planes disposed symmetrically about the geometrical symmetry plane of the structure. Formulas are given[6] permitting the canonical network to be converted into a symmetric one by shifting either the input or the output reference plane to a symmetric location. In the Appendix the methods of analysis described above are employed in connection with an actual set of data in order to demonstrate their applicability.

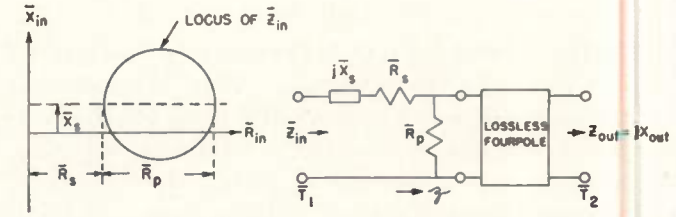


Fig. 1—Circuit proposed by Weissfloch.

Basic Theory

Equivalent Circuit Proposed by Weissfloch

If a series of reactive loads, best realized by a variable short circuiting plunger, is connected to the output terminals of a four-terminal structure, the locus of output impedance terminations Z_{out} in the complex Z_{out}-plane is the imaginary axis, which can be considered as a circle with center at infinity. Since the relation between Z_{in} and Z_{out} is a complex bilinear transformation which transforms circles in the Z_{out}-plane into circles in the Z_{in}-plane, the locus of Z_{in} in the Z_{in}-plane is also a circle which is always located entirely in the right half Z_{in}-plane (for a passive network) and, in general, has a finite radius (see Fig. 1(a)). The size and location of the impedance circle can be specified in a number of ways. The choice of the particular parameters $\bar{R}_s$, $\bar{X}_s$, $\bar{R}_p$ [see Fig. 1(a)] employed by Weissfloch[2] permits the

[6] Details of the precision procedure have been carried out completely and are presented in the report, "The Precision Measurement of the Equivalent Circuit Parameters of Dissipative Microwave Structures," by L. B. Felsen and A. A. Oliner, Microwave Research Institute, Polytechnic Institute of Brooklyn, Report R-282-52, PIB-221; March, 1953.

[7] N. Marcuvitz, *loc. cit.*, pp. 137–138.

simple deduction of an equivalent network. $\bar{R}_s$ and $\bar{X}_s$ are the co-ordinates of the point on the circle closest to the imaginary axis, and $\bar{R}_p$ is the diameter of the data circle. (The $^-$ serves to distinguish the parameters of this network from the network parameters employed later on; it does not denote a complex conjugate.) It is readily checked that the Weissfloch circuit in Fig. 1(b) yields the input impedance locus in Fig. 1(a) if the output terminations are reactive.

The Weissfloch method of separating the network into a lossless part, and a lossy part comprising the resistances $\bar{R}_s$, $\bar{R}_p$ and the reactance $\bar{X}_s$, offers important advantages as far as the analysis of the data is concerned. The three pieces of information obtained from the location and size of the data circle yield directly the three parameters of the lossy part of the network; this is an important advantage which is not realized if the analysis is undertaken directly for an assumed T- or π-network. Upon subtracting out the lossy part, one obtains input-output relations for the lossless part which can then be analyzed by any one of a variety of methods devised for the (precision) analysis of lossless structures. The lossless fourpole does not affect the input impedance circle as such; it simply determines the distribution of data points along the circle for given values of output reactance. For the analysis of the lossless part of the network, Weissfloch suggests a graphical method,[8] which consists essentially of the construction of a nomogram by means of which the input-output relation for a given lossless fourpole is uniquely specified.

It is apparent that a circuit dual to that in Fig. 1(b) can be constructed if the input admittance plane is taken as the starting point. With the circle parameters labeled as in Fig. 2(a), the equivalent circuit is that given in Fig. 2(b). Again, the parameters $\bar{G}_p$, $\bar{B}_p$, $\bar{G}_s$ of the lossy part are obtained directly from the input admittance locus.

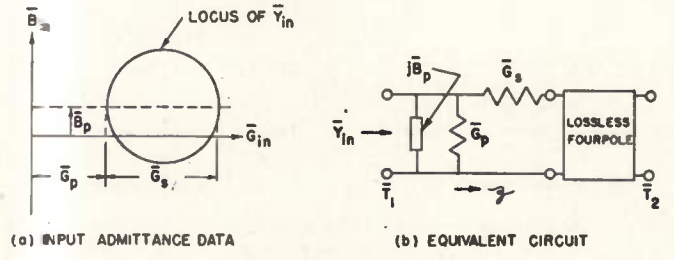


Fig. 2—The dual representation.

The Weissfloch method as outlined above has several features which do not make it well suited to a precision analysis of the data. The analysis proceeds entirely in the impedance (admittance) plane, with the subsequent possibility of sensitive data points with large amplitudes. Impedance values must be obtained for each measured value of VSWR and location of voltage minimum, thereby increasing the necessary number of calculations. Upon subtraction of the lossy part, the lossless network is analyzed in terms of reactances which can also assume very large magnitudes so that a precision analysis here becomes questionable again, although the graphical method advanced by Weissfloch can, in principle, average a set of data points. Finally, the impedance locus becomes very large if the loss in the structure to be measured is small. As pointed out by Weissfloch, it is more desirable for the latter case to perform the analysis in the reflection coefficient plane, where the locus is again a circle. However, the parameters of the lossy part are now given by the co-ordinates of that circle which intersects the data circle and the unit circle orthogonally, and the analytical determination of these points of intersection in terms of the location and size of the data circle is somewhat cumbersome. It can be said in the nature of a general comment on the Weissfloch circuit that the equivalent network is valid between those reference planes with respect to which the measurements are taken, and no attempt is made by Weissfloch to seek out alternative reference planes between which the resulting network might exhibit more desirable properties.

[8] A. Weissfloch, *Hochfreq. und Elektroak.*, vol. 61; April, 1943.

Analysis of the Lossy Part

The observations made in the last paragraph of the preceding section form the starting point in the attempt to modify the Weissfloch procedure into one more suited to the precision and semi-precision analysis of lossy structures. The first step is to carry out the analysis exclusively in the reflection coefficient plane. This eliminates not only the objection to the possible high sensitivity of certain input impedance data points to errors in the measurements, but reduces the total number of calculations involved since the input reflection coefficients Γ are quite simply obtained in terms of the measured values of VSWR and location of voltage minimum:

$$\Gamma_{in} = |\Gamma| e^{j\theta_{\bar{T}_1}}, \qquad |\Gamma| = \frac{(VSWR) - 1}{(VSWR) + 1}. \tag{1}$$

$$\theta_{\bar{T}_1} = 2\kappa_D d + \pi, \quad d = z_{\bar{T}_1} - z_{\min}, \quad \kappa_D = 2\pi/\lambda_{gD}.$$

$z_{\bar{T}_1}$ and $z_{\min}$ denote, respectively, the location of the input reference plane and the voltage minimum, and λ_{gD} is the input guide wavelength; z is the variable along the direction of propagation. The loci of the input data in the impedance and reflection coefficient planes are related by the well-known bilinear transformations

$$\Gamma_{in} = \frac{\bar{Z}_{in} - 1}{\bar{Z}_{in} + 1}, \quad \bar{Z}_{in} = \frac{1 + \Gamma_{in}}{1 - \Gamma_{in}}, \quad \bar{Z}_{in} = \frac{1}{\bar{Y}_{in}}, \tag{2}$$

where the impedance values are normalized to the guide characteristic impedance. The input data circle in the reflection coefficient plane is characterized by r_0, θ_0, the co-ordinates of its center, and its radius ρ.

When this data circle is rotated so that its center lies on the real axis, the expression for the parameters of the

lossy part becomes greatly simplified, and, in addition, the series reactance $\bar{X}_s$ in the equivalence circuit of Fig. 1(b) vanishes. The vanishing of the series reactance follows from the fact that the real axis in the $\bar{\Gamma}_{in}$-plane maps into the real $\bar{Z}_{in}$-axis, and a circle with its center on the real $\bar{\Gamma}_{in}$-axis maps into a circle whose center lies on the real $\bar{Z}_{in}$-axis. The rotation of the data circle is accomplished quite simply by a shift of the input reference plane since the reflection coefficients at the two planes $z_{\bar{T}_1}$ and z_{T_1} are related by:

$$\Gamma(z_{T_1}) = \bar{\Gamma}(z_{\bar{T}_1}) e^{j2\kappa_D(z_{T_1} - z_{\bar{T}_1})}. \tag{3}$$

Thus, if the data circle in the $\bar{\Gamma}_{in}$-plane is rotated so that its center moves about the origin at a radius r_0 until it lies, say, on the negative real axis, this rotation is compensated for by moving the input reference plane $\bar{T}_1$ in Fig. 1(b) to a new location z_{T_1}, where

$$z_{T_1} = z_{\bar{T}_1} - \frac{\theta_0 - \pi}{2\kappa_D}. \tag{4}$$

The shifted data circle and the new equivalent circuit corresponding to it are shown in Fig. 3. The (normalized) resistances R_s and R_p are given in terms of the circle parameters in the Γ_{in}-plane by (cf. (2)):

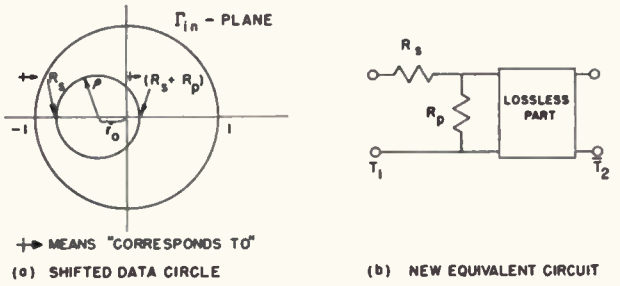


Fig. 3—Effect of the shift of the input reference plane.

$$R_s = \frac{1 - (r_0 + \rho)}{1 + (r_0 + \rho)}, \qquad R_p = \frac{1 - (r_0 - \rho)}{1 + (r_0 - \rho)} - R_s. \tag{5}$$

The lossless part in Fig. 3(b) is different from that in Fig. 1(b). It is seen from (2) and (5) that the series parameter R_s is smaller in magnitude if the center of the data circle is shifted to the negative real Γ_{in}-axis [Fig. 3(a)] than if it is shifted to the positive real axis. Therefore, a shift to the negative real axis is desirable if the "shunt" character of a network is to be stressed, as is the case in the representation in Fig. 3(b).

Besides leading to the simple determination of R_s and R_p in (5) and eliminating the reactive elements $\bar{X}_s$ entirely, the new representation in Fig. 3(b) has additional important advantages as far as a precision analysis of the data is concerned. The elements R_s and R_p are a function only of the radius of the data circle and the distance of its center from the origin of coordinates. They are independent of the location of either the old reference plane $\bar{T}_1$ or the new reference plane T_1 so that they are not affected by errors in the reference plane location (i.e., the value of the reference plane shift) or absolute distance measurements which may be employed to locate the reference planes. Thus, they are distance invariant parameters.[7] By virtue of the simplicity of (5), possible errors in the parameters R_s and R_p are simply related to errors in the initial data [cf. (1)] so that their sensitivities to errors in measurement are readily formulated.

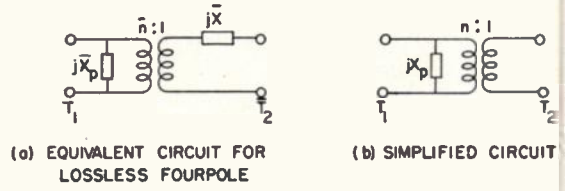


Fig. 4—Network representations for the lossless part.

The Final Canonical Form

From the knowledge of the dissipative network parameters R_s and R_p and the location of the new input reference plane T_1, the remaining lossless portion of the network can be analyzed either by subtracting off the lossy portion and obtaining input-output relations for the lossless part alone, or by dealing directly with the over-all network. The former procedure is employed when a precision analysis of the data is required; the latter, more rapid and elegant approach is used for semi-precision analyses. To apply the two alternative procedures, it is desirable for the former and necessary for the latter that the detailed structure of the lossless part of the network be known. Since there exist a variety of equivalent circuit representations for a lossless fourpole, it is useful to list the desired properties of the equivalent circuit as a guide in the selection of a representation. If the circuit is to have as few elements as possible and, in addition, to be readily combined with the preceding lossy portions, we are led to the representation shown in Fig. 4(a) which may be reduced to that in Fig. 4(b) by a shift of the output reference plane to a new location T_2.[9] (The shift of the output reference plane does not affect the lossy parameters R_s and R_p because these parameters depend only on the size and location of the input data circle which does not vary with the output reference plane location.) The final "canonical" circuit for the lossy fourpole obtained upon combining the separate circuits in Figs. 3(b) and 4(b) is shown in Fig. 5(a).

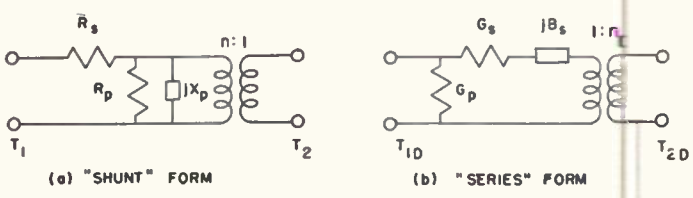


Fig. 5—Canonical networks. (The subscript "D" signifies "dual.")

In direct analogy to the reasoning employed to arrive at the canonical network in Fig. 5(a) from the Weiss-

[9] N. Marcuvitz, *loc. cit.*, Figs. 3.13(b) and 3.17.

flow circuit in Fig. 1(b), we may obtain the dual network (containing admittance elements) in Fig. 5(b) from that in Fig. 2(b). Since the dual network emphasizes the "series" nature of the network, the data circle in the Γ_{in}-plane is shifted to that its center lies on the positive real axis (i.e., the shunt admittance G_p is kept small). Therefore, the locations of T_1 and T_{1D} differ by a quarter wavelength. The (normalized) dissipative network elements G_p and G_s are expressed in terms of the circle parameters r_0 and ρ as in (5) provided that the following duality replacements are made:

$$R_s = G_p, \qquad R_p = G_s. \tag{6}$$

Semi-Precision Methods

This section is concerned with the analysis of the input-output data taken on a particular lossy structure to yield the numerical values of the network parameters characterizing it. It is assumed that the output termination is a variable short-circuit, for each setting S of which the corresponding input measurements of the VSWR and the location of the voltage minimum (corresponding to a reading D on the standing-wave indicator) have been taken. As is customary, the distance measurements are referred to the ends of the physical structure defined by the reference planes $\overline{T}_1$ and $\overline{T}_2$ in Fig. 6, and located at the setting $D=D_R$ and $S=S_R$ on the standing-wave indicator and variable short-circuit scales, respectively. Thus, the actual distance from the short-circuiting plunger to the output reference plane $\overline{T}_2$ is $(S-S_R)$; the actual distance from the corresponding voltage minimum to the input reference plane $\overline{T}_1$ is $(D-D_R)$. It is also assumed that the given sets of values of D, S, VSWR, and the values of D_R and S_R have been corrected from appropriate calibration curves so these data represent effect of structure only and do not include errors inherent in measuring equipment.[10]

The methods of analyzing the data to be discussed in the present section are of a semi-precision type. They make use of special point values in the evaluation of some of the circuit parameters. However, these values do not correspond to individual point measurements; they are taken off a smooth curve drawn through the individual data points so that the average behavior of the data is emphasized. Therefore, the results should be considerably better than those obtained from a three-point analysis, where the entire weight rests on three individual measurements. It will be seen below that very little computational work is involved in a semi-precision analysis. Since input-output data are obtained with little effort, the requirement of a relatively large number of data points does not constitute a major inconvenience. Thus, a semi-precision analysis is definitely to be preferred over a three-point analysis if the results are to be reliable.

[10] Cf. A. A. Oliner, "The Calibration of the Slotted Section for Precision Microwave Measurements," Microwave Research Institute Polytechnic Institute of Brooklyn, Report R-267-52, PIB-206; May 1952. Accepted for publication in the *Rev. Sci. Instr.*

The curves required for a semi-precision analysis of the data are: 1) the D vs. S curve; and 2) the locus in the input reflection coefficient plane. The plot of the D values vs. the values of S to which they correspond is made directly from the data. The input reflection coefficients are computed from (1). In order to minimize the number of steps in the calculations we calculate not $\theta_{\bar{T}_1}$, i.e., the phase of Γ_{in} at $\overline{T}_1$, but simply

$$\theta = 2\kappa_D D, \tag{7}$$

and remember that the locus so obtained must be shifted through a positive angle $(\pi - 2\kappa_D D_R)$ to obtain $\theta_{\bar{T}_1}$. Through the plotted points in the Γ'_{in}-plane ($\Gamma'_{in} = |\Gamma_{in}| \exp j\theta$) we draw a circle with a compass, attempting to accommodate the data points as well as possible.

The Canonical Network

From the data circle, the circle parameters r_0, θ_0' and ρ are obtained directly (θ_0' denotes the value of of θ_0 in the Γ'_{in}-plane). The resistive parameters R_s and R_p are calculated from (5). The location of the input reference plane T_1 is found from (4), provided that we write:

$$\theta_0 = \theta_0' + \pi - 2\kappa_D D_R; \qquad z_{T_1} - z_{\bar{T}_1} \to D_R - D'. \tag{8}$$

The second relationship in (8) follows from the fact that D and z are measured positive in opposite directions (cf. Fig. 6).

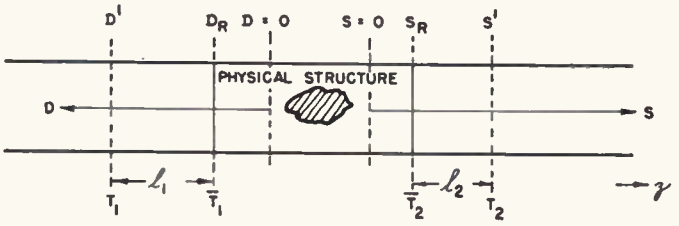


Fig. 6—Specification of the reference planes.

The input reflection coefficient locus produced by a variable lossless termination on the canonical network in Fig. 5(a) is shown in Fig. 3(a). In order to obtain information about the remaining parameters in the lossless part of the canonical network, special points on the input data circle are analyzed and interpreted in terms of the canonical circuit. If a short-circuit is placed at the output reference plane T_2 of the canonical circuit in Fig. 5(a), the input impedance is real and equal to its minimum value R_s. The corresponding input reflection coefficient is $\Gamma_{min} = -(r_0+\rho)$ [cf. Fig. 3(a)]. Since S denotes the location of the short-circuit at the output side of the structure, the special value S' of S which gives rise to the minimum (real) value of input impedance locates the output reference plane T_2. Since the input impedance is real and less than the characteristic impedance Z_0 of the input guide, the voltage minimum occurs at the input terminals of the network, i.e., at $D=D'$. Therefore, S' is found by looking up the value

of S corresponding to D' (and Γ_{min}) on the D vs. S curve.

If the input impedance is real and equal to its maximum value (R_s+R_p), the output termination is $X_{out}=-X_p/n^2$. The corresponding input reflection coefficient is $\Gamma_{max}=-(r_0-\rho)$. A voltage maximum exists at T_1 if Γ_{max} is >0. If we look up the value S_1 corresponding to $D_1=D'\pm(\lambda_{gD}/4)$[11] from the D vs. S curve, we may compute the output reactance X_{out} of the length (S_1-S') of short-circuited transmission line connected to the output terminal plane T_2. Thus, we have evaluated the ratio (X_p/n^2). If $\Gamma_{max}<0$, a voltage minimum occurs at T_1 so that $D_1=D'$. In this case there are two values of S corresponding to D': S' and S_2; S_2 corresponds to Γ_{max}.

The reactance X_p may be determined by placing an open-circuit at T_2, i.e., $S_2=S'\pm(\lambda_{ge}/4)$.[11] From the value D_2 corresponding to S_2 on the D vs. S curve, the phase of the input reflection coefficient Γ_2 is computed. The magnitude of Γ_2 is read off the Γ_{in}-circle. This is done in preference to obtaining it from a VSWR vs. S curve, since the point may fall in a region of large VSWR which usually cannot be measured accurately. In terms of Γ_2, the corresponding input impedance may be formulated from (2) and equated to the input impedance of the open-circuited network in Fig. 5(a). The resulting equation is readily solved for X_p. With X_p known, n may be computed from the previously obtained value of the ratio X_p/n^2.

We now summarize in step form the procedure of analyzing a lossy four-terminal network by a semi-precision method leading to the canonical network in Fig. 5(a). It is assumed that the following data is given: the parameters r_0, θ_0', ρ of the data circle in the Γ_{in}'-plane; the D vs. S curve; the guide wavelengths λ_{gD} and λ_{ge}; the location D_R and S_R of reference planes $\overline{T}_1$ and $\overline{T}_2$, respectively.

1. R_s and R_p:

$$R_s=\frac{1-(r_0+\rho)}{1+(r_0+\rho)};$$

$$R_p=\frac{1-(r_0-\rho)}{1+(r_0-\rho)}-R_s=\frac{4\rho}{(1+r_0)^2-\rho^2}.$$

2. Location of T_1:

$$D'=\frac{\theta_0'}{4\pi}\lambda_{gD}.$$

Shift from $\overline{T}_1$ to T_1: $l_1=D'-D_R$.

3. Location of T_2:

Find S' corresponding to D' from D vs. S curve (if there are two values of S', pick that which corresponds to the larger value of $|\Gamma|$).

Shift from $\overline{T}_2$ to T_2: $l_2=S'-S_R$.

[11] λ_{gD} and λ_{ge}, are the wavelengths in the guides connected to the input and output terminals of the structure, respectively.

4. X_p:

Find D_2 corresponding to $S_2=S'\pm(\lambda_{ge}/4)$ from D vs. S curve.

Find $|\Gamma_2|$ corresponding to $\alpha=4\pi D_2/\lambda_{gD}$ on the data circle in the Γ_{in}'-plane.

Compute $\Phi_2=$ phase of Γ_2 at $T_1=\alpha-\theta_0'\pm\pi$.

$$\Gamma_r=|\Gamma_2|\cos\Phi_2, \qquad \Gamma_i=|\Gamma_2|\sin\Phi_2.$$

Then

$$X_p=\frac{[1-R_s+\Gamma_r(1+R_s)]^2+(1+R_s)^2\Gamma_i^2}{2\Gamma_i}.$$

5. n:

Find S_1 corresponding to $D_1=D'\pm(\lambda_{gD}/4)$ (if $\rho>r_0$) or to $D_1=D'$ (if $\rho<r_0$) from the D vs. S curve (if $\rho<r_0$, pick that value of S_1 which corresponds to the smaller value of $|\Gamma|$).

$$n^2=-X_p\cot(2\pi/\lambda_{ge})(S_1-S').$$

The analysis leading to the dual circuit in Fig. 5(b) may be carried out in an analogous manner. The stepwise procedure above can be applied directly to the analysis of the dual network provided that the appropriate duality replacements are made. In this manner it is found that the network elements in Figs. 5(a) and (b) are related as follows:

$$R_s=G_p, \qquad R_p=G_s, \qquad X_p=B_s, \qquad n^2=n_D^2,$$

$$l_1=l_{1D}\pm\frac{\lambda_{gD}}{4}, \qquad l_2=l_{2D}\pm\frac{\lambda_{ge}}{4}. \tag{9}$$

l_{1D} and l_{2D} are the shifts from the data reference planes $\overline{T}_1$ and $\overline{T}_2$ to the canonical reference planes T_{1D} and T_{2D}. Thus, at reference planes shifted by a quarter wavelength, the (normalized) impedance elements of the "shunt" network are identical with the dual (normalized) admittance elements of the "series" network. In view of this simple general relationship, no separate analyses leading to dual networks need be included in our discussion.

Pure Shunt (or Series) Networks

The analysis for structures which are representable by pure shunt or pure series networks becomes particularly simple. Examples of such structures are zero-thickness transverse resistive sheets (films), symmetrical radiating slots of appreciable wall thickness, etc. Shunt (or series) networks are characterized by $R_s=0$ (or $G_p=0$) in the canonical networks of Fig. 5, i.e., the data circle is tangent to the unit circle at some point. When the structure is in addition symmetrical, the turns ratio n (or n_D) becomes unity, producing the pure shunt or series networks shown in Fig. 7, and appreciable simplification occurs in the analysis.

Dissipation in the guide walls adjacent to the discontinuity being measured serves to prevent the exact tangency between the data and unit circles unless the effect of such dissipation is adequately accounted for. If deviation from exact tangency occurs in spite of ex

pectations due to theoretical considerations concerning the structure, it is advisable to draw the data circle so as to force tangency. When tangency obtains, $r_0+\rho=1$, and the simplications

$$R_s = 0, \qquad R_p = \rho/r_0, \qquad n = 1,$$

follow for pure shunt network. Considerations for dual network follow those given in connection with (9).

The procedure in stepwise form for obtaining the pure shunt network of Fig. 7(a) is presented below. The data

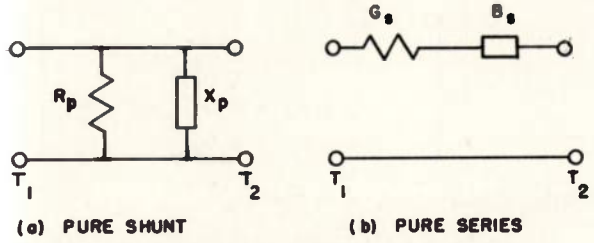


Fig. 7—Pure shunt and pure series networks.

given is the same as assumed in stepwise procedure above.

1. R_p: $R_p=\rho/r_0$.

2.–3. Locations of T_1 and T_2. Same as above.

4. X_p: Find S_1 from the D vs. S curve as in step (5) above. Then:

$$X_p = -\tan\frac{2\pi}{\lambda_{gs}}(S_1 - S').$$

Appendix—Sample Calculation

We now carry through the semi-precision analysis of a set of measurements taken on a symmetric slot radiating from the broad face of ½ inch×1 inch rectangular waveguide. In order to save space, the steps followed in obtaining the different parameters are not repeated here; instead, only the results are listed. The D vs. S curve is plotted in Fig. 8(a). The input locus in the Γ_{in}-plane calculated from the measured values is plotted in Fig. 8(b). The "best" circle drawn through the points with a compass is seen to be specified by the parameters: $r_0=0.310$, $\theta_0'=187.5°$, $\rho=0.670$. It should be stressed that an adequate analysis can be carried out with considerably fewer data points than have been employed here.

Following the stepwise procedure for the canonical network, we calculate the following values for its parameters:

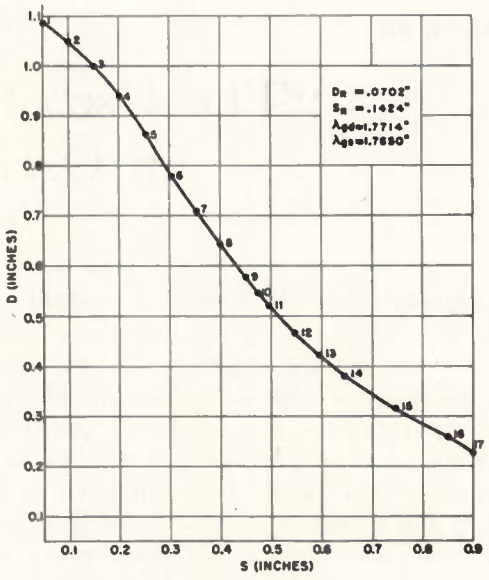
$$R_s = 0.01,\ R_p = 2.12,\ D' = 0.460'',\ l_1 = 0.391''. \quad (10)$$

From Fig. 8(a) it is seen that $S'=0.555''$, $S_2=0.114''$, $D_2=1.037''$. Also, $|\Gamma_2|=0.435$, so that

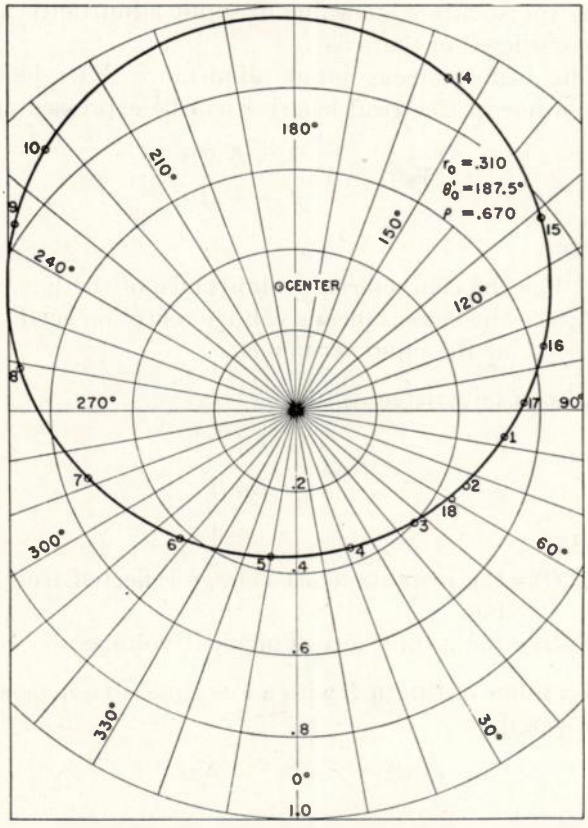
$$X_p = 2.37,\quad n = 0.996,\quad l_2 = 0.413''. \quad (11)$$

Because $R_s \ll 1$ and $n \approx 1$, the equivalent circuit is almost pure shunt.

Since the slot structure itself is symmetrical, it may be desirable to represent the final circuit in symmetrical form. For example, the final network can be presented as a symmetrical Tee between the canonical input reference plane and an appropriately shifted output reference plane.[6]



(a) D vs. S curve.



(b) Input data circle in reflection coefficient plane.

Fig. 8—Data curves.

Acknowledgment

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The authors thank H. M. Altschuler and H. Kurss, of Microwave Research Institute, for their helpful comments and suggestions. We are indebted to H. Kurss for his participation in early stages of this work, and to H. M. Altschuler for his help in testing applicability of these methods, and his critical review.

Discussion on

"The Use of Admittance Diagrams in Oscillator Analysis"*

H. J. REICH

L. I. Algase:[1] There is, I believe, a significant addition that might be made to Dr. Reich's excellent paper above. It concerns the performance of an oscillator during the build-up of oscillations, and would take into account the varying input admittance seen by a generator whose voltage amplitude is varying at a rapid rate, when an electrically long line such as that described by the author in the section "Long-line Effect" is used. Since the voltage amplitude of an oscillator generally does increase rapidly during the build-up of oscillations, it is suggested that the transient rather than the steady-state value of input admittance must be considered in that case.

The instantaneous input admittance $Y_i(t)$ looking into a line of electrical length θ can be expressed as

$$Y_i(t) = Y_0 \frac{1 - K_i(t)\epsilon^{-j2\theta}}{1 + K_i(t)\epsilon^{-j2\theta}} \tag{1}$$

where

Y_0 = the characteristic admittance of the line,
$K_i(t)$ = the instantaneous voltage reflection coefficient at the input terminals.

$K_i(t)$ can be written as

$$K_i(t) = \frac{E_{ri}(t)}{E_{ii}(t)} \tag{2}$$

where

$E_{ri}(t)$ = the instantaneous voltage reflected from the line,
$E_{ii}(t)$ = the instantaneous incident voltage.

For a line of length L with a loss equal to α nepers per unit length,

$$E_{ri}(t) = E_{ii}(t - T)K_L\epsilon^{-2\alpha L} \tag{3}$$

where

T = the round-trip delay in the line,
K_L = the voltage reflection coefficient at the load end of the line $= |K_L|\epsilon^{j\phi}$.

Since the assumption does not affect the rigor of the analysis, a lossless line is assumed, and

$$E_{ri}(t) = E_{ii}(t - T)K_L. \tag{4}$$

Combining (1), (2), and (4), gives for the instantaneous input admittance

$$Y_i(t) = Y_0\left[\frac{1 - \dfrac{|E_{ii}(t - T)K_L|\,\epsilon^{j(\phi-2\theta)}}{|E_{ii}(t)|}}{1 + \dfrac{|E_{ii}(t - T)K_L|\,\epsilon^{j(\phi-2\theta)}}{|E_{ii}(t)|}}\right]. \tag{5}$$

Under steady-state conditions,

$$|E_{ii}(t - T)| = |E_{ii}(t)| \tag{6}$$

and (5) reduces to the usual expression for the steady-state input admittance.

Equation (6) does not hold, however, for input admittance during the build-up of oscillations. During that period, the amplitude of the incident voltage is not constant and is given by[2]

$$E_{ii}(t) = \gamma A\beta \exp\left[-\frac{1}{2C}\int_{t_1}^{t} G(t)dt\right] \tag{7}$$

where

γ = a coupling factor relating the voltage across the resonant circuit of the oscillator to that across the transmission line
A = the amplitude of a signal present when oscillations start
β = a constant depending upon the resonant-circuit parameters
C = the effective shunt capacitance across the resonant circuit
t_1 = the instant at which $G(t)$ becomes negative
$G(t)$ = the total shunt conductance across the resonant circuit defined by Dr. Reich as $G_e + G_c$ in the paper under discussion. It is the sum of the electronic conductance and the circuit conductance.

If it is assumed that there are no discontinuities in $G(t)$, the magnitude of the reflection coefficient at the input terminals of the line during build-up oscillations is

$$K_i(t) = \left|\frac{K_L\gamma A\beta \exp\left[-\dfrac{1}{2C}\displaystyle\int_{t_1}^{t-T} G(t)dt\right]}{\gamma AB \exp\left[-\dfrac{1}{2C}\displaystyle\int_{t_1}^{t} G(t)dt\right]}\right|$$

$$= \left|K_L \exp\left[+\frac{1}{2C}\int_{t-T}^{t} G(t)dt\right]\right|. \tag{8}$$

* H. J. Reich, "The use of admittance diagrams in oscillator analysis," PROC. I.R.E., vol. 41, pp. 522–528; April, 1953.

[1] Airborne Instruments Laboratory, Mineola, N. Y.

[2] J. R. Whitehead, "Super-Regenerative Receivers," Cambridge University Press, Cambridge, England, chap. 3; 1950.

Since $G(t)$ is negative during the build-up of oscillations, there will be a reduction in the effective reflection coefficient during that period. The amount of reduction is determined by the cable delay and the magnitude of $G(t)$. As $G(t)$ approaches zero (steady-state oscillations), the magnitude of the integral decreases, and $|K_i(t)|$ approaches its steady-state value $|K_L|$.

Therefore, during the build-up of oscillations the admittance looking into the line will vary from

$$Y_i(t) = Y_0 \frac{1 - \left| K_L \exp\left[\frac{1}{2C}\int_{t-T}^{t} G(t)dt\right]\right| \epsilon^{j(\phi-2\theta)}}{1 + \left| K_L \exp\left[\frac{1}{2C}\int_{t-T}^{t} G(t)dt\right]\right| \epsilon^{j(\phi-2\theta)}} \quad (9)$$

to

$$Y_i(t) = Y_0 \frac{1 - |K_L| \epsilon^{j(\phi-2\theta)}}{1 + |K_L| \epsilon^{j(\phi-2\theta)}}. \quad (10)$$

The effect of the time-varying input admittance upon oscillator performance is calculated by successive approximations. The total circuit admittance ($Y_c = G_c + jB_c$) contains a component linearly proportional to the line input admittance $Y_i(t)$. However, since $Y_i(t)$ is dependent in turn upon the total circuit conductance $G(t) = G_c + G_e$, it is necessary to assume a value of $G(t)$ in order to evaluate Y_c. Fig. 1, based on Fig. 9 in Dr. Reich's paper, is given as an illustration. Here the variation of circuit susceptance with frequency is assumed to be linear. In this example the magnitude of $G(t)$ at the start of oscillations is assumed to be large enough to make $K_i(t)$ approximately equal to zero and $Y_i(t)$ approximately equal to Y_0. Therefore the locus of Y_c is the straight line shown. The build-up rate and frequency at the start of oscillations are determined by points 1 and 2 rather than 1 and 3. It is necessary to evaluate (5) for $K_i(t)$ to determine if the choice of $K_i(t)$ equal to zero was a good one. If it was not, a value of $K_i(t)$ greater than zero but less than K_L must be selected, Y_c modified accordingly, and (5) re-evaluated until a suitable value is found. This process is repeated over incremental portions of the conductance curves as oscillations build up. During build-up, $G(t)$ will decrease. Therefore $K_i(t)$ will approach K_L, and the locus of Y_c will vary from the straight line assumed at the start of oscillations to the cycloid shown by Dr. Reich. Therefore, steady state $[G(t) = 0]$ occurs at point 5 rather than at point 4.

The steady-state conditions would indeed be those predicted by Dr. Reich. However, it is submitted that accurate prediction of the performance during the transient build-up condition entails careful evaluation of the variation in the input admittance of the long line during that period.

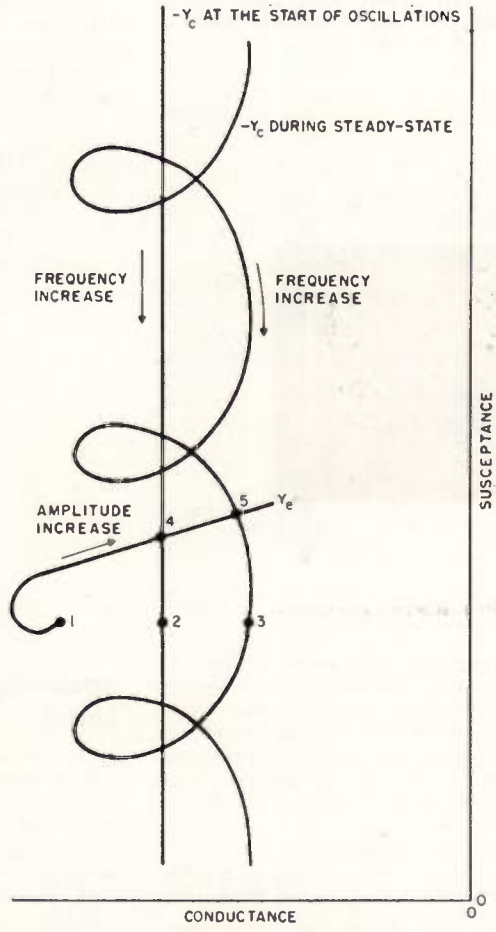

Fig. 1—Modified admittance diagram of a two-terminal oscillator loaded by a long transmission line terminated in an unmatched load.

H. J. Reich: Mr. Algase's extension of the method of oscillator analysis summarized in my paper is interesting, particularly since it helps to emphasize the limitations of this method, which were pointed out in my article. The method of successive approximations suggested by him would seem to be quite laborious and therefore possibly of questionable value as a practical tool in the quantitative prediction of the performance of an oscillator loaded through a long line. It does, however, represent a closer approximation in the solution of an involved problem in nonlinear circuits.

Correspondence

Silicon Surface-Barrier Transistors*

In the group of papers entitled, "The Surface-Barrier Transistor," which appeared in the December 1953 issue of the PROCEEDINGS, we presented the theory, the fabrication techniques, and circuit-application information for germanium Surface-Barrier-Transistors. These Surface-Barrier Transistors

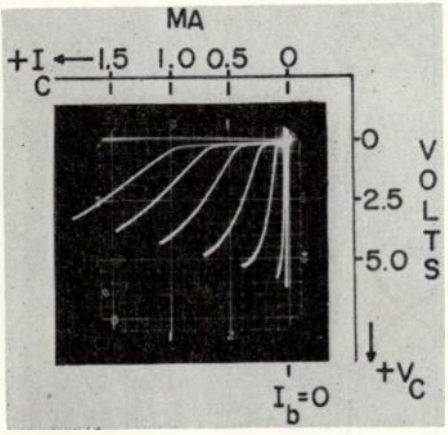

Fig. 1—Typical silicon surface-barrier transistor common emitter collector characteristics as a function of ΔI_b. ΔI_b steps equal 10μa.

employed metal-to-semiconductor interfaces as the minority-carrier-emitter and collector.

The electrochemical jet techniques used on the germanium transistors have been successfully applied to silicon. Experimental silicon Surface-Barrier Transistors have been fabricated with α's greater than .95 and $f_{c\alpha}$ greater than 10 mc.

Very truly yours,

W. E. Bradley

Philco Corporation, Phila., Pa.

* Received by the Institute, December 31, 1953.

A System for Reordering Modes*

The possible solutions to Maxwell's equations for a boundary value problem are usually divided into two classes and called modes, i.e., TE_{pq} and TM_{pq} modes. Associated with each mode, there are two indexes, i.e., p and q. Also the solutions can be divided into classes of progressing in a positive direction and progressing in a negative way.

When it is necessary to consider all of these modes when making a theoretical study of a boundary value problem, the notation and mathematical treatment can often be greatly simplified by combining all modes progressing in one direction into a single class with a single index. This can be done in many ways.

The following system, which is based on a scheme which is used to show that there are a countable number of rational numbers,[1] is a convenient one. The arrays shown in Table I can be used for denoting the TE_{pq} and TM_{pq} modes by a single index. An index in the arrays denotes the pair of indexes whose first index is the row in which it appears and whose second index is the column in which it appears. For example, mode 15 is the $TE_{1,2}$ mode.

It is realized that the fields associated with some modes may be zero. This does not require that these modes be removed from Table I since their inclusion in any theoretical study introduces no error.

TABLE I
ARRAYS FOR DENOTING *TE* AND *TM* MODES BY A SINGLE INDEX

	0	1	2	3	4
0	1	3	11	13	29
1	5	9	15	27	·
2	7	17	25	·	·
3	19	23	·	·	·
4	21	·	·	·	·

For *TE* Modes

	0	1	2	3	4
0	2	4	12	14	30
1	6	10	16	28	·
2	8	18	26	·	·
3	20	24	·	·	·
4	22	·	·	·	·

For *TM* Modes

H. F. MATHIS

University of Oklahoma

Norman, Oklahoma

Attenuation Calculations for UHF Lines*

In making accurate direct attenuation measurements on a section of low loss transmission line at UHF, severe requirements are placed on the transducers and, in general, a long length of line is required. In working with experimental sections of line both requirements frequently become a problem. In recent investigations of some experimental lines an adaptation of the incremental inductance rule[1,2] was made which allowed hf resistance to be determined using relatively short sections of line, and without concern for transducers. Once hf resistance is determined, total attenuation is calculated in the conventional way.[3]

The hf resistance, taking "proximity effect" into account of parallel conductors supporting TEM waves, is related to the derivative of the external inductance with respect to the conductor radii:

$$R_i = \frac{R_s}{\mu}\frac{dL}{dn} \qquad (1)$$

where

R_i = resistance of transmission line per unit of length

R_s = intrinsic resistance of conductor

μ = permeability of conductor

L = inductance of transmission line per unit of length

n = distance normal to conductor axis.

If we take r_1 and r_2 as the cross-sectional conductor dimensions:

$$R_i = \frac{R_s}{\mu}\left[\left|\frac{\partial L}{\partial r_1}\right| + \left|\frac{\partial L}{\partial r_2}\right|\right]. \qquad (2)$$

Let

Z_0 = characteristic impedance of line

v = phase velocity of propagation in the line.

For TEM propagation $L = Z_0/v$ and since for this mode v is independent of r_1 and r_2

$$R_i = \frac{R_s}{\mu v}\left[\left|\frac{\partial Z_0}{\partial r_1}\right| + \left|\frac{\partial Z_0}{\partial r_2}\right|\right]. \qquad (3)$$

Z_0 is determined from phase velocity and static capacity measurements as a function of the involved radii from which R_i is determined graphically from the Z_0 vs r_1 and Z_0 vs r_2 plots.

JAMES J. NAIL

Bell Telephone Labs.

Whippany, N. J.

RC Oscillators*

In this note on RC oscillators, Wray[1] pointed out that putting $R_1 = RA/s$ (Fig. 1) imposed the restriction that $R_2 = R_3$; $C_2 = C_3$, for optimum tracking.

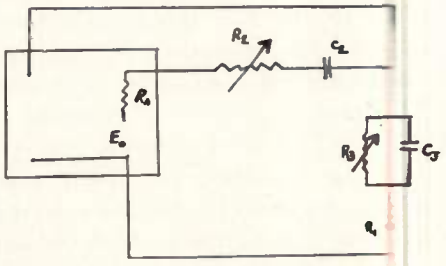

Fig. 1

However R_1 can be chosen such that the only condition imposed on the network is $R_2C_2 = R_3C_3$ and the frequency of oscillation will still be given by $\omega_0 = 1/R_2C_2$. The frequency of oscillation determined by the network in Fig. 1 is given by:

$$\omega_0^2 = \frac{1}{R_3C_2C_3\left[R_2 + R_A\frac{R_1}{R_1+R_3}\left(R_2 + R_A\frac{R_3C_3}{C_2}\right)\right]}.$$

On imposing the condition $R_2C_2 = R_3C_3$ this reduces to:

$$\omega_0^2 = \frac{R_1+R_3}{R_2^2C_2^2\left[\frac{R_AR_3}{R_2} - R_1 + R_3\right]}.$$

If R_1 is now chosen so that

$$R_1 = \frac{R_AR_3}{2R_2} = \frac{R_AC_2}{2C_3}$$

then the above reduces to $\omega_0 = 1/R_2C_2$.

The case noted by Davidson[2] is the particular case of the above when $C_2 = C_3$.

B. J. O'BRIEN

c/o Dominion Physical Laboratory,

Private Bag,

Lower Hutt,

New Zealand

* Received by the Institute, August 5, 1953.

[1] E. W. Hobson, "The Theory of Functions of a Real Variable and the Theory of Fourier's Series," Cambridge University Press, Cambridge, England, p. 66; 1907.

* Received by the Institute, Aug. 28, 1953.

[1] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., Inc., New York, N. Y., p. 285; 1943.

[2] H. A. Wheeler, "Formulas for the skin effect," PROC. I.R.E., vol. 30, pp. 412–424; Sept., 1942.

[3] "Reference Data for Radio Engineers," Federal Telephone and Radio Corp., 3rd ed., page 319; 1949.

* Received by The Institute, Oct. 12, 1953.

[1] W. J. Wray, Jr., "More on the RC oscillator," PROC. I.R.E., vol. 41, p. 801; June, 1953.

[2] J. A. B. Davidson, "A note on a precision decade oscillator," PROC. I.R.E., vol. 40, pp. 1124–1125; Sept., 1952.

Contributors

A. G. Emslie was born in Aberdeen, Scotland, on November 28, 1907. He received the M.A. degree in physics from Aberdeen University in 1928, and the Ph.D. degrees from Cornell University in 1933 and Cambridge University in 1937. He has been a research lecturer in electronics at Harvard University, professor of physics at Williams College, and member of the Radiation Laboratory staff at Massachusetts Institute of Technology.

A. G. EMSLIE

Since 1951, Dr. Emslie has been associated with the physics group at Arthur D. Little, Inc., where he has conducted numerous research studies in atomic, solid state, and theoretical physics; in particular, statistics infrared spectroscopy, quantum mechanics and information theory.

Dr. Emslie is a member of the American Physical Society.

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L. B. Felsen (S'47–A'53) was born on May 7, 1924, in Munich, Germany, and entered the United States in 1940. After arriving here, he attended the Evening School Division of the Polytechnic Institute of Brooklyn, N.Y., until 1943, at which time he entered the armed forces. During his service in the U. S. Army from 1943–1946, he was in the Army Specialized Training Program for one year and subsequently was assigned to operate and maintain electronic equipment on an Ordnance Ballistic Calibration Team which was assigned to serve in the Pacific Theater of Operations.

L. B. FELSEN

After receiving his discharge from the service, he returned to the Polytechnic Institute of Brooklyn, receiving the B.E.E., M.E.E., and D.E.E. degrees in 1948, 1949, and 1952, respectively. His work toward the M.E.E. degree was carried out on a research fellowship.

Since 1949, Dr. Felsen has been employed by the Microwave Research Institute of the Polytechnic Institute of Brooklyn, where he has been engaged in work on electromagnetic diffraction problems and microwave measurement techniques. He is presently a Research Assistant Professor and teaches at the Graduate School of the Institute.

Dr. Felsen is a member of Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

J. J. Hupert (SM'48) was born on May 6, 1910, in Tarnopol, Poland. He received the M.S. (El. Eng.) degree from the Polytechnic Institute of Warsaw, Poland, in 1933 and the Ph.D. degree from Northwestern University in 1951.

J. J. HUPERT

Dr. Hupert's considerable experience in research and development in the field of communications was acquired as head of the Transmitter Department, Bureau of Research of the State Tele- and Radio Establishment in Warsaw before the outbreak of World War II, and as temporary Senior Scientific Officer (Naval Service assignment) at the British Admiralty Signal Establishment during World War II.

Dr. Hupert has also been active in the teaching field as lecturer (radio transmitter design) at the Polish University College in London (1943–1947) and, after his arrival to U.S.A., in 1947, as a member of the faculty of the Physics Department of De Paul University in Chicago, where he conducts graduate courses and thesis work in electronics.

Since 1947, Dr. Hupert has also been associated with A.R.F. Products, Inc., River Forest, Ill., as consulting engineer and project engineer on research and development projects, mostly in the field of electronic instrumentation. He now holds the position of Director of Research of A.R.F. Products, Inc.

Dr. Hupert is a member of the American Institute of Physics and of Sigma Xi.

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F. G. Kear (A'24–M'31–SM'43–F'53) was born in Minersville, Pa., on October 18, 1903. He received the E.E. degree (cum laude) from Lehigh University in 1926, the S.M. degree in electrical engineering in 1928, and the Sc.D. in electrical engineering from Massachusetts Institute of Technology in 1933.

F. G. KEAR

Dr. Kear was associated with development of the product integraph and the differential analyzer. From 1928 to 1933 he was a physicist in the Aeronautical Radio Group at the National Bureau of Standards. For the next eight years he was Chief Engineer of the Washington Institute of Technology, in charge of development of radio aids to air navigation. Since 1941 he has been a senior partner in the consulting engineering firm of Kear and Kennedy. During 1941–1945 he headed the Radio Section, Electronics Division, Bureau of Aeronautics, U. S. Navy.

Dr. Kear is a member of the Society of Motion Picture and Television Engineers, the Association of Federal Communications Consulting Engineers, Eta Kappa Nu, Tau Beta Pi, Phi Beta Kappa, and Sigma Xi.

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E. R. Kretzmer (S'46–A'49–SM'52) was born on December 24, 1924, in M. Gladbach, Germany, and came to the United States in 1940. He received the B.S. degree in electrical engineering from Worcester Polytechnic Institute in 1944, and the S.M. and Sc.D. degrees in the same field from the Massachusetts Institute of Technology in 1946 and 1949, respectively. From 1944 to 1949 he was a member of the electrical engineering staff of MIT, engaged in communication research during the greater part of this period.

E.R.KRETZMER

In September, 1949, Dr. Kretzmer joined the technical staff of the Bell Telephone Laboratories, where he has been engaged in television research and more recently also in transistor application work. He is a member of Sigma Xi.

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C. W. Mueller (S'35–A'36–VA'39–SM'45) was born in New Athens, Ill., in 1912. He received the B.S. degree in electrical engineering from the University of Notre Dame in 1934, and the S.M. degree in electrical engineering from the Massachusetts Institute of Technology in 1936. From 1936 to 1938 Dr. Mueller was associated with the Raytheon Production Corporation, first in the engineering supervision of factory production of receiving tubes, and then in the development of gas-tube voltage regulators and cold-cathode thyratrons. In 1938 he returned to M.I.T. where he continued his studies and received the degree of Sc.D. in physics in 1942. While at M.I.T., he worked on a government contract on the development of gas-filled special-purpose tubes for counting operations. Since 1942 he has been a member of the RCA Research Laboratories, where he has been engaged in research on high-frequency receiving tubes and secondary electron emission phenomena.

C. W. MUELLER

Dr. Mueller is a member of the American Physical Society and Sigma Xi.

Contributors

G. F. Montgomery (M'48) was born in Oakmont, Pennsylvania, on May 1, 1921. He graduated from Purdue University with the B.S.E.E. degree in 1941. From 1941 to 1944, he worked at the Naval Research Laboratory and conducted research on development of radar and the radio equipment. During the period from 1944 to 1946, he served in the U. S. Army Signal Corps, working on experimental ionospheric research during part of that time. Later he was sent to operate an overseas ionospheric measurement station. Since 1946, he has been employed by the Central Radio Propagation Laboratory, National Bureau of Standards as a radio engineer in experimental radio and ionospheric research.

G. F. Montgomery

Mr. Montgomery has been actively interested in amateur radio for many years. He has held a radio license since 1935. He holds membership in the following organizations: Tau Beta Pi, Eta Kappa Nu, Sigma Pi Sigma, and Sigma Xi.

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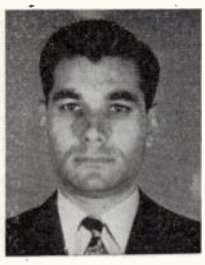
A. A. Oliner (M'47–SM'52) was born in Shanghai, China, on March 5, 1921. He received the Bachelor of Arts degree from Brooklyn College in 1941, and the Ph.D. degree in physics from Cornell University in 1946. During the time from 1941 to 1944, he was a graduate-teaching assistant in the physics department at Cornell University, and later participated as a research assistant in an Office of Scientific Research and Development project at Cornell University during the period from 1944–1945.



A. A. Oliner

Since 1946, Dr. Oliner has been with the Microwave Research Institute of the Polytechnic Institute of Brooklyn, where he has been engaged in research in the field of waveguide theory. While on a leave of absence, he spent the summer of 1952 at the Microwave Research Department of the Hughes Aircraft Company in Culver City, California. He has also taught several graduate courses in physics and electrical engineering, and is a Research Associate Professor at the Institute.

Dr. Oliner holds membership in the American Physical Society, and also in Sigma Xi.

J. I. Pankove (A'45–S'47–A'49) was born in Russia on November 23, 1922. He grew up in France where he completed his secondary education. During the time from 1942 to 1944 he attended the University of California in Berkeley, Calif., where he received his B.S. degree and also was a laboratory assistant and instrument reader.

J. I. Pankove

The following two years he served with the United States Army. He then returned to the University of California where he received his M.S. degree in 1948. Since then, he has been working as a research engineer engaged in semi-conductor research at the RCA Laboratories in Princeton, N. J.

Mr. Pankove is a member of Sigma Xi and also holds membership in the AIEE.

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J. G. Preston (S'35, A'37, VA'39) was born in Binghamton, N. Y., on January 16, 1913. He received the B.S. degree in electrical engineering (cum laude) from the Georgia Institute of Technology in 1935.

J. G. Preston

Mr. Preston entered the broadcasting field in 1930 as a transmitter engineer, and continued this association during his undergraduate years until 1937, when he became a member of the field service of the Federal Communications Commission. In 1940, he joined the FCC Broadcast Engineering Division in Washington. He was called to active military duty in 1941, and served on the staffs of the Chief Signal Officer and of the Signal Officer, Army Ground Forces. In 1946, Mr. Preston joined the American Broadcasting Company as Chief Allocations Engineer. He is now Director of Engineering Facilities and General Services for that network.

Mr. Preston is an associate member of the Association of Federal Communications Consulting Engineers, and also holds membership in Phi Kappa Phi and the Briaerean Society.

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R. M. Ryder (SM'43) was born on March 8, 1915. He received the Bachelor of Science degree in Physics from Yale University in 1937, and the Ph.D. degree in Physics in 1940. He has worked at the Bell Telephone Laboratory since 1940. He has worked in the following fields: microwave circuits, especially for low noise radar preamplifiers, wideband IF amplifiers, and fast video amplifiers. He also worked in vacuum tube development, specifically klystrons, exploratory traveling wave tubes, and close-spaced triodes for microwave radio relay. He is now working as a Transistor Development Engineer, with special interest in high-frequency and other types of transmission applications. He is a past Chairman of Committee 7.7 on Semiconductor Devices and also of Committee 7.6.3. on Noise in Electron Devices.

R. M. Ryder

He belongs to S.M.I.R.E., Yale Eng. Assn. and AAAS.

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J. M. Salzer (S'46–A'51) was born in Vienna, Austria, on September 12, 1917 He received the B.S. and M.S. degrees in electrical engineering from Case Institute of Technology in 1947, 1948, and the Sc.D. degree in electrical engineering from Massachusetts Institute of Technology in 1951.

J. M. Salzer

From 1941 to 1945, he was with the U. S. Army doing technical work on anti-aircraft computers, instruments, radio and radar sets.

In 1947 Dr. Salzer helped develop electronic controls for General Electric Co. in Schenectady, N. Y. He taught mechanics and electronics at Case (1947–48) and in 1950 circuits at M.I.T. He consulted for Commercial Television Corp., Cleveland, Ohio, on installation and servicing in 1948. From 1948 to 1951 he was Research Associate at the Digital Computer Laboratory of M.I.T. Since 1951 he has been with the Research and Development Laboratories of Hughes Aircraft Co., Culver City, Calif., in the analysis and development of digital-analog control systems. He is also Lecturer in Engineering at the University of California Los Angeles.

Dr. Salzer is a member of A.I.E.E., Association for Computing Machinery, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi

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C. B. Sharpe (A'52) was born on April 8, 1926, in Windsor, Ontario, Can. He received the B.S. degree from the University of Michigan in 1947, and the S.M. degree in electri-

Contributors

cal engineering from the Massachusetts Institute of Technology in 1949, and the Ph.D. degree in electrical engineering from the University of Michigan in 1953. While at M.I.T. he was associated with the High Voltage Research Laboratory.

C. B. SHARPE

In 1949 Mr. Sharpe returned to the University of Michigan, where he has since been associated with the Willow Run Research Center and the Electronic Defense Group.

Dr. Sharpe is now on active duty in the U. S. Navy, Bureau of Ordnance, Research and Development Division.

❖

W. R. Sittner (M'52) was born in Cape Girardeau, Mo., on August 21, 1919. He received the B.A. and M.A. degrees in physical chemistry from the University of Missouri in 1940 and 1942.

W. R. SITTNER

From 1942 until 1945, Dr. Sittner was engaged in work on magnetron-type vacuum tubes for the Western Electric Company. From 1946 until 1949, he was a research associate in the Physics Department of Northwestern University, where he worked on light sources for optical communications. He received his Ph.D. in physics there in 1949; then joined Bell Telephone Laboratories to do development work on transistors. In 1953 Dr. Sittner became the head of the Semicondutor Research Department of Hughes Aircraft Co.

Dr. Sittner is a member of the Institute of Radio Engineers, of the American Physical Society and Sigma Xi.

C. A. Swenson was born in Minneapolis, Minn., in 1923. He received the B.S. in physics from Harvard University in 1944, and did field research in the Explosives Division of the University of California. He was awarded a Henry Fellowship to study low-temperature physics at Oxford, receiving his D.Phil. in 1949.

C. A. SWENSON

During his three years at Clarendon Laboratory, Oxford, he conducted experimental research on ortho-para conversion in liquid hydrogen and on liquid-solid transformation in helium. He returned to Harvard as an instructor in physics, and set up a low-temperature laboratory to work on liquid-solid transformation in helium and research on application of high-pressure techniques to low temperature.

Since June, 1952, Dr. Swenson has been at the M.I.T. Cryogenic Laboratory.

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

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L. B. Valdes (A'46–M'52) was born in Marianao, Cuba, on June 12, 1928. He moved to New Orleans for his high school and college education and received the B.E. degree in electrical engineering from Tulane University in 1946. That fall he began graduate studies in electrical engineering at Northwestern University, where he remained for three years. He received the M.S. degree in 1947.

L. B. VALDES

Mr. Valdes then joined the transistor-development technical group at Bell Telephone Laboratories, where he has been engaged in developmental work and research projects on transistors since 1949.

Mr. Valdes holds membership in Eta Kappa Nu and Sigma Xi.

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L. Weinberg (M'52) was born on July 15, 1949, in Brooklyn, N. Y. He graduated from Brooklyn College in 1941 receiving the Bachelor of Arts degree. In 1947 he received the M.S. degree in engineering sciences and applied physics from Harvard University, and in 1951 the Sc.D. degree in electrical engineering from the Massachusetts Institute of Technology.

L. WEINBERG

During the period from 1942–1943, Dr. Weinberg worked for the Office of the Chief Signal Officer as Chief Radio Engineer in the Controlled Items Section. He left to join the Army Air Forces, where he remained until 1946, serving as a Radar and Electronics Officer, and as Technical Advisor to the American Embassy at The Hague, Netherlands. From 1947 to 1951 he was an instructor in the Department of Electrical Engineering at M.I.T. Since 1951 he has worked as a research physicist in the Systems Planning and Analysis Department of the Hughes Research and Development Laboratories in Culver City, California and also as a lecturer in engineering at the University of California, Los Angeles.

Dr. Weinberg is the West Coast Representative of the I.R.E. Professional Group on Circuit Theory, a member of the Circuits Committee, the Electronic Computers Professional Group, and the Transistor Circuitry Subcommittee, and Chairman of the Linear Lumped-Constant Passive Circuits Subcommittee. He holds membership in the Propylaea, and in Sigma Xi.

The biographies of following contributors were inadvertently omitted from the November, 1953 issue, in which their papers appeared.

E. L. Ginzton (S'39–A'40–SM'46–F'51) was born in Russia in 1915. He received the B.S. degree from the University of California in 1936, and the M.S. degree in 1937. He received the E.E. degree in 1938, and the Ph.D. degree in 1940 from Stanford University. He is now a professor of applied physics and electrical engineering at Stanford University, Stanford, California.

E. L. GINZTON

I. R. Neilsen (A'49) was born in Rulison, Colorado, on August 12, 1915. He received the B.A. degree in physics from Pacific Union College in 1936, the M.S. and Ph.D. degrees in Physics from Stanford University in 1948 and 1952, respectively.

I. R. NEILSEN

He joined the staff of Pacific Union College in 1943, where he is at present associate professor of physics.

S. Sonkin (M'49) was born in New York in 1901. He received the B.S. degree from the College of the City of New York in 1920, the degree of E.E. from Brooklyn Polytechnic Institute in 1922 and the Ph.D. degree from Columbia Univ. in 1933.

S. SONKIN

Dr. Sonkin taught at C.C.N.Y. many years. At present he is Professor in the Microwave Laboratory at Stanford University.

I·R·E News and Radio Notes

Mrs. Kinzie Honored

Mrs. Martha E. Kinzie, Secretary of the National Television System Committee, who was recently awarded the Radio Fall Meeting Plaque (inset) for her efforts in connection with the NTSC work.

Martha E. Kinzie, of the General Electric Co. and Secretary of the National Television System Committee, was honored recently at the annual IRE-RETMA Radio Fall Meeting, held in Toronto, Canada, for her untiring efforts in connection with NTSC work. On behalf of the industry's engineers, she was awarded a plaque which was inscribed as follows: "Mrs. Martha E. Kinzie, awarded for her untiring efforts on behalf of the N.T.S.C. in its work of formulation of a successful compatible standard for color television."

Mrs. Kinzie has been affiliated with the radio industry since 1920, when she worked with Dr. L. W. Chubb of the Westinghouse Electric Corp. During the period from 1920–1930 she also acted as secretary of the industry committee working on the development and design of the early radio receivers. Since 1936 she has been Secretary to Dr. W. R. G. Baker of the General Electric Co. In addition to this work she assisted L. F. C. Horle when he was Secretary of the first National Television System Committee, and she has been Secretary for the Radio Technical Planning Board and the second NTSC.

Mrs. Kinzie has been an Associate Member of the IRE since 1947, and was recently elected to the Member Grade.

B.I.R.E. 1954 Convention Announced

The 1954 Convention of the British Institution of Radio Engineers, entitled "Electronic Aids to Production," will be held in Oxford, England, from July 8 through 12. Those attending the meeting will be in residence at Christ Church, Oxford, while the technical meetings will be held in the lecture theaters of the University.

The subject of the Convention was chosen as being one of considerable general interest, and also as an opportunity to demonstrate the contribution of the radio and electronics engineer towards improving industrial production.

The General Council of the Institution has expressed the hope that engineers from the United States and other parts of the world will attend the meeting. The Council has also inquired if there are any members of the I.R.E. who might be able to present papers reviewing the progress in electronic aids to production in this country. Anyone interested in information about attending the Convention or in presenting a paper should contact F. W. Sharp, Publications Officer, British Institution of Radio Engineers, 9 Bedford Square, London, W.C. 1, England.

IRE National Convention Set for March 22–25

Plans for the 1954 IRE National Convention, to be held at the Waldorf-Astoria Hotel and the Kingsbridge Armory in New York City from March 22–25, are now nearing completion.

Fifty-one technical sessions, sponsored by the 21 IRE Professional Groups, are scheduled to be held over the four-day period. A total of 244 papers will be presented. Two special symposia will be held on Tuesday evening, March 23: "Engineering Based on Biological Design," sponsored by the Professional Group on Medical Electronics, and the Audio Seminar on "High Fidelity in Audio Engineering." The 1954 Radio Engineering Show, held at the Kingsbridge Armory, will include over 600 exhibits, a 50 per cent increase over last year's exhibit.

The Annual Meeting of the Institute is scheduled to be held on Monday morning, March 22. John D. Ryder, Head of the Electrical Engineering Department at the University of Illinois, will be the principal speaker.

Two social activities will be held during the Convention—the Annual "Get Together" Cocktail Party on Monday evening March 22, in the Grand Ballroom of the Waldorf-Astoria, and the Annual Banquet on Wednesday evening, March 24. At the Banquet Dr. Alfred N. Goldsmith, cofounder and Editor Emeritus of the Institute, will be presented with the Founders' Award, and he will deliver the keynote address, "IRE—Past and Future." The other Annual IRE Awards will also be presented at this time, and those members awarded the Fellow Grade in 1954 will be introduced. Dr. Maurice J. H. Ponte, Vice-President of the Institute for 1954, will be the spokesman for the newly-elected Fellows.

The full program of the Convention, including 100-word abstracts of all technical papers and a list of exhibitors and products, will be published in the March issue of the PROCEEDINGS OF THE I.R.E.

Calendar of COMING EVENTS

IRE-AIEE Conference on Transistor Circuits, Philadelphia, Pa., February 18–19

Society of Women Engineers, National Convention, Mayflower Hotel, Washington, D. C., March 5–7

1954 IRE National Convention, Waldorf-Astoria Hotel and Kingsbridge Armory, New York, N. Y., March 22–25

The Association of Radioelectricians, International Sound Recording Congress, Paris, France, April 5–10

Stanford Research Institute-United State Air Force Symposium on Automatic Production of Electronic Equipment, Fairmont Hotel, San Francisco, Calif., April 19–20

Electrochemical Society Semiconductor Symposium, La Salle Hotel, Chicago, Ill., May 2–6

Society of Motion Picture & TV Engineers, 75th Semi-Annual Convention, Hotel Statler, Washington, D. C. May 2–7

IRE-AIEE-RETMA Electronic Components Symposium, Washington, D. C., May 4–6

IRE Seventh Region Technical Conference, Multnomah Hotel, Portland, Ore., May 5–7

IRE New England Radio Engineering Meeting (NEREM), Sheraton Plaza Hotel, Boston, Mass., May 7–8

IRE National Airborne Electronics Conference, Biltmore Hotel, Dayton, Ohio, May 10–12

IRE-AIEE-IAS-ISA National Telemetering Conference, Morrison Hotel, Chicago, Ill., May 24–26

IRE Symposium on Global Communications, Washington, D. C., June 10–12

IAS Annual Summer Meeting, IAS Building, Los Angeles, Calif., June 21–24

British Institute of Radio Engineers, 1954 Convention, Christ Church, Oxford, England, July 8–12

IAS Turbine-Powered Air Transportation Meeting, Seattle, Wash., August 9–11

Society of Motion Picture & TV Engineers Seventy-sixth Semi-Annual Convention, Ambassador Hotel, Los Angeles, Calif., October 17–22

IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., October 18–20

I·R·E News and Radio Notes

San Francisco Science Center Proves Successful

In April, 1953, the directors of the Co-operative Research Foundation approved a program to establish and operate a number of international science centers for the service of foreign scientists and engineers visiting the major centers of scientific activity in the United States. The primary objective of the program is to stimulate direct working relationships between foreign scientists and their colleagues in the regions to be served by the centers.

The centers, which are to be established with the co-operation of state or local schools of science, will provide facilities for visiting scientists from overseas, especially those whose visits to the United States are sponsored by private interests, and will offer such assistance to these men as other institutions of the area may not be equipped to give. The centers will also serve as a co-ordinating agency for international scientific activities in the region.

The plan for the establishment of the centers was developed jointly by the Office of International Relations of the National Academy of Sciences—National Research Council, and the Co-operative Research Foundation. Since the Policy Committee of the O.I.R. felt that the role of the Academy should be that of a co-operating, rather than an operating agency, the Co-operative Research Foundation offered to assume the responsibility for the establishment of the centers, and to function as the common co-ordinator. The several agencies felt that this plan would assure the greatest possible conjunction of the activities of all centers, and the present program was then evolved.

The first of the international science centers was established several months ago in San Francisco, Calif., with the co-operation of the California Academy of Sciences. It has already begun to have visitors, and has received enthusiastic support from the scientific organizations in the San Francisco area.

Foreign members of the I.R.E. who may be traveling in the United States will be interested to know that the services of the San Francisco Center may be received by telephoning BAyview 1-7314 on arrival in San Francisco. The Center is located at the Morrison Planetarium of the California Academy of Sciences, Golden Gate Park, and is open from 9:00 A.M. to 5:00 P.M. every Monday through Saturday.

Semiconductor Symposium Planned

The Semiconductor Group of the Electronics Division of the Electrochemical Society is sponsoring a symposium for the Chicago Meeting of the Society, to be held May 2–6, at the La Salle Hotel, Chicago, Ill. The symposium will feature intermetallic compound semi-conductors such as the antimonides, arsenides, and phosphides. Sessions are also planned on germanium, silicon, and other semi-conductors.

The papers under consideration include new and old semi-conductors, their electrical, chemical, and metallurgical properties, and preparation and fabrication methods; closely associated geometrical and electrical requirements on imperfections in semiconductors, and mechanical and chemical combinations of semiconductors with other materials to achieve a particular solid-state electrical device; techniques of producing conducting films and compositions, and devices made from them, as well as analytical properties and structure of semiconductors and semiconductor devices.

1954 National Telemetering Conference

The 1954 National Telemetering Conference under the joint sponsorship of the I.R.E., the American Institute of Electrical Engineers, the Institute of the Aeronautical Sciences, and the Instrument Society of America will be held in Chicago, Ill., on May 24–26.

This year's meeting promises to provide subject matter of wider scope and deeper interest as a result of a considerable amount of missile telemetering and remote control information just recently developed.

Mr. W. J. Mayo-Wells of the Applied Physics Laboratory of Johns Hopkins University is Chairman of this year's meeting. The meetings and exhibit will be held and presented at the Hotel Morrison in Chicago.

Are You Eligible for a Higher Grade of IRE Membership?

Every member of the IRE should be registered as a member in the highest grade for which he is qualified. If you feel you are eligible for a higher grade, contact your local Section Officers, or write to the IRE Headquarters, for a transfer form and full details, at the following address: The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

Sound Recording Congress Announced

The Association of Radioelectricians plans to hold an International Congress from April 5–10, on the general subject of Soundtrack Recording Processes and Their Extension to Other Fields of Application.

It is planned to divide the meeting into four sections: Mechanical Recording, Photographic Recording, Magnetic Recording, and Development and Standardization Problems. In certain cases several sections may hold joint sessions. The registration fees, fixed at 1500 francs, will entitle each participating member to receive a brief analysis of the papers at the opening of the Convention and a copy of the minutes of the meetings, and to participate in the visits and receptions provided for in the program. The program will also include a public exhibition of equipment. The conditions for participation in this exhibition will be determined later.

Anyone who wishes to obtain further information concerning this Congress should contact the Société des Radioelectriciens, 10–14 Avenue Pierre-Labrousse, Malakoff (Seine).

"Aviation Electronics Days" Scheduled

"Aviation Electronics Days," consisting of a special luncheon and three full technical sessions, have been scheduled for March 22–23, during the IRE National Convention in New York City, to commemorate the 50th anniversary of powered flight. Plans for the event, in which the Institute of Aeronautical Sciences will participate, were announced by Dr. K. C. Black, chairman of the Professional Group on Aeronautical and Navigational Electronics.

Fifteen technical papers, most of them in the field of aviation, are scheduled for presentation, beginning on Monday afternoon, March 22. The Tuesday luncheon and the sessions which precede and follow it, will all be held at the Hotel Shelton, 49th Street and Lexington Avenue, New York City. Details on the luncheon speaker and how to obtain tickets will be announced shortly.

Color Television Symposium

The Philadelphia Section of the I.R.E. announces a Color Television Symposium which will be held on February 11, 18, 25, March 11, 19, and April 1. An integrated program has been prepared to present various facets of this timely subject. Twelve nationally-known speakers have been scheduled, including three Chairmen of NTSC Panels. For further information about the Symposium, write the Institute of Radio Engineers, P.O. Box 2113, Philadelphia 3, Pa.

Cumulative Index to I.R.E. Publications, 1948–1953

A cumulative index of Proceedings of the I.R.E., Transactions of the I.R.E. Professional Groups, and the Convention Record of the I.R.E. for the period 1948–1953 will be available shortly. Copies may be purchased at $2.50 each from the Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

Cumulative indexes for 1913–1942 and for 1943–1947 are still available at $1.50 per copy.

I-R-E News and Radio Notes

N.Y.U. Symposium Planned

New York University has announced its First Annual Symposium on Recent Advances in Science: Physics and Applied Mathematics for the Scientist and Engineer, which will be held each Monday from 7–9:30 P.M. from February 8 to May 24. Registration may be by mail or in person through February 8. The tuition for the full course is $47.00.

The Symposium will offer the following lectures: February 8, Richard Courant, New York University, "Methods of Applied Mathematics"; February 15, Claude Shannon, Bell Telephone Laboratories, "Information Theory"; March 1, John Von Neumann, Institute for Advanced Study, "Computing Machines"; March 8, P. M. Morse of Massachusetts Institute of Technology, "Methods of Operations Research"; March 15, I. I. Rabi, Columbia University, "Atomic Structure"; March 22, C. H. Townes, Columbia University, "Microwave Spectroscopy"; March 29, H. A. Bethe, Cornell University, "Nuclear Structure and Transmutation"; April 5, V. F. Weisskopf, Massachusetts Institute of Technology, "Elementary Particles"; April 12, Leland J. Haworth, Brookhaven National Laboratory, "Electronuclear Machines"; April 19, Norman F. Ramsey, Harvard University, "Neutron Physics"; April 26, Frederick Seitz, University of Illinois, "Solid State Physics"; May 3, William Shockley, Bell Telephone Laboratories, "Semiconductor Electronics"; May 10, R. M. Bozarth, Bell Laboratories, "Magnetism"; May 17, F. G. Brickwedde, National Bureau of Standards, "Cryogenics"; and May 24, Edward U. Condon, Corning Glass Works, "Physics and the Engineer."

Further information may be obtained by writing to Sidney G. Roth, Division of General Education, New York University, Washington Square, New York 3, N. Y.

Technical Committee Notes

The **Circuits** Committee convened on December 4th under the chairmanship of W. R. Bennett. Professor Lynch distributed copies of a report of a meeting of Subcommittee 4.7 held November 30th. This meeting made notable progress toward breaking the deadlock which had developed in defining basic terms such as "feedback loop," "loop transmission," and "stability." The proposed definitions were agreeably received by the Committee, subject to clarification of certain terms used in phrasing them. One of these was "unilateral device" which did not convey a satisfactory meaning to the Committee. The Committee proceeded to discuss "unilateral" and "feedback." As a result of their discussion a number of definitions were approved. Professor Lynch volunteered to compile an up-to-date list of the Subcommittee 4.7 terms and their definitions.

On November 13th the **Electron Devices** Committee met under the chairmanship of G. D. O'Neill. The Annual Review material was reported by T. J. Henry, Chairman, Subcommittee 7.1, to be virtually complete. Chairman E. O. Johnson, Subcommittee 7.3, reported that Task Group 7.3.1 has completed 80 per cent of its planned work on definitions and has begun work on methods of test. G. A. Espersen, Chairman, Subcommittee 7.5 was not present. A written report of the subcommittee work was discussed. P. A. Redhead, Chairman, Subcommittee 7.6 (Physical Electronics) reported on the formative work which has been necessary in this new subcommittee prior to the assumption of specific standards responsibilities. R. M. Ryder, reported for A. E. Anderson, Chairman, Subcommittee 7.7, that a set of definitions had been readied for submission to the main committee. A major part of this work was done before activation of the Subcommittee on Physical Electronics. It was requested that the proposed definitions be submitted immediately to Mr. Redhead for consideration by Subcommittee 7.6 with the hope that the material could be made available for processing by the main committee at its January meeting. W. J. Pietenpol has resigned as Chairman of Subcommittee 7.7.2 and has been replaced by C. L. Rouault of the General Electric Company, Schenectady, New York. R. S. Burnap reported, after scrutinizing the comments returned with the Phototube Definitions from the Standards Committee, that questions were raised which will require discussion with Chairman Janes of Subcommittee 7.4.

Under the chairmanship of Henry Burkhard the **Facsimile Committee** convened on November 13th. A number of terms were reviewed and action was taken by the Committee.

On November 12th the **Feedback Control Systems** Committee convened under J. E. Ward the acting Chairman. A memorandum prepared by G. A. Biernson "Recommendations Concerning Definition of Basic Feedback Control Loop" was distributed to the Committee. Mr. Biernson's memorandum was discussed by the Committee and compared with the report of the Feedback Control Systems Subcommittee on Terminology dated November 2, 1953. Chairman Ward suggested that the Committee defer discussion of their report "Recommended Symbols and Terminology for the Basic Feedback Control Systems" until the December meeting and instead discuss the report of the Subcommittee on Terminology. Since the Committee's report was to be based on the subcommittee's terminology, it seemed desirable to obtain general agreement between the Committee and the Subcommittee on the subject of terminology. The comments of the Committee members on the various sections of the Subcommittee's report was discussed.

The **Navigation Aids** Committee convened on November 23rd under the chairmanship of P. C. Sandretto. At the October 26th meeting of the Navigation Aids Committee it had been decided that the remaining Navigation terms should be divided into five equal parts and that Messrs. Metz, Mimno, Palmer, Sherer and Gray would each edit one of the five parts. Each of these people gave a report on their assigned list of definitions, and action was taken by the Committee on each of these terms.

Due to the absence of the Chairman, Dr. Friend, the **Sound Recording and Reproducing** Committee met under M. S. Corrington, the Vice Chairman. The IRE Standard of the proposed American Standard "Method for Determining Flutter Content of Sound Recording and Reproducing, Z571.1.6" has been approved by the Standards Committee. Dr. Peterson will exchange definitions and procedures for measurement of distortion and intermodulation with Subcommittee 3.1. They have made many helpful suggestions for changes and may be invited to co-sponsor resulting standards. Mr. Fraser, reporting for Subcommittee 19.3 on Optical Recording and Reproducing, indicated that the SMPTE Committee on Nomenclature will work with his committee. They have gathered definitions of terms and compared them with the master list of the IRE. A preliminary copy of the Proposed Standards on Intermodulation Distortion: Definitions and Procedures for Measurement dated November 16, 1953 was submitted by Dr. Peterson. They were studied for general comments and were referred to the Committee with some recommendations.

Professional Group News

NEW CHAPTERS APPROVED

The following Chapters were approved by the Executive Committee of the I.R.E. at their meeting on December 8: The Washington D. C. Section Chapter of the Professional Group on Information Theory; the Philadelphia Section Chapter of the Professional Group on Circuit Theory; the Boston Section and the Detroit Section Chapters of the Professional Group on Electronic Computers; and the Los Alamos Section Chapter of the Professional Group on Nuclear Science.

Component Parts

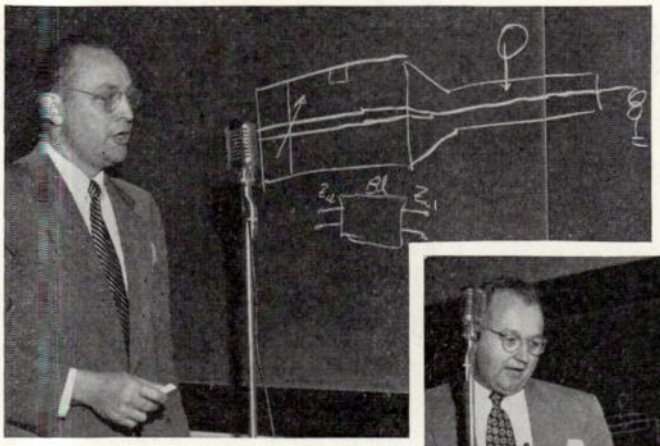
The Professional Group on Component Parts has announced its schedule of activities for early 1954. There are four major programs on which the Group is concentrating its efforts. The first is an issue of Professional Group Transactions, which will be published early in 1954. The Group is also sponsoring an Electronic Components Symposium with the AIEE and RETMA, which will be held in Washington, D. C. on May 4–6. (See the January issue of the Proc. I.R.E. for story.) A third program is emphasizing a scholarship proposal to initiate action in colleges and universities which will make engineering electronic graduates more aware of details and parts applications. The final project, and perhaps one of the most important, is the project to establish more Component Parts Chapters throughout the country. The Group, which is the newest of the IRE Professional Groups, is now under the chairmanship of F. A. Paul, and has a total membership of 284, with Chapters in Los Angeles, Calif., and Dayton, Ohio.

3rd ANNUAL
BROADCAST SYMPOSIUM

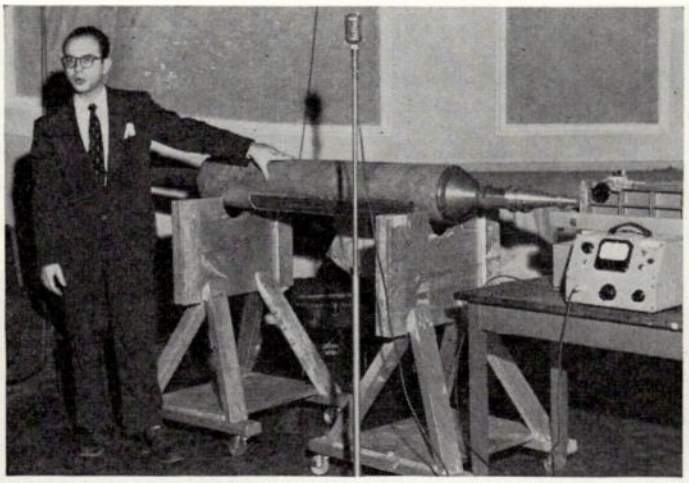
Held Recently at Franklin Institute, Philadelphia, Pa.,
and Sponsored by the Professional Group on
Broadcast Transmission Systems

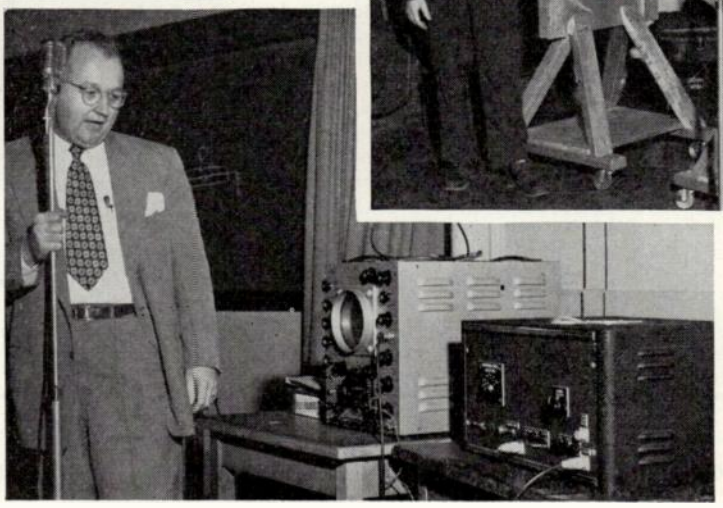
Above—Lewis Winner, Chairman of the Professional Group on Broadcast Transmission Systems, who also served as chairman of the Symposium.

Below—Owen O. Fiet, RCA, who discussed uhf directional antennas at the Symposium.



Right—William Behrend, RCA, with step generator and scope set-up used in color to waveform analysis work. Operation and use of equipment were discussed by George H. Brown, RCA Laboratories, and R. E. Shelby, in their paper on compatible color television.





Above—D. C. Stork, RCA, shown at the Broadcast Symposium with actual test set-up, used for measuring impedance and velocity in the design of a particular high gain directional array for uhf television; described in a paper by O. O. Fiet and E. H. Shively, both of RCA, on directional antennas for uhf television.

Below—John B. Grund, KG2XDU and KG2XEL, owned and operated by Sylvania, who discussed uhf to satellite operation.

Left—Robert E. Shelby, NBC, (inset) shown below discussing compatible color television and its relationship to the broadcaster, which followed a report on the subject by George H. Brown of RCA Laboratories.

I-R-E People

For his outstanding paper, "Compensation of Feedback-Control Systems," **Dr. George Cheney Newton, Jr.** (M'47), Associate Director of the Servomechanisms Laboratory, Massachusetts Institute of Technology, has been awarded the Louis E. Levy Medal by the Franklin Institute of Pennsylvania. The article appeared in the October and November, 1952, issues of the *Journal of The Franklin Institute.*

G. C. NEWTON, JR.

The presentation of the medal was made on the occasion of the Institute's annual Medal Day ceremonies.

Born in Milwaukee, Wisconsin, in 1919, Dr. Newton was graduated in 1941 from the Massachusetts Institute of Technology with a B.S. degree, and received his Sc.D. in Electrical Engineering in 1950. In that Institute he was appointed Instructor in Electrical Engineering, 1946; Assistant Professor, 1949; and Associate Professor in 1952, the position he now holds. In addition to teaching Dr. Newton carries on a consulting practice in the field of automatic control. He also has served as an engineer of the York Safe and Lock Co., and as project engineer of the Sperry Gyroscope Co.

He is a member of The Franklin Institute, and the American Institute of Electrical Engineers.

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Kerim Onder (J'34–A'35–M'45–SM'45) has been appointed engineer-in-charge of the newly established Circuit Research Laboratory, 617 West 113 Street, New York, N. Y. The activities of the laboratory are devoted to research and development in radio, television, audio, and electronics.

Mr. Onder attended the University of Istanbul and the American Collegiate Institute, and received his Master's degree in Electrical Engineering from Stanford University. For the past fifteen years he has been affiliated with several companies, including Marconi's Wireless Telegraph Company, Ltd., of England, and the RCA Laboratories in Princeton, N. J.

He is a member of the American Standards Association.

❖

William J. Morlock (A'43–SM'46) has been recently appointed general manager of the new Commercial Equipment Department of the Electronics Division of General Electric Company.

W. J. MORLOCK

Mr. Morlock is a native of McKeesport, Pa., and a graduate of Ohio State University. Prior to joining General Electric, he was engaged in the development of interior communication and sound equipment for the U. S. Navy and other government agencies. During the war he was a committee member of the Office of Scientific Research and Development. For several years before the war he was manager of distributive products engineering for the Radio Corporation of America.

He joined G. E. at Syracuse in January, 1948, and served in several executive engineering positions until June, 1950, when he was appointed assistant manager of the Commercial and Government Equipment Department. He was named general manager for commercial equipment activities of the department in 1951. Since then, these activities have grown through their designation as a sub-department last year, until the recent announcement making them a separate product department.

Mr. Morlock is a member of the Society of Motion Picture Engineers, Pi Tau Pi Sigma, Theta Kappa Phi, and the Radio Oldtimers Association.

Lt. Commander John E. Fredricks (A'15–VA'39), USN (Ret.), recently passed away.

A native of Brush Prairie, Wash., in 1905 he enlisted in the Navy. He was retired in 1946 after 41 years of continuous service. During that period he held various technical and administrative positions in the radio and electronics field at Naval Stations and at sea. During World War II he was a Communications Officer for the Pacific Fleet, and from 1943 to 1945 he was the Radar Installations Officer at the U. S. Naval Base in San Diego, Calif. From 1945 to 1946 Mr. Fredricks participated in the storage, disposal, and repair of electronics equipment during the demobilization of the Navy.

Mr. Fredricks was a leader in the campaign to perfect and install radar on Naval Ships during the early development of that instrument. He also participated in the applications of electronically-controlled gun-laying and the development of adequate intercommunication systems on Naval Ships.

Books

Principles of Transistor Circuits, ed. Richard F. Shea

Published (1953) by the John Wiley & Sons, Inc., 440 Fourth Ave., New York, N. Y. 508 pages+9-page index+18 pages Appendixes+xxx pages. 445 figures. 6×9¼. $11.00.

Mr. Shea is a member of the engineering staff of the Electronics Laboratory, General Electric Co., Syracuse, N. Y.

This book is a new volume written by nine members of the General Electric Electronics Laboratory. It is intended for the practicing engineer and graduate student. The treatment of the subject matter seems to be more in the line of the practicing engineer than student, although the latter half of the book may well be fitted into graduate studies. The more detailed analyses can then be assimilated as the need arises.

Some material is included on almost all the applicative phases of transistors. A brief but adequate treatment of semiconductor principles constitutes the first chapter. The authors have attempted and succeeded to a reasonable degree to make each chapter self-explanatory. Problems of notation and definitions have been simplified by tabulations of symbols, definitions of terms, and tabulation of phrases and device characteristics. In addition to references at the end of each chapter, there is an extensive chronological bibliography at the end of the book covering the period from 1945 to the end of 1952. Problems without answers are given at the end of each chapter.

Although the authors make it amply clear that transistors are power-operated devices whose operating parameters will in general have both a resistive and reactive component, the conditions for conjugate impedance match are not considered. Conditions for iterative impedance match have also been omitted. Too great an emphasis is placed upon the grounded base (emitter input) connection. Thus, the consideration of power amplifiers is largely devoted to the grounded base circuit which is the least used of the three possible circuits. Complementary symmetry circuits are treated in one brief paragraph. Too little attention is also given to the driver for power amplifiers.

The authors have done a commendable job in embracing the various aspects of the subject in view of the limited amount of published information available to them at the time the material was prepared. It is the opinion of the reviewer that all those giving serious thought to the application of transistors will want to familiarize themselves with the contents of this book.

L. J. GIACOLETTO
RCA Laboratories Division
Princeton, N. J.

Books

Ultra High Frequency Propagation by Henry R. Reed and Carl M. Russell

Published (1953) by John Wiley & Sons, Inc., 440 Fourth Ave., New York, N. Y. 554 pages+8-page index+xiv pages. 14 figures. 6×9¼. $9.50.

The primary purpose of this book, according to a statement in the preface, is to present current information on UHF radio wave propagation. The authors have also attempted to integrate propagation data with system design and they have done a rather good job in this direction. Much of the book deals with aircraft-to-aircraft systems and aircraft-to-ground systems, perhaps because the authors' background and experiences have been in this region.

The book suffers from the lack of references to existing literature. For example, in their discussion of circularly-polarized antenna systems, the authors state on page 324 "In addition to the crossed-dipole antenna, there are available a number of other types of antennas that produce circularly polarized radiation." Since no further hint is given as to the nature of these antennas, references to the existing literature might prove helpful to the reader who wishes to pursue the subject.

The authors' word descriptions of the physics of propagation is the best part of the book, while the weakest treatment is in the field of antennas. In their treatment of colinear arrays, they neglect entirely the impedance change due to mutual coupling between elements, and as a result conclude on page 259 that the voltage gain of a two-element colinear array is $\sqrt{2}$ and fail to point out that the gain is a function of separation between the elements. When two half-wave dipoles are placed in line and the gap is small, the voltage gain is 1.21 instead of $\sqrt{2}$. Likewise, on page 262, they prove that the gain of an n-element colinear array is $\sqrt{n}$, while a value of $0.83\sqrt{n}$ is a more generally accepted figure.

In the preface, the authors frankly state that the book is not written for any single level of experience. They evidently hoped to cover a wide level, for on page 466, they say

$$\epsilon = 400/159.9 = 2.5° = 400/9162 = 0.0437 \text{ radian},$$

a manipulation that would hardly tax the capabilities of the average high-school graduate, while on page 66, they produce the solution of the vector potential from the attendant differential equation by "heuristic" reasoning. To this unsophisticated reviewer, this procedure is equivalent to waving a wand—"Behold!"

The authors devote at least sixteen pages to a method of calculating the UHF field strength in the diffraction region around a smooth spherical earth, using an adaptation from K. A. Norton's original paper. This adaptation includes a lengthy piece of calculation which involves tucking values of dielectric constant and conductivity into a vast array of equations. Several years ago, this reviewer discovered a well-kept secret, namely, that for a wide range of dielectric constants and earth conductivities, one gets the same value of field intensity since the parameters cancel out as one proceeds through the maze. This observation enables one to make the calculation in a relatively simple fashion, using greatly simplified formulas.

Using the following notation:

h = height of the transmitting antenna, feet,
a = height of the receiving antenna, feet,
f = frequency, megacycles,
d = distance between transmitting and receiving antennas, statute miles,
E = field intensity, microvolts per meter for one kilowatt of effective radiated power,

one calculates a reference distance d_0 which is in the diffraction region.

$$d_0(\text{miles}) = 216/f^{1/3}$$

Now, to calculate the field intensity at this distance, we first establish two parameters.

$$h_1 = 8.56\times10^{-5}\, h.f^{2/3}$$

and

$$h_2 = 8.56\times10^{-5}\, a.f^{2/3}.$$

Then two functions ($F(h_1)$ and $F(h_2)$ are found by reading from the curve marked "$K\geq0.01$" in Fig. 5–26 in the Reed-Russell book.

For the reader who does not have access to this book and who may wish to plot his own curve, the following table is provided.

h	$F(h)$
0.01	0.04
0.02	0.08
0.05	0.2
0.1	0.41
0.2	0.82
0.5	2.2
1.0	5.3
2.0	16.4
5.0	137.
10.0	1,410.
20.0	39,000.
50.0	29,000,000.
100.0	51,000,000,000.

Then at d_0 miles, the field intensity for one kilowatt of effective radiated power is

$$E_0(\text{microvolts per meter}) = 13.03\, f^{1/3}\cdot F(h_1)\cdot F(h_2)$$

To find the field intensity E at other distances in the diffraction region, one uses the following table.

d/d_0	E/E_0
0.5	10.5
0.75	3.05
1.0	1.0
1.25	0.35
1.5	0.125
2.0	0.0175
2.5	0.0025
3.0	0.00035

These simplified procedures have been found to be in excellent agreement with the more complex manipulations found in Norton's papers and in this book and, one might add, any disparity between the two methods becomes insignificant when one recalls that neither method agrees very well with actual measurements of waves propagated over the surface of a world nodulated with buildings and trees.

GEORGE H. BROWN
RCA Laboratories
Princeton, N. J.

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Television Fundamentals by Kenneth Fowler and Harold B. Lippert

Published (1953) by the McGraw-Hill Book Company, Inc., 330 West 42nd St., New York 36, N. Y. 520 pages+4-page index+vi pages. 425 figures. 6¼×9¼. $7.00.

Mr. Fowler is the Supervisor of Field Engineering, Radio and Television Department, Electronics Division, of the General Electric Company. Mr. Lippert is the Supervisor of Product Service, Commercial and Government Equipment Department, Electronics Division of the General Electric Company.

A non-engineering presentation of the basic principles of installing and servicing television receivers. Each element of the receiver system, from antenna to picture tube, is taken up in turn, without mathematical or heavy theoretical treatments. It includes practical installation rules for receivers and antennas, including making antennas for various reception conditions.

After a brief description of the present television details (black and white) and of cathode ray tube principles, the book reviews the reasons for each function of typical receivers, and the equipment details whereby those functions are accomplished.

These descriptions are concise and accurate, and cover many of the common methods used in present receivers. They are not specific as to circuit values, relying on previous knowledge of radio circuits on the part of the readers. The author's desires are to give the radio technician enough fundamental knowledge to analyze troubles from either test instrument measurements or by an inspection of the resulting pictures and sound. About one-fourth of the book covers these subjects.

This book is not particularly suited to engineers. Most of the chapters are best suited as a course of study either in a trade school or for home study use. The final two chapters provide reference data for getting service men started until the symptoms accompanying troubles are learned.

The book is relatively up to date as far as VHF monochrome receivers are concerned.

R. R. BATCHER
Radio-Electronics-Television Manufacturers Association
New York, N. Y.

Books

Basic Electronic Test Instruments by Rufus P. Turner

Published (1953) by Rinehart Book Co. Inc., 232 Madison Avenue, New York 16, N. Y. 245 pages+6-page-index+xiv pages. 172 figures. 6×9. $4.00.

Intended as a general survey for service technicians, experimenters and amateurs, this book discusses in simple, non-mathematical style common types of measurements and measuring devices peculiar to communications and electronics. The treatment is descriptive, rather than fundamental, and is not concerned with analyzing principles or tracing historical development. The bibliography is "practical" rather than "professional," and the circuit schematics and photographs used as illustrations are directed to specific instruments and circuits in current use. Considerable emphasis is laid upon the test instruments now available in kit form, and the descriptions of how to build equipment found in the author's earlier *Radio Test Instruments* are correspondingly minimized. In a field that is rapidly becoming complex the concept of what is basic changes rapidly, and it is noteworthy today that the omission of any discussion of uhf measurements and electronic counters strikes one as disappointing, even in an elementary manual. The book offers many pointers for practical everyday use, but a more philosophical treatment would seem desirable for students and others planning to acquire a thorough grasp of the subject.

D. B. Sinclair
General Radio Company
Cambridge, Mass.

Fields and Waves in Modern Radio by Simon Ramo and John R. Whinnery

Published (1953) by John Wiley & Sons, Inc., 440 Fourth Ave., New York, N. Y., and Chapman & Hall Ltd., London, England. 565 pages+2-page appendix +8-page index+vii pages. 6×9¼. $8.75.

This is the second edition of this test which first appeared in 1944. It differs only slightly from the first edition, but the changes go far toward removing most criticisms levelled at the original work. The principal changes include the use of the rational mks system of units throughout the text, rather than in part; and the inclusion of a new chapter on microwave networks. A number of miscellaneous minor changes have been made, including new examples, the addition of the Smith chart in the treatment of transmission lines, several new aspects of propagating waves, the added discussion of horns, slot antennas. receiving antennas, and additional material on arrays.

The book opens with several alternative solutions for free and forced oscillations. This is followed by a clear discussion of attenuation, transmission, reflections, and standing waves. A clear picture of group velocity is given.

The equations of electrostatics and magnetostatics in Chapter 2 serve to introduce the basic properties of vector analysis. The scalar product and the divergence is presented through Gauss's law. The vector cross product is introduced through Ampere's law. The concept of potential, potential difference, gradient, Laplace's and Poisson's equations, are clearly discussed. The curl is introduced through a discussion of magnetomotance. The scalar and vector potential, and the integrals for them, the energy in the electrostatic and in the electromagnetic fields, Stoke's theorem, orthogonal coordinates, are all well-presented.

Chapter 3 contains a resume of potential theory. It includes, with examples, many of the techniques used in the solution to static field problems; graphical field mapping, method of conformal transformations, and a resume of the important concepts of complex function theory, the product solution, and problems solved in terms of rectangular harmonics, Bessel function, Hankel functions, spherical harmonics.

Maxwell's equations are discussed in Chapter 4, with a discussion of the propagation equations for fields, and electromagnetic potentials. The integral expressions for the retarded scalar and vector potentials are clearly presented. A clear distinction is made between the voltage between two points, and the potential difference. All too frequently this distinction is lost. This receives further attention in the next chapter.

The circuit concepts and their derivation from the field equations receive attention in Chapter 5. The generalizations include the effects of displacement current; also the modifications in the ideas of inductance and capacitances to include allowance for the propagation time between circuit elements. This is followed in Chapter 6 by a detailed discussion of skin effect, self- and mutual inductance, self- and mutual capacitances.

This reviewer feels it necessary to note, in agreement with a reviewer of the first edition, that the first six chapters contain many examples of expert exposition, and originality of presentation. The remainder of the text, while competently accomplished, is generally ordinary in presentation.

Chapter 7 discusses the propagation and reflection of electromagnetic waves. Poynting's vector is introduced; plane waves, polarization, the concept of wave impedance, receive attention. The reflection of waves from conductors and dielectrics receives considerable discussion. The features of waves in imperfect conductors and dielectrics are also considered.

Chapter 8 is devoted to guided electromagnetic waves. The general aspects of field guidance are well discussed in terms of parallel plane guide, with good discussions of the transverse magnetic and electric waves between planes. A general analysis of guided waves follows, with the propagation equations for guides of rectangular and cylindrical shape being obtained.

The next chapter discussed the characteristics of common wave guides, and transmissions lines. Coaxial lines, rectangular wave guides, circular wave guides, dielectric, radial, conical, ridge guides, helix and other slow-wave structures, receive attention. A number of charts giving detailed results for many of these guide properties and characteristics are included.

Resonant cavities are discussed in Chapter 10. Much is left to be desired here, since attention is confined principally to the general problem of the Q of simple resonators. Since resonators are generally used as elements of klystrons, some discussion should have been included of the shunt resistance between poles of the resonator. Also more attention might have been given to the resonant frequencies of resonators.

Microwave networks, the title of the next chapter and the major new addition to the book, contains a discussion of the network formulation of microwave networks, which is then followed by a review of the standards properties of networks, principally 2-, 4-, and 6-terminal networks. The adaptation to microwave networks is very interesting. A very inadequate treatment of the analytical techniques used in microwave networks follows.

The final chapter discusses radiation with details of the dipole antenna, the long antenna, the half-wave dipole, and others. An interesting calculation of the radiation resistance by the induced emf method is given. Arrays of elements are considered in some detail; two dipoles, the rhombic linear arrays, supergain. The formulation of the radiation problem in terms of equivalent electric and magnetic surface currents is adequate to illustrate the methods used. The antenna impedance problem is presented, but is not adequately explored. Receiving antennas and Reciprocity appears adequate for introducing these topics.

On the whole, the book maintains good balance between subject matter and technical detail. It will undoubtedly take its place as one of the important texts for senior and first-year graduate students, as did its predecessor.

Samuel Seely
Syracuse University
Syracuse, New York

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

Note: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

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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.

A new section has been introduced covering the general technique of electromagnetic waves, oscillations and pulses (i.e. transmission lines and circuits), as distinct from specific applications to telecommunications. The new section is numbered 621.37, with subdivisions. Full details of the new classification, and of the numbers which become obsolete as a result of its introduction, are given in PE Note 535, obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1, England.

Section 621.396.67, dealing with Antennas, has been modified and expanded; details of the new classifications are given in PE Note 519.

Section 621.396.96, with subdivisions, has been introduced to cover Radar; details of the new classifications are given in PE Note 518.

New Subject Section

A section headed Automatic Computers has been introduced.

The Index to the Abstracts and References published in the PROC. I.R.E. from February 1953 through January 1954 is published by *Wireless Engineer* and included in the March 1954 issue of that journal. Copies of this issue may be purchased for $1 (including postage) from the Institute of Radio Engineers, 1 East 79th Street, New York 21, N.Y. As supplies are limited, the publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

ACOUSTICS AND AUDIO FREQUENCIES

534.231 1

Acoustic Radiation Pressure of Plane Compressional Waves—F. E. Borgnis. (*Rev. Mod. Phys.*, vol. 25, pp. 653–664; July, 1953.) See 1878 of 1953.

534.4+681.142:519.272.1 2

The Correlatograph—Bennett. (*See* 58.)

534.3 3

The Audibility of Nonsinusoidal Variations of Pitch—E. Zwicker. (*Funk u. Ton*, vol. 7, pp. 342–346; July, 1953.) A recording audiometer was used in investigations of the limit of audibility of pitch variations due to modulation by triangular or rectangular pulses. The results obtained are shown graphically, together with comparative results for 4-cps sinusoidal FM.

534.833.4:534.321.9-14 4

Thermal Investigations on the Absorption of Ultrasonic Waves in Liquids—S. Parthasarathy, D. Srinivasan, and S. S. Chari. (*Z. Phys.*, vol. 135, pp. 395–402; July 31, 1953.) Measurements of the heat generated in various organic liquids exposed to ultrasonic radiation of frequencies between 340 and 660 kc indicate that the absorption coefficient over this range has the values given by the Stokes-Kirchhoff formula.

534.833.4:534.321.9-14 5

The Relation of the Temperature Effect, produced by Ultrasonic Waves in Liquids, to the Absorption Coefficient—S. Parthasarathy, D. Srinivasan and S. S. Chari. (*Z. Phys.*, vol. 135, pp. 403–405; July 31, 1953.) Measurements indicated that at 5 mc the quantity of heat produced was proportional to the coefficient of absorption for 13 out of 17 organic liquids tested. At 15 mc the quantity of heat produced in 6 liquids was constant and not proportional to the coefficient of absorption. See also 2543 of 1953.

534.84 6

Subjective Comparison of Concert Halls—T. Somerville. (*B.B.C. Quart.*, vol. 8, pp. 125–128; Summer, 1953.) A piece of music with a wide range of tone color and loudness was recorded, using an omnidirectional microphone, in St. Andrew's Hall, Glasgow, the Usher Hall, Edinburgh, the Royal Festival Hall, London, and the Civic Hall, Wolverhampton. Excerpts from the records were presented in a random order to 119 listeners, who were asked for opinions as to (*a*) tone quality, (*b*) definition (the ability to hear the majority of the instruments), (*c*) over-all preference. Only groups of skilled listeners gave results having statistical significance. The only valid results of the analysis show a preference for St. Andrew's Hall, with the Royal Festival Hall second.

534.843 7

The Interpretation of Directional-Characteristic Diagrams of Sound Emitters and Receivers—N. Meyer. (*Funk u. Ton*, vol. 7, pp. 398–404; Aug., 1953.) A method is shown of calculating the effect of the directional characteristic of the source (e.g. voice), the receiver (e.g. microphone) and the reverberation characteristic of the hall, on the characteristics of the sound transmitted. A practical example is analyzed.

621.395.623.7.001.4 8

Conditions for Acoustic Testing of Loudspeakers—P. Chavasse and R. Lehmann. (*Ann. Télécommun.*, vol. 8, pp. 226–237; July, 1953.) To specify the acoustic quality of a loudspeaker the following characteristics should be determined: (*a*) frequency response, (*b*) directivity, (*c*) harmonic and intermodulation distortion, (*d*) impedance as a function of frequency, (*e*) efficiency, (*f*) transient response. Specifications for the test chamber and measurement apparatus are suggested and a critical review is made of measurement methods.

621.395.625.3 9

Studies on Magnetic Recordings Part 2—Field Configuration around the Gap and the Gap-Length Formula—W. K. Westmijze. (*Philips Res. Rep.*, vol. 8, pp. 161–183; June, 1953.) Analysis is given for heads with (*a*) infinite gap, (*b*) thin gap, (*c*) semi-infinite gap. Part 1: 3471 of 1953.

621.395.625.3 10

Correction of Frequency-Response Variations caused by Magnetic-Head Wear—K. Singer and M. Rettinger. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 61, pp. 1–7; July, 1953.) Wear on a magnetic recording head reduces the front-gap to pole-face depth and thereby produces an increase of the gap reluctance. This results in a higher effective bias flux, which tends to attenuate the high frequencies. By reducing the bias current the optimum frequency response can be restored. The relations between the frequency response, head inductance and bias current are shown graphically and the change in sensitivity with change in head inductance is also given.

621.395.625.3:621.385.832 11

Electron-Beam Head for Magnetic Tape Playback—A. M. Skellett, L. E. Leveridge, and J. W. Gratian. (*Electronics*, vol. 26, pp. 168–171; Oct., 1953.) A reproducing head which is responsive to the instantaneous value of the magnetic flux rather than to its rate of change has a composite magnet system comprising a core structure external to a small cathode ray tube (length about 3 inches) with the pole pieces inside the tube and serving to deflect the beam over a split-target signal electrode. The tape record runs over the smoothly curved wings of the core. Typical frequency response curves are shown.

621.84/.85 12

Design Interrelations of Records and Reproducers—H. I. Reiskind. (PROC. I.R.E. [*Australia*], vol. 14, pp. 196–200; Aug., 1953. *Trans. I.R.E.*, No. PGA-5; Feb., 1952.) The dimensional characteristics of the groove in 33⅓- and 45-rpm gramophone records are discussed in relation to the recording characteristics. The quality of the reproducer pickup is assessed from intermodulation-distortion measurements made using a 400/4000-cps intermodulation signal recorded at various levels.

ANTENNAS AND TRANSMISSION LINES

621.315.212.4:[621.395+621.397.24 13

The Coaxial Cable of the New Italian Communication Network—R. Monelli. (*Alta Frequenza*, vol. 22, pp. 72–97; April, 1953.) The cable system described handles both telephone and television transmissions. Specifications, test methods and preliminary performance figures are given.

621.372.8 14

Determination of Reflection Coefficients and Insertion Loss of a Waveguide Junction—G. A. Deschamps. (*Jour. Appl. Phys.*, vol. 24, pp. 1046–1050; Aug., 1953.) A short-circuit in one of the waveguides is displaced by equal steps of, say, $\lambda/8$ and the corresponding reflection coefficients are measured in the other waveguide, their values, when plotted on the complex plane, falling on a circle. The required parameters are found from the position of points on an image circle derived by application of a bilinear transformation corresponding to the characteristics of the junction.

621.372.8 15

"Pipe-Lines" for Microwaves—(*Elec. Rev. [London]*, vol. 153, pp. 15–18; July 3, 1953.) Investigations of multichannel communication and power transmission along tubular waveguides are briefly reviewed. Practical applications are discussed.

621.396.67:[621.397.5+621.396.97.029.6 16

Combined Transmitting Aerials for Television and U.S.W. Broadcastings Part 2—W. Berndt. (*Telefunken Ztg*, vol. 26, pp. 268–279; Aug., 1953.) Decoupling between the vision and sound transmitters is considered. Eight different antenna-feeding arrangements are discussed. Two practical arrangements of combined television-u.s.w. antennas on one 30-m mast are illustrated and the mismatch /frequency and decoupling/frequency characteristics are shown graphically. Part 1: 37 of 1953.

621.396.674.3.029.62/.63 17

Television Aerials of the Future—F. R. W. Strafford. (*Wireless World*, vol. 59, pp. 506–508; Nov., 1953.) (Correction, *ibid.*, vol. 59, p. 580; Dec., 1953.) A discussion of problems to be faced in designing antennas for reception over frequency bands 1–5 (television and vhf sound), taking into account that horizontally as well as vertically polarized transmissions will be used. Difficulties in respect of propagation losses and energy collection at the higher frequencies are indicated. Possible alternatives are (*a*) a wide-band array covering the whole frequency range, (*b*) separate arrays for selected frequency bands, all permanently connected to the receiver via isolating networks, (*c*) separate arrays with switching arrangements. U.S.A. experience is referred to.

621.396.677 18

Gain, Effective Area and Gain Efficiency of Directional Aerials, and Methods for their Determination—G. F. Koch. (*Telefunken Ztg*, vol. 26, pp. 292–308; Aug., 1953.) A critical survey of the methods used for the determination of gain shows that gain efficiency cannot exceed 100 per cent, though gain efficiencies up to 180 per cent have been reported.

621.396.677.3 19

A New Approach to the Design of Superdirective Aerial Arrays—A. Bloch, R. G. Medhurst and S. D. Pool. (*Proc. IEE*, part III, vol. 100, pp. 303–314; Sept., 1953.) The current distribution required for maximum directivity of an array with a finite number of elements and any specified geometrical configuration is completely defined by the self and mutual resistances of the elements, and by the "resistance-voltage" component of the voltage across the terminals of each element, this component varying over the array in a specified manner. The maximum gain of the array is expressed either as a double sum containing the mutual conductances between the individual elements multiplied by trigonometrical factors depending on their spacing, or as an expression identical, except for a numerical factor, with that for the distant field of the array. This theory has been applied to the numerical calculation of certain simple arrays. Directivities greater than those obtained by conventional design methods can be achieved without excessive losses, for arrays of a given size. This was confirmed by experiments on a 75-mc four-element array.

621.396.677.32.029.63 20

A New Type of Aerial Radiating Longitudinally—J. C. Simon and G. Weill. (*Ann. Radioélect.*, vol. 8, pp. 183–193; July, 1953.) The theory of end-fire radiation due to variations in phase velocity along a dielectric-rod antenna (1244 of 1953) is developed and the design of all-metal antennas on this principle is discussed. Using a rod of length 4 λ with coaxial disks of different diameter at regular intervals along it, a gain of 16 db with swr <1.5 was maintained throughout the frequency range 1.7–2 kc. Values of gain and main-lobe width are given for antennas of length 4, 6, 20 and 80 λ.

621.396.677.83 21

Tuned Microwave Reflectors—S. Freedman. (*Radio and Telev. News, Radio-Electronic Eng. Suppl.*, vol. 50, pp. 13–15, 27; Aug., 1953.) A flat reflector is provided with specially shaped phase-shifting ends which are tuned by appropriate dimensioning according to the frequency employed and the desired angle of reflection. The resultant phase shift increases the efficiency of the reflector to that of an infinite plane. The relations between efficiency, frequency and direction of reflection are shown graphically. Design considerations are discussed.

AUTOMATIC COMPUTERS

681.142 22

Computing Bit by Bit or Digital Computers Made Easy—A. L. Samuel. (Proc. I.R.E., vol. 41, pp. 1223–1230; Oct., 1953.) An introduction to the subject, with explanations of the specialized terminology.

681.142 23

Can Machines Think?—M. V. Wilkes. (Proc. I.R.E., vol. 41, pp. 1230–1234; Oct., 1953. *Discovery*, vol. 14, pp. 151–154; May 1953.) A general survey of the capabilities and limitations of digital computers.

681.142 24

Computers and Automata—C. E. Shannon. (Proc. I.R.E., vol. 41, pp. 1234–1241; Oct., 1953.) Recent developments in the field of automata and nonnumerical computers are reviewed; some typical machines, including games-playing machines are described. A comparison is made between the operation of computers and of the brain.

681.142 25

Fundamentals of Digital Computer Programming—W. H. Thomas. (Proc. I.R.E., vol. 41, pp. 1245–1249; Oct., 1953.) The process of programing is explained in relation to a typical but simplified stored-program computer, using 8 basic instructions. Elementary arithmetical and logical computation are illustrated. The ability of the computer to modify its own program is demonstrated.

681.142 26

Influence of Programming Techniques on the Design of Computers—G. M. Hopper and J. W. Mauchly. (Proc. I.R.E., vol. 41, pp. 1250–1254; Oct., 1953.) New techniques are discussed which take advantage of the high-speed checked operation of the computer to perform the routine work of programming.

681.142 27

Analogue vs. Digital Computers—A Comparison—M. Rubinoff. (Proc. I.R.E., vol. 41, pp. 1254–1262; Oct., 1953.) The operation of analogue computers is described. The problem of simulation in real time is used as a basis for comparing analogue and digital computers in respect of speed, accuracy, convenience, etc. The mathematical operations discussed include addition, subtraction, multiplication, division, evaluation of polynomials and other functions, extraction of square roots, and solution of differential equations.

681.142 28

The System Design of the IBM Type 701 Computer—W. Buchholz. (Proc. I.R.E., vol. 41, pp. 1262–1275; Oct., 1953.) Attention is directed specially to the new features of this machine, which include improved arithmetic and logical facilities and the control of the extensive input and output arrangements directly from the stored programme. 18-bit instructions cover 33 distinct operations.

681.142 29

Engineering Description of the IBM Type 701 Computer—C. E. Frizzell. (Proc. I.R.E., vol. 41, pp. 1275–1287; Oct., 1953.) A detailed description is given of this high-speed computer for large-scale scientific work. A complete installation includes analytical control unit, e.g. storage unit, magnetic-tape and magnetic-drum readers and recorders, punched-card reader, alphabetic printer, punched-card recorder, power distribution unit and power supply unit. A note on maintenance is included. *See* also 28 above.

681.142 30

The Arithmetic Element of the IBM Type 701 Computer—H. D. Ross, Jr. (Proc. I.R.E., vol. 41, pp. 1287–1294; Oct., 1953.) A storage circuit giving a 1-μsec delay is described; its output is at either +10 v or −30 v. Used in conjunction with direct-coupled diode switching, this arrangement leads to simplifications in shifting registers and in performing division and testing for zero in the accumulator.

681.142 31

The SWAC [National Bureau of Standards Western Automatic Computer]—Design Features and Operating Experience—E. D. Huskey, R. Thorensen, B. F. Ambrosio and E. C. Yowell. (Proc. I.R.E., vol. 41, pp. 1294–1299; Oct., 1953.) This high-speed computer uses a Williams-tube main storage system, an auxiliary magnetic-drum storage system, and punched-card input and output systems. A review is presented of problems solved by the computer during the preceding year. For a less detailed account see *Tech. News Bull. Nat. Bur. Stand.*, vol. 37, pp. 145–150; Oct., 1953.

681.142 32

SEAC [National Bureau of Standards Eastern Automatic Computer]—S. Greenwald, R. C. Haueter and S. N. Alexander. (Proc. I.R.E., vol. 41, pp. 1300–1313; Oct., 1953.) A comprehensive description (see also 385 of 1951) is given of this computer, which has been in operation almost continuously since the spring of 1950. It is a synchronous machine, operating at 1 mc. It has been expanded from its original form, with increase in the number of thermionic valves from about 750 to 1300, and of Ge diodes from about 10,500 to 16,000. Circuit modifications include the introduction of ac coupling using pulse transformers.

681.142 33

Electronic Circuits of the NAREC [Naval Research Laboratory Computer]—P. C. Shertz. (Proc. I.R.E., vol. 41, pp. 1313–1320;

Oct., 1953.) Crystal diodes are used where practical for all nonlinear functions in this digital computer, thermionic valves being used only for inversion or amplification. All circuits are designed to provide low-impedance output signals whose maximum and minimum levels are held to prescribed potentials. The circuits are packaged in standard 4-tube plug-in units. See also *Elect. Eng.*, vol. 70, p. 111; Feb., 1951 (Gridley and Sarahan).

681.142 34
Diagnostic Programmes for the ILLIAC [University of Illinois' digital computer]—D. J. Wheeler and J. E. Robertson. (Proc. I.R.E., vol. 41, pp. 1320–1325; Oct., 1953.) A detailed description is given of a "leapfrog" program for detecting faults; this ensures that every component of the computer is tested under all the different conditions of use.

681.142 35
The Logistics Computer—R. S. Erickson. (Proc. I.R.E., vol. 41, pp. 1325–1332; Oct., 1953.) Construction and operation of a digital computer for logistics research are described. Consideration of the particular requirements for this work led to a design incorporating high-speed magnetic-tape (in addition to punched-tape) input and output equipment, fixed sequence operation, magnetic-drum storage with capacity for 180,000 decimal digits, and plug-board programming.

681.142 36
The Remington Rand Type 409-2 Electronic Computer—L. P. Crosman. (Proc. I.R.E., vol. 41, pp. 1332–1340; Oct., 1953.) Construction and operation are described of a machine of moderate size, for accounting purposes, using 1476 thermionic valves and 1128 cold-cathode gas diodes. Programing is directed by means of plug-boards. There are 40 calculating-steps and 6 special steps; each step is automatically tested by a reverse process. A "reproduce" feature is included.

681.142 37
The Design of the Bendix Digital Differential Analyzer—M. Palevsky. (Proc. I.R.E., vol. 41, pp. 1352–1356; Oct., 1953.) "The error reduction scheme employed by the Bendix Digital Differential Analyzer, which utilizes both ternary intercommunication between integrators and trapezoidal integration, is described. A device for the variation and automatic resetting of initial conditions is discussed together with an example illustrating its operation. Finally, techniques employed for minimizing the size of the computer are considered."

681.142 38
Theory of Logical Nets—A. W. Burks and J. B. Wright. (Proc. I.R.E., vol. 41, pp. 1357–1365; Oct., 1953.) Two-valued logic is applied to the study of digital-computer circuits. A "logical net" is an array constructed from a "strike element," representing circuit components performing logical functions, and a "delay element," representing storage components. Various types of net are defined, their properties are correlated with those of the associated set of equations, and a study is made of the operations they can perform. The classification of digital-computer circuits in this system is discussed.

681.142 39
Dynamic Circuit Techniques used in SEAC and DYSEAC—R. D. Elbourn and R. P. Witt. (Proc. I.R.E., vol. 41, pp. 1380–1387; Oct., 1953.) The circuits of these digital computers are standardized on the basis that Ge diodes are used for all logical operations, delay lines for storage, and thermionic valves for amplification. Manufacture is simplified by using only two types of plug-in unit.

681.142 40
The Design of Logical OR-AND-OR Pyramids for Digital Computers—S. E. Gluck, H. J. Gray, Jr., C. T. Leondes and M. Rubinoff. (Proc. I.R.E., vol. 41, pp. 1388–1392; Oct., 1953.) Recent improvements in design technique are described, leading to reduction of number of diodes required, reduction of voltage and power requirements, and speeding up of operation. Basic considerations involved in the choice of elements are discussed.

681.142 41
A Survey of Digital Computer Memory Systems—J. P. Eckert, Jr. (Proc. I.R.E., vol. 41, pp. 1393–1406; Oct., 1953.) A survey limited to systems whose reaction time is shorter than human reaction time.

681.142 42
The Logical Principles of a New Kind of Binary Counter—W. H. Ware. (Proc. I.R.E., vol. 41, pp. 1429–1437; Oct., 1953.) Auxiliary temporary storage in a binary counter is provided, during intervals over which the main storage element is in a state of change, by elements having the same time-indefiniteness as the main storage elements. This leads to a device which has "true" and "false" ranks of storage elements. The basic device is generalized to one with four modes of operation, and the rules of number formation are discussed. Application to a computer is described.

681.142 43
Combined Reading and Writing on a Magnetic Drum—J. H. McGuigan. (Proc. I.R.E., vol. 41, pp. 1438–1444; Oct., 1953.) A technique is described which uses a single head for storing information and reading it back. Design of a suitable amplifier for this method of operation is discussed. The process can be performed with successive cells at the repetition rate of 60,000/seconds. The technique extends the use of magnetic drums to data processing as well as storage.

681.142 44
Coded Decimal Number Systems for Digital Computers—G. S. White. (Proc. I.R.E., vol. 41, pp. 1450–1452; Oct., 1953.) "From the very large numbers of possible ways of coding the ten decimal digits into sets of binary elements, a restricted group of number systems is presented as being particularly adaptable for representing the decimal digits in digital computers. As an example, one number system of this group is presented which permits the construction of a very simple addition circuit."

681.142 45
An Electromagnetic Clutch for High Accelerations—S. M. Oster and L. D. Wilson. (Proc. I.R.E., vol. 41, pp. 1453–1455; Oct., 1953.) Unidirectional and bidirectional models are discussed of a clutch with many possible applications in the digital-computer field.

681.142 46
A Survey of Analog-to-Digital Converters—H. E. Burke, Jr. (Proc. I.R.E., vol. 41, pp. 1455–1462; Oct., 1953.)

681.142 47
An Analog-to-Digital Converter for Serial Computing Machines—H. J. Gray, Jr., P. V. Levonian and M. Rubinoff. (Proc. I.R.E., vol. 41, pp. 1462–1465; Oct., 1953.) A two-stage process is used, the analogue number being first converted into a "reflected-binary" or "cyclic" code number, which is then converted into the binary form as a time sequence with the least significant digit first.

681.142 48
Effectiveness of Two-Step Smoothing in Digital Control Computers—R. E. Spero. (Proc. I.R.E., vol. 41, pp. 1465–1469; Oct., 1953.) The use of a digital computer as an element in a control system, to smooth observational data, is discussed. Analysis is presented for a two-stage system, in which the first stage has a high data-handling rate and short smoothing time while the second has a low data-handling rate and a long smoothing time.

681.142 49
An Input-Output Unit for Analog Computers—P. R. Vance and D. L. Haas. (Proc. I.R.E., vol. 41, pp. 1483–1486; Oct., 1953.) A device is described which operates as a curve-follower type of function generator to provide an input unit, or as an X-Y recorder when used as output unit.

681.142 50
Application of Electronic Differential Analyzers to Engineering Problems—C. A. Meneley and C. D. Morrill. (Proc. I.R.E., vol. 41, pp. 1487–1496; Oct., 1953.)

681.142 51
The Solution of Partial Differential Equations by Difference Methods using the Electronic Differential Analyzer—R. M. Howe and V. S. Haneman, Jr. (Proc. I.R.E., vol. 41, pp. 1497–1508; Oct., 1953.)

681.142:621.396.822 52
Analog Computing Applied to Noise Studies—R. R. Bennett. (Proc. I.R.E., vol. 41, pp. 1509–1513; Oct., 1953.) Methods suitable for studying both linear and nonlinear systems are discussed.

681.142 53
EDVAC Drum-Memory Phase System of Magnetic Recording—D. Eadie. (*Elect. Eng.*, vol. 72, pp. 590–595; July, 1953.)

681.142 54
Dead Programmes for a Magnetic Drum Automatic Computer—W. L. van der Poel. (*Appl. Sci. Res.*, vol. B3, pp. 190–198; 1953.) Breakdown due to a wrong instruction can be prevented by blocking part of the magnetic-drum memory for writing, and placing in this "dead" part standard sub-programs which occur frequently.

681.142 55
Electrical Analogues—G. Liebmann. (*Brit. Jour. Appl. Phys.*, vol. 4, pp. 193–200; July, 1953.) Description of the principles, operation and applications of analogue techniques using conducting paper, the electrolyte tank, resistive networks, or networks with both resistive and reactive components.

681.142:33 56
Economic Analogs—O. J. M. Smith. (Proc. I.R.E., vol. 41, pp. 1514–1519; Oct., 1953.) Discussion of the possibilities of using analogue computers of various types for solving problems of commerce, industry, etc.

681.142:512 57
Elements of Boolean Algebra for the Study of Information-Handling Systems—R. Serrell. (Proc. I.R.E., vol. 41, pp. 1366–1380; Oct., 1953.) Terminology, symbols, definitions, and proofs of basic theorems of Boolean algebra are given as a preliminary to the use of this algebra for studying methods of minimizing the number of physical elements required in information-handling or computing systems. The method is illustrated by examples of design of computer circuits.

681.142:519.272.1:534.44 58
The Correlatograph—W. R. Bennett. (*Bell Sys. Tech. Jour.*, vol. 32, pp. 1173–1185; Sept., 1953.) A paper-tape-recorder analogue device for the continuous display of the short-term correlation function by means of a time/time-lag/correlation-factor plot (x-y-intensity) is described. It was designed for the analysis of 200-cps—4-kc signals in a magnetic-tape recording. The electromechanical design is based

on the af spectrograph [3517 of 1946 (Koenig et al.)].

681.142:538.221 59

A Myriabit Magnetic-Core Matrix Memory—J. A. Rajchman. (PROC. I.R.E., vol. 41, pp. 1407–1421; Oct., 1953.) Description of an experimental information-storage system providing random access to any one of 10,000 bits in a few μsec. The system is an extension of that previously described by Papian (2258 of 1952). Details are given of the cores, which are of ferromagnetic ceramic material and of diameter 0.054 inches, and of the operation of the array.

681.142:621.318.5 60

Machine Aid for Switching Circuit Design—C. E. Shannon and E. F. Moore. (PROC. I.R.E., vol. 41, pp. 1348–1351; Oct., 1953.) An experimental special-purpose nonnumerical computer for analysis of relay circuits has as its input a relay circuit together with the specification the circuit is intended to satisfy. The analyzer checks the circuit against the specification, makes systematic attempts to simplify the circuit by removing redundant contacts, and establishes lower limits for the numbers and types of contact needed to satisfy the specifications. This method is compared with the method in which the same operations are performed by coding them on a general-purpose digital computer.

681.142:621.374.5 61

The Mercury-Delay-Line Storage System of the ACE Pilot Model Electronic Computer—E. A. Newman, D. O. Clayden, and M. A. Wright. (*Proc. IEE*, part II, vol. 100, pp. 445–452; Aug., 1953.)

681.142:621.375.3 62

The Magnetic Amplifier as an Analog Computer Component—L. J. Craig. (PROC. I.R.E., vol. 41, pp. 1477–1482; Oct., 1953.)

681.142:621.376.332 63

An A.M.-F.M. Electronic Analog Multiplier—W. A. McCool. (PROC. I.R.E., vol. 41, pp. 1470–1477; Oct., 1953.) A multiplier is described in which frequency deviation and carrier amplitude are made respectively proportional to the two quantities to be multiplied. The product is obtained by means of a Foster-Seeley discriminator. Stabilization is provided by means of feedback. The long-term error is <1 per cent of full-scale output.

681.142:621.395.34 64

Electronic Computers and Telephone Switching—W. D. Lewis. (PROC. I.R.E., vol. 41, pp. 1242–1244; Oct., 1953.) A discussion of the possibility of using digital-computer techniques in automatic telephone switching systems.

681.142:621.395.34 65

An Automatic Telephone System employing Magnetic Drum Memory—W. A. Malthaner and H. E. Vaughan. (PROC. I.R.E., vol., 41, pp. 1341–1347; Oct., 1953.) The DIAD (drum information assembler and dispatcher) system is described. A capacitive scanner acts as a commutator for conveying signals from subscribers' sets to storage on magnetic drums. These signals are combined with permanently stored information, processed in accordance with built-in programs, and used to control call-switching circuits. The technical feasibility of the system has been demonstrated by large scale laboratory experiments.

681.142:778.3 66

Photographic Techniques for Information Storage—G. W. King, G. W. Brown, and L. N. Ridenour. (PROC. I.R.E., vol. 41, pp. 1421–1428; Oct., 1953.) Storage devices with capacities of 10^7–10^9 bits and useful access times can be produced by means of photographic techniques. The difficulty that rapid erasing and rewriting are not possible with photographic media is overcome by providing a small auxiliary unit such as a magnetic-drum store. Flying-spot scanners can be used for reading.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.2.012.3 67

Audio Transformer Design Charts—T. Halabi. (*Electronics*, vol. 26, pp. 193–194, 196; Oct., 1953.)

621.314.22.015.7 68

The Design of a Peaking Transformer—A. B. Thomas. (*Jour. Brit. IRE*, vol. 13, pp. 486–489; Nov., 1953.) Reprint. See 2233 of 1953.

621.314.222.026.444:621.384.612 69

A Wide-Band High-Power Radio-Frequency Transformer—L. U. Hibbard, W. Raudord, and L. Riddiford. (*Jour. Sci. Inst.*, vol. 30, pp. 245–250; July, 1953.) Detailed description of a transformer forming part of the rf system of the Birmingham synchrotron (2092 of 1953). It is designed for operation from a push-pull amplifier and will deliver 10 kv throughout the frequency range 0.2–10 mc with a duty cycle of 1 second in every 10 seconds.

621.314.7:621.375.427 70

A Crystal-Triode Push-Pull Amplifier—J. I. Missen. (*G.E.C. Jour.*, vol. 20, pp. 144–150; July, 1953.) The amplifier described uses two point-contact transistors and is capable of delivering up to 420 mw output power at an efficiency of 28 per cent. A power gain of 12 db at 10 per cent harmonic distortion is obtained.

621.314.7–713 71

Kerosene-Cooled Transistors—J. E. Maynard and R. L. Brock. (*Electronics*, vol. 26, pp. 202, 204, 206; Oct., 1953.) The temperature rise in the contact regions of a point-contact transistor is considerably reduced by circulating kerosene around the junctions; circulation is maintained by the natural convection currents set up.

621.318.4.011/.012 72

The Design of High-Q Iron-Cored Inductors—N. H. Crowhurst. (*Electronic Eng.*, vol. 25, pp. 478–482; Nov., 1953.) Design charts are presented for inductors not carrying dc; the attainment of maximum Q requires the reduction of over-all losses, including those due to core magnetization, to a minimum. For data on inductors carrying dc, see 1096 of 1951.

621.318.435.3:621.314.12 73

The Study of a Magnetic Inverter for Amplification of Low-Input-Power D.C. Signals—E. H. Frost-Smith. (*Proc. IEE*, part II, vol. 100, pp. 362–371; Aug., 1953. Discussion, pp. 371–375. Digest, *ibid.*, part III, vol. 100, p. 318; Sept., 1953.) The principle of operation of the magnetic inverter differs from the conventional magnetic amplifier in that the load current corresponds to the even-harmonic mmf's present in the asymmetrically excited iron-core system. Power gains up to 10^3 can be obtained with time constants of about 1 second, when operating at 50 cps; the zero stability is within 10^{-12} w.

621.318.57:621.314.63 74

Semiconductor Diode Gates—L. W. Hussey. (*Bell. Sys. Tech. Jour.*, vol. 32, pp. 1137–1154; Sept., 1953.) The general properties of gate circuits are discussed and a simple design analysis of transmission-type and switching-type gates is given. The results of an experimental check are noted, and the suitability of point-contact and junction-type rectifiers for different purposes is discussed.

621.319.43:621.317.3 75

Study and Evaluation of the Microphony due to Vibration of the Plates of a Variable Capacitor—E. Briganti. (*Poste e Telecomunicazioni*, vol. 21, pp. 363–368; Aug., 1953.) Capacitance variations due to vibrations are investigated by observing the resulting distortion, the latter being evaluated from the ratio between the third harmonic and the fundamental when a signal of amplitude about 100 v is applied. This ratio is expressed as a function of the Q of the measuring circuit, details of which have been given previously [1035 of 1943 (Brunetti and Greenough)]. The effect of variations of supply voltage is shown in curves.

621.372 76

The Fundamental Theorem of Electrical Networks—J. L. Synge. (*Quart. Appl. Math.*, vol. 11, p. 215; July, 1953.) A fuller proof is given than that presented in the original paper (347 of 1952).

621.372 77

Application of Complex Symbolism to Linear Variable Networks—A. P. Bolle. (*Tijdschr. ned. Radiogenoot.*, vol. 18, pp. 231–246; July, 1953. In English.) The theory developed is based on frequency domain analysis. Equations are derived for networks containing one linear variable element the value of which varies periodically and is capable of being developed in a Fourier series. The usefulness of these equations for computations of magnetic and dielectric modulators and amplifiers is indicated.

621.372.413 78

The Utilization of Degenerate Modes in a Spherical Cavity—M. R. Currie. (*Jour Appl. Phys.*, vol. 24, pp. 998–1003; Aug., 1953.) Formulas are derived for calculating the coupling coefficients between an arbitrary number of modes excited simultaneously in a cavity resonator. The perfect symmetry of the sphere gives rise to a high degree of degeneracy, i.e. many independent field configurations with the same natural frequency. Experiments on a spherical cavity gave results in good agreement with the theory. Practical coupling devices discussed include irises, loops and small volumes of conductor placed within the cavity. See also 625 of 1952 (Lin).

621.372.413 79

Microwave Cavity Resonators. Some Perturbation Effects and their Applications—S. K. Chatterjee. (*Jour. Brit. IRE*, vol. 13, pp. 475–484; Nov., 1953.) First-order perturbation theory is used to examine the change in resonance frequency and Q of a cavity due to a small deformation of the wall. An expression is derived for the coupling coefficient between the two degenerate modes H_{01} and E_{11}; in the absence of perturbation these two modes can co-exist. Applications of the theory to measurements of dielectric properties, study of ferromagnetic resonance etc. are described. 33 references.

621.372.5 80

Simplification of the Tensor Analysis of Networks—J. Thouzéry. (*Radio. Tech. Dig. Édn franç.*, vol. 7, pp. 145–161; 1953.)

621.372.5 81

Power and Efficiency of Passive Quadripoles—C. Bordone and G. G. Sacerdote. (*Alta Frequenza*, vol. 22, pp. 98–107; April, 1953.) Families of circle diagrams are used to study the output power, efficiency, insertion loss and attenuation of quadripoles.

621.372.5.029.64:538.614 82

Investigation of Nonreciprocal Quadripoles in the Centimetre Wave Range—A. A. T. M. van Trier. (*Tijdschr. ned. Radiogenoot.*, vol. 18, pp. 211–229; July, 1953.) Devices based on the Faraday rotation of guided waves are investigated [see e.g. 1233 of 1952 (Hogan)]; a method is developed for determining the quadripole

parameters of these devices. A waveguide arrangement adapted for measurements on gyrators is described; results are given for a gyrator using ferroxcube IV E. Factors to be considered in designing gyrators for particular applications are indicated. See also 2890 of 1953.

621.372.54 83

The Design of Zig-Zag Filters—T. Laurent. (*Ericsson Tech.*, vol. 9, pp. 83–108; 1953.) Formulas are derived for the computation of a filter with a known distribution of cut-off and peak-attenuation frequencies and given image impedances. The matching conditions between the sections and the method of determining the order of sections in a ladder-network by means of a zig-zag diagram is explained. A numerical example of lf band-pass filter design is given.

621.372.54 84

Old and New Methods for designing Composite High-Frequency Filter Circuits and their Application to Filter Circuits with Low Relative Bandwidth—R. Rücklin. (*Arch. elekt. Übertragung*, vol. 7, pp. 363–374; Aug., 1953.) A calculation based on that given by Edelmann for Tchebycheff-type filters (88 of 1952) is used to extend Schienemann's formulas for multistage filters (2016 of 1939) to the case of finite ripples of the response curve. For low relative bandwidths, simple formulas are derived for the form factors and time constants of the double-circuit stages. As the ripples tend to zero, further simplifications can be introduced. Modification of the method is indicated to deal with cases where large ripples are permitted for the sake of selectivity.

621.372.54:517.727 85

Practical Calculation of Jacobi's "sn" Elliptic Function, with Special Reference to the Determination of the Cauer Parameter for Filters with Operative Attenuation Characteristics—V. Fetzer. (*Arch. elekt. Übertragung*, vol. 7, pp. 393–401; Aug., 1953.) Elliptic functions encountered in calculations of Tchebycheff-type filters are considered. The method of deriving these functions and hence the attenuation/frequency curve is shown in detail for the case of an antimetrical low-pass filter. Formulas for calculating the functions are tabulated in an appendix.

621.372.54:621.372.412 86

Quartz Crystals for Filters—R. A. Spears. (*A.T.E. Jour.*, vol. 9, pp. 149–157; July, 1953.) Outline account of the characteristics of crystals and their use in filter networks.

621.372.542.4:621.396.67 87

Electrical High-Frequency Separating Filter Networks—R. Becker. (*Telefunken Ztg.*, vol. 26, pp. 280–291; Aug., 1953.) Frequency-separating, diplexer and notch-diplexer networks are described. Two classes of frequency-separating networks are distinguished: (*a*) networks with series-tuned circuits in parallel (F_{SP}) and parallel-tuned-circuits in series (F_{PS}), (*b*) parallel-parallel (F_{PP}) and series-series (F_{SS}) tuned-circuit networks. The former are of the frequency-rejector type, the latter of the frequency-acceptor type. The application of theory is shown in several numerical examples of design of transmitter hf separating filter networks; the results are compared with the observed characteristics.

621.372.6 88

The Necessary and Sufficient Conditions for the Physical Realizability of a $2m$-Rank Field Impedance (Admittance) Matrix, in the Form of a Passive Multipole—V. A. Taft. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 84, pp. 499–501; May 21, 1952. In Russian.) It is proved that every positive real matrix of $2m$-rank, where m is a finite integral number, can be physically realized in the form of a passive multipole.

621.373.2.029.65 89

A Critical Review of Researches into Millimetric-Wave Spark Generators—M. H. N. Potok. (*Jour. Brit. IRE*, vol. 13, pp. 490–497; Nov., 1953.) Methods using cylindrical and spherical electrodes and arrays of these are discussed; other methods are mentioned. Difficulties involved are indicated. 66 references.

621.373.4:621.316.726 90

Magnetic Control of the Frequency of Valve Transmitters by means of an Electrodeless Gas Discharge—B. Koch and H. Neuert. (*Z. angew. Phys.*, vol. 5, pp. 249–251; July, 1953.) The tuned-circuit inductor of an oscillator was wound round an argon-filled spherical container in which an electrodeless discharge could be maintained, and a magnetic field was applied parallel to the plane of the coil. The oscillator frequency was found to vary almost linearly over a certain range of field strengths, and was also dependent on the gas pressure in the discharge vessel.

621.373.42.029.4/.51 91

Bridge-Stabilized Ultrasonic Oscillator—L. W. Erath. (*Electronics*, vol. 26, pp. 174–175; Oct., 1953.) Good frequency stability and low harmonic distortion are obtained by arranging a Wien-bridge *RC* oscillator so that no arm of the bridge is shunted by a low impedance. An oscillator with five overlapping ranges covering 1 cps–120 kc is described.

621.373.421.13:621.372.412 92

Frequency Stability and Accuracy of a Crystal Oscillator—E. J. Post and J. W. A. van der Scheer. (*Tijdschr. ned. Radiogenoot.*, vol. 18, pp. 183–210; July, 1953.) Properties of crystals relevant to their use for frequency stabilization are reviewed. The occurrence and elimination of crystal oscillations at undesired frequencies are discussed. Among the factors causing deviation of frequency from the nominal value the most important is the circuit arrangement; the requirements for satisfactory crystal oscillater circuits are examined in detail. A table indicates the frequency range, maximum instability, maximum inaccuracy, maximum total frequency deviation, and energy output of some known circuits.

621.373.423 93

Frequency Stabilization of a Reflex-Klystron Oscillator—F. Bruin. (*Appl. Sci. Res.*, vol. B3, pp. 199–200; 1953.) Part of the klystron output is fed to a cylindrical resonator, one of whose ends is constituted by a diaphragm which is vibrated at about 1 kc. The detected output from the resonator, which varies with shift of the klystron frequency, is fed to a 1-kc lock-in amplifier to produce a control voltage which is added to the klystron reflector voltage.

621.373.431 94

The Miller Integrator as Sawtooth-Voltage Generator—A. Nowak. (*Radio Tech.*, [*Vienna*], vol. 29, pp. 235–240; July, 1953.) The theory of the Miller integrator is discussed and practical circuits of the transitron and sanatron (an integrator with separate controlling tube) are given. The choice of components and component tubes is considered in detail.

621.375.024 95

The Parallel-T D.C. Amplifier: A Low-Drift Amplifier with Wide Frequency Response—P. S. T. Buckerfield. (*Proc IEE*, part II, vol. 100, pp. 375–376; Aug., 1953.) Discussion on 982 of 1953. Please note change in U.D.C. number.

621.375.13.018.756:621.314.7 96

A Transistor Pulse Amplifier using External Regeneration—J. H. Vogelsong. (Proc. I.R.E., vol. 41, pp. 1444–1450; Oct., 1953.) Regenerated pulses having desired waveforms, as required in synchronous serial computers, are obtained by means of a circuit using a point-contact transistor operated as an over-driven amplifier with external feedback. Provision is made for synchronizing the output pulses with a master clock. Transformer coupling provides dc restoration.

621.375.221 97

Neutralizing Pentodes in Radar I.F. Stages—J. C. Tellier. (*Electronics*, vol. 26, pp. 184–186; Oct., 1953.) The effects of grid-plate capacitance are cancelled without use of additional components, tapped coils or balanced tuned circuits by using as neutralizing capacitor a screen-bypassing capacitor providing slightly less than complete bypassing. Experimental methods are described for determining the required values of the neutralizing capacitors for radar and television IF amplifiers.

621.375.3 98

The Magnetic Amplifier—W. C. Johnson. (*Elect. Eng., N.Y.*, vol. 72, pp. 583–588; July, 1953.) Comparison of the elementary magnetic amplifier with the type using external feedback.

621.375.3 99

Design of Magnetic Amplifiers—W. Schmidt. (*Funk u. Ton*, vol. 7, pp. 347–358; July, 1953.) An outline of basic design principles.

621.375.5 100

Nonlinear Capacitors for Dielectric Amplifiers—G. S. Shaw and J. L. Jenkins. (*Electronics*, vol. 26, pp. 166–167; Oct., 1953.) The operation of the dielectric amplifier depends on the variation of dielectric constant with applied voltage. Requirements for materials suitable for high-frequency small-signal applications are discussed. A two-stage amplifier with RF power supply and gain of 30 db is briefly described.

621.396.6 101

New Constructional Techniques—G. W. Dummer and D. L. Johnston. (*Electronic Eng.*, vol. 25, pp. 417–421 and 456–461; Oct./Nov., 1953.) Survey of modern methods for producing radio equipment, including potting, circuit printing and automatic soldering. 61 references. See also 2926 of 1953.

621.3.066.6:537 102

Radio Research Special Report No. 24. Fundamental Processes of Electrical Contact Phenomena. [Book Notice]—Jones. (*See* 122.)

GENERAL PHYSICS

534.01 103

On Interaction of Nonlinear Oscillations—N. Minorsky. (*Jour. Franklin Inst.*, vol. 256, pp. 147–165; Aug., 1953.) General considerations regarding the existence, stability and certain topological rules of selection between several oscillations are outlined. The results are used to investigate the differential equation

$$x+e(x^2-1)x+[1+(a-cx^2)\cos 2t]x=0$$

using the "stroboscopic" method (2951 of 1951). Some special cases are examined.

535.215 104

The Photoeffect in a Uniform Electric Field—H. Überall. (*Acta Phys. austriaca*, vol. 7, pp. 14–22; April, 1953.) It is shown theoretically that even very high external fields have little effect on the angular distribution of the photoelectrons or on the total effective atomic cross section.

537.122:[537.291+538.691 105

Free Electron in an E and H Field—A. Spork. (*Öst. Z. Telegr. Teleph. Funk Fernschtech.*, vol. 7, pp. 85–98; July/Aug., 1953.) The general equations of motion are obtained for an electron in electric and magnetic fields. Particular cases, for various angles between the field and velocity vectors, are considered.

537.122:551.510.535 106

Experimental Studies of the Motions of Slow Electrons in Air, with Application to the Ionosphere—R. W. Crompton, L. G. H. Huxley, and D. J. Sutton. (*Proc. Roy. Soc. A.*, vol. 218, pp. 507–519; July 23, 1953.) From experiments on the motion of slow electrons in air, empirical formulas were deduced from which the electronic temperature, collision frequency, drift velocity and the average energy loss per collision (ΔQ) can de derived. The collision cross section of molecules of air and nitrogen at electron temperatures up to 2600 degrees K is directly proportional to the speed of the electrons. The electron collision frequency in the ionosphere up to 94 km is linearly proportional to the pressure. The important result, that ΔQ is proportional to $[(Q-Q_0)/Q]^2$ and not to $(Q-Q_0)$, as believed hitherto, where Q is the agitational energy of the electron and Q_0 the energy of thermal agitation of the gas molecule, is discussed briefly.

537.224 107

The Case of Charged Density Distribution versus Semipermanent Polarization as a Basis for Electret Behavior—W. F. G. Swann. (*Jour. Franklin Inst.*, vol. 256, pp. 167–175; Aug., 1953.) It is shown that if an electret were composed of ordinary charge distributions, its effects would diminish by 50 per cent in a day or two; the observed effects indicate the existence of a semipermanent polarization, and can be regarded as arising from a lag in the cancelling action of conductivity.

537.226:539.11 108

Investigation of the Interaction between an Electron and a Lattice Oscillator by means of a [one-dimensional] Model—H. Haken. (*Z. Phys.*, vol. 135, pp. 408–430; July 31, 1953.)

537.226.2/.3:546.212 109

The Dielectric Properties of Water in Solutions—J. B. Hasted and S. H. M. El Sabeh. (*Trans. Faraday Soc.*, vol. 49, pp. 1003–1011; Sept., 1953.) Measurements have been made of the microwave dielectric constants and losses of water and some aqueous solutions over the temperature range 0–60 degrees C, using methods described by Collie et al. (2508 of 1948). The dielectric constant of water at cm λ rises from 5.0 ± 0.5 at 0 degrees C to 5.9 ± 0.5 at 60 degrees C. Results for the solutions are tabulated and their interpretation is discussed.

537.228.1 110

The Linear Piezoelectric Equations of State—R. Bechmann. (*Brit. Jour. Appl. Phys.*, vol. 4, pp. 210–212; July, 1953.) The different forms of the equations of state obtained from various combinations of stress, strain, electric field, electric displacement and polarization are given in tabular form.

537.311.33 111

Electro-thermal Behaviour of Point Contacts to Semiconductors—A. D. Stuckes. (*Proc. Phys. Soc.*, vol. 66, pp. 570–587; July 1, 1953.) Joule heating effects, due to current flow through a point contact to a semiconductor with no barrier layer, were investigated theoretically and the results were confirmed experimentally for tungsten points in contact with sintered disks of Mg_2TiO_4. The values of the contact radius calculated from dc and ac experiments are in fair agreement, as are the calculated and measured variations with frequency of the impedance to sinusoidal ripple superimposed on steady current. Minor anomalies in the dc characteristic are assumed to be due to field emission around the contact.

537.52 112

The Ignition Voltage of Gas Discharges in a Transverse Magnetic Field over the Pressure Range $10-10^{-8}$ Torr—R. Haefer. (*Acta Phys. austriaca*, vol. 7, pp. 52–90; April, 1953.) Experiments were made using a discharge space bounded by coaxial cylinders, with a coaxial magnetic field. A voltage of 2.2 kv was sufficient to start a discharge at 1.5×10^{-8} Torr with a suitable magnetic field strength. On the basis of these measurements and those of other workers, expressions are derived relating ignition voltage to magnetic field strength, electrode separation (d) and configuration, electrode material, nature and pressure (p) of the gas. Above a critical value of pd (1.07 Torr cm for air) the magnetic field raises the ignition voltage; below this critical value the magnetic field reduces the ignition voltage.

537.525.72 113

The Growth of the High-Frequency Electrodeless Discharge—G. Francis and A. von Engel. (*Phil. Trans. A.*, vol. 246, pp. 143–180; July 9, 1953.) The various stages in the growth of electrodeless discharges are discussed in detail, with particular consideration of electron-multiplication processes, increase of the number of positive ions, and the effect of charges on the wall of the discharge tube. A new experimental technique for measuring the discharge current is described in which the large capacitive current across the external electrodes is balanced in a bridge circuit, so that when a discharge occurs the unbalanced voltage, after amplification and rectification, can be used to give a proportionate representation of the value of the discharge current on a cro. For measurement of the growth of the current with time, hf voltage pulses are applied to the discharge tube and the timebase of the cro is synchronized with the pulses. Oscillograms show how the growth of current depends on the pressure ($2-35\mu$) the nature of the gas (H and He), the excess voltage and the frequency of the applied field (10–20 mc). In the initial stage the properties of the tube walls mainly control the multiplication process and thus the starting field. In the later stages the properties of the gas become important and so determine the total time of growth of the discharge. For previous work see 2500 of 1949 (Gill and von Engel).

537.533.8:[546.46–31:548.55 114

Secondary Electron Emission of Crystalline MgO—J. B. Johnson and K. G. McKay. (*Phys. Rev.*, vol. 91, pp. 582–587; Aug. 1, 1953.) Single crystals of MgO in bulk form, cleft along the (100) plane were used. The maximum secondary-emission ratio is about 7 at room temperature and with a bombarding voltage of about 1.2 kv. No correlation between conductivity and yield was established. Increase in temperature led to a decrease in yield. The most probable energy of emission is about 1 ev. Theoretical considerations based on band theory support these observations.

537.56 115

Collisional Effects and the Conduction Current in an Ionized Gas—K. C. Westfold. (*Phil. Mag.*, vol. 44, pp. 712–724; July, 1953.) A first approximation to the transport equation for the conduction current in a binary ionized gas is derived. The collision damping factor derived is a weighted mean of the electron and ion collision frequencies, which is 4/3 times the electron collision frequency. Corresponding approximations made to the equations of conservation, motion and thermal energy, when combined with Maxwell's equations, yield a set of equations for the investigation of the interactions between an ionized gas in motion and the associated radiation field. The results are applicable to the solar atmosphere and the H I, H II and lower ionosphere regions.

538.122 116

The Magnetic Fields Produced by Uniformly Magnetized Ellipsoids of Revolution—H. J. Peake and N. Davy. (*Brit. Jour. appl. Phys.*, vol. 4, pp. 207–209; July, 1953.) Formulas, graphs and tables are given for the numerical evaluation of the field strengths in the equatorial plane and on the magnetic axis of prolate and oblate spheroids.

538.24:548.55 117

The Influence of Domain Structure on the Magnetization Curves of Single Crystals—E. W. Lee. (*Proc. Phys. Soc.*, vol. 66, pp. 623–630; July 1, 1953.) Néel's model for the domain structure of a single crystal of iron in the form of a strip parallel to the (110) direction was used in the calculation of magnetization curves. The I/H curves depend explicitly on the width of the crystal. Comparison between the calculated curve and experimental results for a 3.85 per cent Si/Fe crystal shows good agreement, particularly at low field values.

538.52 118

Induction Phenomena consequent on the Movement of Material in Primary Magnetic Fields, and their Experimental Applications: Part 1—Experimental Bases—H. Hintenegger. (*Acta Phys. austriaca*, vol. 7, pp. 1–13; April, 1953.) Discussion of "unipolar induction," using this term in its broad original meaning, viz., induction of electricity due to movement of material in a magnetic field. Simple experiments demonstrating the basic phenomena are described.

538.521 119

Induction in a Conducting Sheet by a Small Current-Carrying Loop—J. R. Wait. (*Appl. Sci. Res.*, vol. B3, pp. 230–236; 1953.) Analytical solutions are derived for the cases where the loop axis and the plane of the conducting sheet are (a) mutually perpendicular, (b) parallel. The results are applicable to problems of electrical shielding and prospecting.

538.56:517.9 120

An Identity leading to a Solution of Kirchhoff's Problem for Damped Waves—É. Durand. (*Compt. Rend. Acad. Sci., [Paris]*, vol. 237, pp. 647–649; Sept. 28, 1953.)

539.232:537.311.1 121

Electron Currents in Thin Oxide Films on Aluminium—A. Charlesby. (*Proc. Phys. Soc.*, vol. 66, pp. 533–541; July 1, 1953.) At low electric-field intensities, the electron current in films formed on Al in suitable electrolytes depends on the temperature and on the hyperbolic sine of the product of the field strength and a constant dependent on barrier dimensions. The oxide-film barrier height and trough-to-peak width are 0.61 ev and 1.1–1.4 Å respectively. The theoretical current/voltage curves given show good agreement with experimental results. The origin of the electron current and the presence of trapped charges in the oxide layer near the electrolyte surface are discussed.

537:621.3.066.6 122

Radio Research Special Report No. 24. Fundamental Processes of Electrical Contact Phenomena—[Book Notice]—F. L. Jones. Publishers: H. M. Stationery Office, London, 3s, 1953. (*Govt. Publ., [London]*, p. 27; July, 1953.) For a summarized account see *Beama Jour.*, vol. 60, pp. 293–295; Sept., 1953.

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

521.038:537.525 123

Conditions for the Occurrence of Electrical Discharges in Astrophysical Systems—J. W. Dungey. (*Phil. Mag.*, vol. 44, pp. 725–738; July, 1953.) Theoretical considerations of a large ionized mass of gas in motion show that neutral pints of the magnetic field are unstable, so that a small perturbation will start a discharge. The orbits of particles in the field are considered; particles with very large energies

may be involved. Such discharges may account for aurorae and may also occur in solar flares and the interstellar gas.

523.5 124
The Length of Ionized Meteor Trails—L. A. Manning, O. G. Villard, Jr, and A. M. Peterson. (*Trans. Amer. Geophys. Union*, vol. 34, pp. 16–21; Feb., 1953.) The length of the meteor ionization column is defined as the distance between the most widely separated points along the trail from which normal-incidence radio reflections can be obtained, using a radar system of prescribed characteristics. A statistical study is made of the length distribution of trails, based on measurements made at two observation stations spaced 100 km apart and operating with a cw output power of about 1 kw or 23 mc. The mean trail length found was 25–30 m; meteors up to the sixth magnitude were detected.

523.8 621.396.822 125
Cosmic Origin of Radiation at Radio Frequencies—F. Hoyle. (*Nature*, [*London*], vol. 172, pp. 296–297; Aug. 15, 1953.) Discussion of possible ways in which rf radiation from stellar sources may be produced.

523.854:621.396.822 126
Some Fundamental Results and Problems of Radio Astronomy—H. H. Klinger. (*Funk u. Ton*, vol. 7, pp. 359–368; July, 1953.) A clear and concise survey of measurement techniques is given, together with a discussion of the theories of the origin of rf radiation from the sun and the radio stars. The results of measurements of the line emission of interstellar hydrogen on a wavelength of 21 cm are briefly discussed in relation to the problem of the structure of the Milky Way.

537.5– 127
Collisional Effects and the Conduction Current in an Ionized Gas—Westfold. (*See* 115.)

538.7 128
Note on a Proposed Three-Component Aeromagnetometer—V. B. Gerard. (*Z. N. Jour. Sci. Tech.*, vol. 35, pp. 1–3; July, 1953.) By identifying stars photographed with a small nonmagnetic camera mounted on the fluxgate gimbal mechanism of a total-force aeromagnetometer, it is possible to find the magnetic declination and dip.

551.5 621.396.11 129
Conference on Radio Meteorology—(Proc. I.R.E. vol. 41, pp. 1534–1541; Oct., 1953.) Summaries are given of 61 papers for presentation at the conference at the University of Texas, November 1953.

551.510.535 130
Exploring the Ionosphere by Means of Rockets—E. Burgess. (*Electronic Eng.*, vol. 25, pp. 469–470; Nov., 1953.) Summaries are given of papers presented at a conference held at Oxford under the auspices of the Royal Society.

551.510.535 131
The Charge on the Ionosphere—J. A. Chalmers. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 345–346; July, 1953.) The total charge on the inner side of the ionosphere is shown to be zero.

551.510.535 132
Study of the Ionospheric D Region using Partial Reflections—F. F. Gardner and J. L. Pawsey. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 321–344; July, 1953.) 2.28-mc pulse-echo observations were made with a transmitter of peak power ~1 kw and with the receiver located where the noise level was very low. Results showed that below the E layer there were two reflection regions, (*a*) a region of greater electron density around 90 km extending up to the normal E layer, and having minimum ionization at noon, (*b*) a region around 70 km where a distinct ionized layer forms during the day, attains a maximum ionization at noon and disappears at night. In echoes from this region both magneto-ionic components were present, the extraordinary being the stronger. The ratio of the two magneto-ionic components of a discrete echo at a particular range was, within the uncertainty of observation, independent of fading, but varied with time of day and from day to day. From this ratio the collision frequency and electron density for this range of heights were determined.

551.510.535 133
Two Cases of Large F_2-Region Disturbance associated with Small Magnetic Disturbance—B. Hardwick. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 347–349; July, 1953.) During Oct. 26–27, 1947, at Canberra and Brisbane, and, to a lesser extent, at other places in this area of the world, hourly values of f_0F_2 and $h'F_2$ showed departures from monthly median values typical of those observed during a considerable magnetic storm. The days in question were international quiet days. Two other, though less striking, cases have been found from analysis of the records.

551.510.535 134
The Distribution of F_2 Region-Ionization at High Latitudes—J. C. W. Scott. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 289–294; July, 1953.) The linear relation $f_0F_2 = a + bs$ between critical frequency and sunspot number is found to hold with reasonable accuracy in the Arctic. Contour charts of a and b have been prepared in coordinates of latitude and longitude for fixed hours and months. Typical examples are shown, and the general features, particularly the effects of the auroral zone and geographical factors, are pointed out.

551.510.535:523.74 135
Some Statistical Results on the Phenomena associated with Ionospheric Storms—D. W. G. Chappell, L. B. Hainsworth, and I. M. Moorat. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 301–320; July, 1953.) Data covering the years 1941–1949 on (*a*) f_0F_2, (*b*) height of maximum ionization of the F_2 layer, and (*c*) absorption in lower layers, were analyzed to determine the probability that a particular day will be a storm day for a given set of solar conditions. A day is considered to be a storm day if the mean variation of these three parameters studied separately in relation to the effects of sunspots, M regions and solar flares, is such as would cause a 10 per cent fall in muf or a 10 per cent rise in luf. Positive results were obtained only for the parameter f_0F_2. Except for storms associated with M regions at sunspot minimum, the probability that a given day will be a storm day does not exceed 0.5 for the first day of a storm but may reach 0.7 on subsequent days. It is concluded that only during the sunspot minimum period can ionospheric storms be forecast on solar data alone.

551.510.535:523.78 136
The Ionosphere during the Partial Solar Eclipse observed at Buenos Aires, 20th August 1952—I. Ranzi. (*Alta Frequenza*, vol. 22, pp. 59–71; April, 1953.) The decrease in the number of free electrons in the F_1 layer during the eclipse was mainly due to ion recombinations. A simultaneous decrease in the F_2 layer was preceded by an appreciable increase. The two layers were clearly separated, indicating a high recombination coefficient above the F layer. Rapid decrease of the critical frequencies of both layers was observed during the occultation of two sunspots.

551.510.535:537.122 137
Experimental Studies of the Motions of Slow Electrons in Air, with Application to the Ionosphere—Crompton, Huxley, and Sutton. (*See* 106.)

551.510.535:621.396.11 138
Typical Scattering of Radio Waves by Electron Clouds in the Sporadic-E Layer—W. Becker. (*Arch. elekt. Übertragung*, vol. 7, pp. 375–378; Aug., 1953.) The three main types of scattering deduced by Eckersley (2881 of 1940) from fixed-frequency observations are confirmed by specially selected records of swept-frequency echo soundings made at Lindau.

551.578.11 139
The Relation between Rain Current and Rate of Rainfall—A. C. Best. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 285–288; July, 1953.) Consideration of the drop-size distribution in rain (2215 of 1950 and *Quart. J. R. Met. Soc.*, vol. 76, p. 302; 1950.) leads to a relation very similar to the first of Simpson's formulas (*Geophys. Mem.* No. 84).

551.594.2 140
The Effective Separation of Discharging Points—J. A. Chalmers. (*Jour. Atmos. Terr. Phys.*, vol. 3, pp. 346–347; July, 1953.) The effective separation, i.e. that spacing of points in rectangular array which would give the same current density as actually occurs with an earth-connected point in discharges of atmospheric electricity, is 11 m for high field strength.

551.594.5 141
The Aurorae—V. C. A. Ferraro. (*Advances Phys.*, vol. 2, pp. 265–320; July, 1953.) General characteristics of aurorae are described and the connection between aurorae and both solar activity and geomagnetism is shown. The auroral spectrum is discussed and various theories of the aurora are critically reviewed.

551.594.6 142
An Investigation of Whistling Atmospherics—L. R. O. Storey. (*Phil. Trans. A.*, vol. 246, pp. 113–141; July 9, 1953.) A comprehensive report of an an experimental and theoretical study of whistling atmospherics, at frequencies <15 kc is given. Whistlers may or may not be preceded by ordinary atmospherics, produced by lightning strokes at a distance of ~2000 km. The diurnal and annual variations of the properties of both types were investigated. Explanatory theory of their origin advanced by Eckersley (958 of 1935) is developed. Measurements of the degree of dispersion indicate an electron density in the upper atmosphere considerably larger than expected. This result is explained on the assumption that electrons are falling in from outside, and this might account for the relation between the occurrence of whistlers and magnetic activity.

LOCATION AND AIDS TO NAVIGATION

621.396.93 143
Some Practical Measurements of the Relative Performances of a Cyclical Phase-Comparison Type of Direction-Finder and a U-Adcock Instrument—H. G. Hopkins and E. N. Bramley. (*Proc. IEE*, part III, vol. 100, pp. 263–267; Sept., 1953.) Simultaneous bearing observations were made with a cr U-Adcock direction finder and a multi-aerial wide-aperture instrument (PV2) using cyclical differential measurement of phase. The transmitters observed operated in the range 8–12 mc and were located at distances between 750 and 1500 km. In the analysis of the data, where particular attention was paid to the rapidly varying component of the bearing changes, the spread of bearings was, in general, significantly greater with the Adcock direction finder. The superiority of the PV2 instrument was, by inference, due to the relatively large aerial spacing. Although the Adcock direction finder

gave the better performance in the presence of interfering signals, a higher proportion of observations were lost with it owing to deep fades.

621.396.93:621.396.677 144

The Calculation of Wave-Interference Errors on a Direction-Finder employing Cyclical Differential Measurement of Phase—W. C. Bain. (*Proc. IEE*, part III, vol. 100, pp. 253–261; Sept., 1953.) Expressions are derived for the mean-square error due to wave interference as the phase difference between two incident rays varies, and also for the case where the bearing separation of the two rays varies. From the expressions obtained, graphs are constructed showing the variation of rms error (*a*) with system aperture, (*b*) with amplitude ratio of incident waves, (*c*) with bearing separation of rays, (*d*) with angle of elevation of the weaker ray. The corresponding errors for an Adcock direction finder are also plotted for comparison. When angles of elevation of the two rays are the same, apertures greater than 20 λ will be required to give a substantial reduction in these errors but where the angles of elevation differ, apertures of 4 λ will give sufficient error reduction.

621.396.93:621.396.67 145

Effective Polarization on Elevated-Aerial Direction Finders—D. W. G. Byatt. (*Marconi Rev.*, vol. 16, pp. 128–133; 3rd Quarter, 1953.) In general no simple standard-wave error can be quoted for an elevated aerial of U-Adcock or similar type due to the rotation of the plane of polarization. Field conditions at an elevated receiving antenna have been investigated, using a transmitter at different heights operating on 130 mc. For an H-type antenna system at a height 4 λ a signal tends to preserve its original plane of polarization for angles of incidence below ~20 degrees if the reflecting surface is of high conductivity and permittivity the corresponding angle is ~6 degrees.

621.396.93.088 146

The Octantal Error in a Phase-to-Amplitude Conversion Circuit—D. H. Shinn and D. W. Watson. (*Marconi Rev.*, vol. 16, pp. 121–127; 3rd Quarter, 1953.) The two direct currents derived in the conversion circuit for operating the bearing meter of the Marconi vhf df equipment Type AD200, are not exactly proportional to the sine and cosine of the phase angle between the 25-cps input voltages. Analysis shows that the error involved is approximately octantal, depends on the phase difference and the amplitude ratio of the input voltages, and has a maximum value of 3.16 degrees. Operating conditions are chosen such that this error cancels most of that due to antenna spacing.

621.396.933:526 147

The Use of Loran for Survey Work in the Pacific—J. O. Clarke. (*Jour. Inst. Nav.*, vol. 6, pp. 307–312; July, 1953.) The errors likely to occur in loran fixes are discussed. Loran was found particularly useful in survey work between Japan and New Guinea.

621.396.96:534.241 148

Audar—R. H. James. (*Electronic Eng.*, vol. 25, pp. 451–455; Nov., 1953.) An audio system designed to demonstrate radar principles is described. 1-ms pulses of sound of frequency 6 kc are radiated by a highly directional loudspeaker, which also serves as receiver for echo signals. These are rectified and used to provide a conventional cro display of either ppi or range-amplitude type. The loudspeaker is rotated at 3 rpm. A 6-inch metal corner reflector can be detected at 25 feet. Range resolution is about 6 inches and angular resolution a few degrees.

621.396.96:621.372.54 149

Filters for Detection of Small Radar Signals in Clutter—H. Urkowitz. (*Jour. Appl. Phys.*, vol. 24, pp. 1024–1031; Aug., 1953.) Signal /clutter ratio is defined as the ratio of the peak signal to the rms value of the clutter. The synthesis is studied of a linear filter with a frequency characteristic such that the signal/clutter ratio is a maximum at its output; such a filter is inserted before the second detector to improve reception of small signals. The frequency characteristic of the optimum filter is given by the conjugate of the voltage spectrum of the transmitted pulse divided by the power spectrum of the clutter. The influence of such filters on signal/noise ratio and the influence of the second detector on signal/clutter and signal/noise ratios are also discussed.

621.396.962.3:621.396.822 150

Detection of Pulse Signals in Noise—D. G. Tucker and J. W. R. Griffiths. (*Wireless Eng.*, vol. 30, pp. 264–273; Nov., 1953.) Methods described by previous workers are reviewed; the coherent detector, modified rectifier circuits, phase-modulation display, pulse-to-pulse integration, pulse-to-pulse correlation and optimum filtration are discussed. Visual display methods of interest apart from cro methods include the chemical recorder. For conditions where the pulse can just be detected, these methods give better results than the method of extracting the signal envelope by plain rectification.

621.396.963:621.314.7 151

Transistorized Radar Scope Display Unit—R. S. Markowitz. (*Electronics*, vol. 26, pp. 182–183; Oct., 1953.) By substituting four point-contact transistors and a crystal diode for four thermionic valves, an airborne display unit has been produced having a power consumption one tenth that of the standard unit.

621.396.969:527 152

The Use of Radar for Preventing Collisions at Sea—F. J. Wylie. (*Jour. Inst. Nav.*, vol. 6, pp. 271–281; July, 1953. Discussion, pp. 281–293.) Discussion of the problem of applying the international regulations for preventing collisions at sea when one or both ships have radar equipment.

621.396.969:551.594.21 153

The Radar Detection of Dangerous Storms—R. E. Perry. (*Jour. Inst. Nav.*, vol. 6, pp. 238–239; July, 1953.) Technical data are tabulated of equipment in regular use at Entebbe airport. With an operating frequency of 9.32–9.5 kmc and a peak pulse power of 10 kw, moderately large thunderclouds can be detected at ranges up to about 80 miles, while larger storm formations may be detectable at ranges up to 120 miles.

621.396.969.34 154

New Airfield Radar Equipment—(*Wireless World*, vol. 59, pp. 547–548; Nov. 1953.) Brief description of Decca Type-424 equipment, which provides most of the facilities offered by a full-scale G.C.A. The parabolic cylinder-type scanner is 14 feet across and rotates at 24 rpm The magnetron generator gives a peak pulse output of 30 kw at 3.2 cm λ; pulse lengths of 0.1 and 0.5 μsec are available. Maximum range is 25 miles.

MATERIALS AND SUBSIDIARY TECHNIQUES

535.215:546.78 155

The Photoemission of Tungsten in the Region of Predicted Schottky Deviations—E. E. Buder, J. J. Ruddick, and A. H. Weber. (*Phys. Rev.*, vol. 91, pp. 561–566; Aug. 1, 1953.) Experimental evidence has been obtained for the existence of quasi-periodic deviations of the photoemission with variation of the applied electric field, as predicted by Guth and Mullin (2766 of 1941).

535.215.1/.5:621.383.4 156

Infrared Photoconductors—R. A. Smith. (*Advances Phys.*, vol. 2, pp. 321–369, July, 1953.) Work carried out, mainly since 1947, on semiconductor research and cell development for the 1.5–10-μ range, is reviewed and discussed. The group PbS, PbTe, and PbSe is considered in detail and compared with Ge and Si. The photoconductivity of (*a*) compounds involving S, Se or Te, (*b*) intermetallic compounds, and (*c*) the elements Ge, Si, Te, Sn, Sb, is discussed and theory of the photo-effects in the PbS group of substances is presented.

535.215.5:537.29 157

Interpretation of Photoconductivity Experiments at High Field Strengths—D. Curie. (*Compt. Rend. Acad. Sci.* [*Paris*], vol. 237, pp. 791–794; Oct. 12, 1953.) Strong photocurrents observed by Kallmann and Kramer (438 of 1952) with ZnS phosphors are discussed. The theoretical study of electroluminescence shows that, even at field strengths much lower than those causing breakdown, "avalanches" of electrons are produced due to ionization at donor levels.

535.37:546.86 158

Research on Luminescent Antimony Oxide—J. Janin and R. Bernard. (*Compt. Rend. Acad. Sci.* [*Paris*], vol. 237, pp. 798–800, Oct. 12, 1953.) Properties of Sb_2O_2 with and without Mn are investigated.

537.226.2 159

The Dielectric Constant of a Liquid containing Spherical Particles—S. H. M. El-Sabeh and J. B. Hasted. (*Proc. Phys. Soc.*, vol 65, pp. 611–612; July 1, 1953.) The results of dielectric-constant determinations at wavelengths of 1.26–9.2 cm for 5 per cent and 10 per cent solutions of lustrex latex (a suspension of spherical particles of polystyrene in dilute soap solution) agree with Lewin's theory (2139 of 1947).

537.311.31 160

Increased Electrical Resistance and Thermoelectric Power due to Lacunae in Metals—F. Abelès. (*Compt. Rend. Acad. Sci.* [*Paris*], vol. 237, pp. 796–798; Oct. 13, 1953.) The changes of resistance and thermoelectric power due to lattice defects are evaluated in terms of electron moments and impurity concentrations, using a simple model according to which the metal consists of an electron gas obeying Fermi-Dirac statistics.

537.311.33 161

Semiconductor Statistics—P. T. Landsberg. (*Proc. Phys. Soc.*, vol. 66, pp. 662–663; July 1, 1953.) The theoretical derivation of the number of electrons in impurity centers by the free-energy argument, or equivalent, is to be preferred to the approach using the Fermi-Dirac distribution function. See also 18 of 1953 and 1960 of 1953 (Guggenheim).

537.311.33 162

Current Carrier Mobility Ratio in Semiconductors—L. P. Hunter. (*Phys. Rev.*, vol. 91, pp. 579–581; Aug. 1, 1953.) A method is proposed for measuring the ratio of electron mobility to hole mobility directly, using a single sample of either *p*- or *n*-type conductivity. The logarithm of resistivity is plotted as a function of the reciprocal of absolute temperature. The approximately linear portions of the two branches of the curve are extrapolated to their intersection. If r is the ratio of the resistivity at the intersection point to the measured resistivity at the same temperature, the quantity $[1/(r-1)-r]$ gives the mobility ratio for *p*-type samples, and its reciprocal the corresponding ratio for *n*-type samples. With reasonable care r can be determined to within about 0.5 per cent.

537.311.33 163

On the Temperature Dependence of the Mobility of Electrical Charge Carriers in Semi-

conductors—E. D. Devyatkova, Yu. P. Maslakovets, L. S. Stilbans, and T. S. Stavitskaya. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 84, pp. 681–682; June 1, 1952.) A new law, $u = AT^{-3}$, for the temperature dependence of the mobility u of charge carriers was found as a result of an investigation of the temperature dependence of the Hall coefficient and the conductivity of PbSe. This law applies in the region from 20 to 500 degrees C, which is above the estimated Debye temperature of ~ -50 degrees C. This result is not in agreement with existing theory of the interaction of electrons with the thermal vibrations of the lattice.

537.3 L.33:534.39 164

Modulation of the Electrical Resistivity of Germanium by a High-Frequency Stationary Acoustic Wave—G. Mayer. (*Jour. Phys. Radium*, vol. 14, p. 492; July/Sept., 1953.) An experiment is described in which a rod of *n*-type Ge is made to vibrate in resonance with a 1.05-mc quartz resonator. When dc is passed through the rod, the output contains a component alternating at the frequency of the vibrations. From the magnitude and phase of this signal deductions can be made regarding the variation of the number of conduction electrons with the applied pressure and the noise spectrum in semiconductors.

537.311.33:538.632 165

Diffusion Currents in the Semiconductor Hall Effect—R. Landauer and J. Swanson. (*Phys. Rev.*, vol. 91, pp. 555–560; Aug. 1, 1953.) When a magnetic field is applied across a semiconductor sample, it tends to deflect all charge carriers in the same direction. Under certain circumstances a gradient in carrier concentration is set up, thus giving rise to diffusion currents, whose magnitude depends on the recombination velocity at the surface and on the lifetime of the excess carriers in the interior. Assuming small magnetic fields, formulas are derived giving the Hall coefficient in the general case. Circular and rectangular cross-sections are considered. The possibility of measuring the lifetime of excess carriers in near-intrinsic semiconductors on the basis of Hall-effect measurements alone is discussed.

537.311.33:546.289 166

***r*-Type Surface Conductivity on *p*-Type Germanium**—W. L. Brown. (*Phys. Rev.*, vol. 91, pp. 518–527; Aug. 1, 1953.) A positive charge on the surface of a *p*-type Ge crystal induces a net negative space charge within the crystal adjacent to the surface. This space charge consists of ionized acceptor atoms and also under certain conditions, of electrons. When electrons are present they provide a layer of *n*-type conductivity below the *p*-type Ge surface. Experimental evidence for the existence of such layers on the *p*-type region of some *n-p-n* transistors is presented, the layer of electrons forming an additional conducting path or channel across the *p*-type material between the two *n*-type ends. The conductance and capacitance of some channels were measured and compared with values derived from simple theory.

537.311.33:546.289 167

Evaporation of Copper from Germanium—G. Finn. (*Phys. Rev.*, vol. 91, pp. 754–755; Aug. 1, 1953.) Experiments show that (*a*) the number of acceptors introduced is greater when heating takes place in a He atmosphere, and (*b*) the evaporation of Cu from a Ge specimen corresponds, within experimental error, to the loss of acceptors as given by Hall-effect measurements.

537.311.33:546.289 168

Oxygen-Induced Surface Conductivity on Germanium—E. N. Clarke. (*Phys. Rev.*, vol. 91, pp. 756–757; Aug. 1, 1953.) Experiments are described, the results of which can be interpreted in terms of the build-up of a thin oxide layer on the Ge surface.

537.311.33:546.289:535.34:535.61-15 169

Injected Absorption in Germanium—A. F. Gibson. (*Proc. Phys. Soc.*, vol. 66, pp. 588–596; July 1, 1953.) The modification of infrared absorption by carrier injection was observed in *n*-type Ge crystals. The effect was investigated theoretically and experimentally as a function of wavelength, injected current, current frequency and other parameters, and good agreement was found between the basic theory and experimental results. The possible use of light beams in place of the collector and emitter contacts in a "light triode," and the design of Ge modulators, are discussed briefly.

537.311.33:546.289:548.55 170

Thermally Induced Acceptors in Single Crystal Germanium—R. A. Logan. (*Phys. Rev.*, vol. 91, pp. 757–758; Aug. 1, 1953.) When precautions were taken to avoid Cu contamination, rapid quenching was still accompanied by formation of acceptor centers throughout the sample, but the cause of formation was not determined.

537.311.33:546.289:621.396.822 171

A Simpler Explanation for the Observed Shot Effect in Germanium Filaments—A. van der Ziel. (*Jour. Appl. Phys.*, vol. 24, p. 1063; Aug. 1953.) An explanation of the effect [2035 of 1953 (Mattson and van der Ziel)] is advanced which does not conflict with Hall-effect observations on such filaments and which is based on the assumption that hole and electron traps play only a minor part. The theory is supported by experimental results.

537.311.33:546.46-31 172

The Electrical Conductivity and Thermoelectric Power of Magnesium Oxide—R. Mansfield. (*Proc. Phys. Soc.*, vol. 66, pp. 612–614; July 1, 1953.) Report of measurements on sintered specimens of MgO which indicate that it is a defect semiconductor.

537.311.33:621.316.86 173

Nonlinear Semiconductor Resistors—F. A. Schwertz and J. J. Mazenko. (*Jour. Appl. Phys.*, vol. 24, pp. 1015–1024; Aug., 1953.) A phenomenological theory for the nonlinear I/V characteristic exhibited by a granular aggregate of SiC is based on the assumption that the resistance is located mainly at the grain contacts. Constants appearing in the equation derived for the characteristic have values depending on the concentration of impurities, the elastic constants and the grain shape. The validity of the equation is supported by experimental results. The theory should apply equally well to aggregate of other semiconductors over the pressure ranges in which the grains behave as elastic bodies.

537.312.8:669.14.018.583 174

The Change of Electrical Resistance of Single Crystals of Transformer Steel in a Magnetic Field—T. D. Zotov and Ya. S. Shur. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 86, pp. 267–269; Sept. 11, 1952. In Russian.) An observed anomalous variation, of opposite sign to that usually obtained with ferromagnetic materials, can be explained on the basis of the electronic and crystallographic parameters of the material.

538.221 175

A Review of New Magnetic Phenomena—R. E. Alley, Jr. (*Bell. Sys. Tech. Jour.*, vol. 32, pp. 1155–1172; Sept., 1953.) A simple account is given of magnetic phenomena as observed in ferrites at frequencies up to uhf. Domain-wall motion, dimensional resonance, ferromagnetic resonance, the Faraday effect and the effect of transverse fields on wave propagation are discussed. Applications noted include the gyrator, a one-way transmission system at uhf, the polarization circulator and the measurement of magnetic field strength. 28 references.

538.221 176

Effect of a Demagnetizing Field on Magnetic Viscosity—R. V. Telesnin and E. F. Kuritsyna. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 84, pp. 477–478; May 21, 1952. In Russian.) Experimental results indicate that magnetic viscosity is a function of the magnetic characteristics depending on the geometry, not on the material, of the sample.

538.221 177

Thermal Effects accompanying Magnetization of a Ferrimagnetic Material—L. F. Bates and N. P. R. Sherry. (*Proc. Phys. Soc.*, vol. 66, pp. 609–610; July 1, 1953.) The large thermal changes which occur in ferroxcube IV when taken through a hysteresis cycle are explained adequately in terms of the magneto-caloric effect of Weiss and Forrer. Domain-boundary phenomena do not appear to make a significant contribution to the energy changes.

538.221 178

Developments in Sintered Magnetic Materials—J. L. Salpeter. (*Jour. Brit. IRE*, vol. 13, pp. 499–575; Nov., 1953.) Reprint. See 3334 of 1953.

538.221 179

Ferromagnetic Domain Walls in Ferroxdure—H. P. J. Wijn. (*Physica*, vol. 19, pp. 555–564; July, 1953.) A study is made of the magnetic spectrum of ferroxdure samples in which the ratio of the contributions of wall displacements and domain rotations to the initial permeability is varied. Evidence is found of a resonance of domain walls.

538.221 180

Investigation of Ferromagnetic Resonance in a CrTe Alloy—T. M. Perekalina. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 84, pp. 475–476; May 21, 1952. In Russian.)

538.221 181

Temperature Dependence of Magnetic Viscosity of Fe-Ni Alloys—E. F. Kuritsyna. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 84, pp. 687–688; June 1, 1952. In Russian.) The relaxation times τ of wires of (*a*) invar (35 per cent Ni), (*b*) hypernic (50 per cent Ni) and (*c*) permalloy (78.5 per cent Ni) were measured at temperatures from 80 to 673 degrees K for a change in the applied field of 0.5 oersted for (*a*) and (b) and 0.1 oersted for (*c*). τ is approximately constant for both (*a*) and (*b*) but increases with temperature in sample (*c*). The results are discussed in relation to Telesnin's laws (2409 of 1951).

538.221:548.55 182

The Preparation in Sheet Form of Large Single Crystals of Silicon-Iron of Predetermined Orientation for Magnetic Purposes—R. G. Martindale and D. A. Langford. (*Proc. IEE*, part II, vol. 100, pp. 417–426; Aug., 1953.

538.221:621.3.042.14 183

The Assessment of Core Material Variability—M. O. Williams. (*A.T.E. Jour.*, vol. 9, pp. 120–133; July, 1953.) The meaning of complex permeability, complex reluctivity and loss parameter of ferromagnetic cores is discussed. Statistical methods used in the assessment of core material variability are described, with examples, and their application to the determination of realistic tolerances and adjustment of design specifications is indicated.

538.221:621.318.2 184

Oxidic Permanent Magnets with Preferred Orientation—H. Fahlenbrach. (*Elektrotech. Z., Edn A*, vol. 74, pp. 388–389; July 1, 1953.) A short report of investigations carried out by the author, in collaboration with W. Heister, on ceramic magnetic materials composed mainly

of $BaO \cdot 6Fe_2O_3$. Results similar to those reported by Rathenau et al. (751 of 1953) were obtained.

538.221:669.862.5.721 185

Gadolinium-Magnesium Alloys with High Gadolinium Content: Preparation and Magnetic Properties—F. Gaume-Mahn. (*Compt. Rend. Acad. Sci.* [*Paris*], vol. 237, pp. 702–704; Oct. 5, 1953.)

538.222 186

Paramagnetic Resonance Absorption in Metals—S. A. Altshuler, V. Ya. Kurenev and S. G. Salikhov. (*Compt. Rend. Acad. Sci.* [*U.R.S.S.*], vol. 84, pp. 677–679; June 1, 1952. In Russian.)

546.23 187

The Supply of Selenium—(*Beama Jour.*, vol. 60, pp. 223–227; July, 1953.) Extracts from "Review of the Technical and Economic Position of Selenium," B.E.A.I.R.A. Technical Report Z/T90, by A. M. Cassie. The supply of Se is discussed and its principal uses are considered. It is recommended that economies in the use of Se be made, particularly by using alternative materials wherever possible, and that a search for as yet untapped sources be made.

546.28+546.289]:532.61 188

The Surface Tension of Liquid Silicon and Germanium—P. H. Keck and W. Van Horn. (*Phys. Rev.*, vol. 91, pp. 512–513; Aug. 1, 1953.) Results of measurements by the drop-weight method gave 600 dynes/cm for Ge and 720 dynes/cm for Si at their freezing points.

546.28+546.289]:539.32 189

Measurement of Elastic Constants at Low Temperatures by Means of Ultrasonic Waves—Data for Silicon and Germanium Single Crystals, and for Fused Silica—H. J. McSkimin. (*Jour. Appl. Phys.*, vol. 24, pp. 988–997; Aug., 1953.)

621.314.634 190

Direct-Currect Characteristics of the Selenium Rectifier: Part 1—Theory of Rectification—M. Tomono. (*Jour. Phys. Soc. Japan*, vol. 8, pp. 477–483; July/Aug., 1953.) A model of a barrier layer, consisting of a *p*-type and an *n*-type semiconductor with an intermediate insulating film, is considered in the development of a new theory of rectification which is valid over a large range of applied voltage.

621.315.612 191

Ceramic and Dielectric Properties of the Stannates—W. W. Coffeen. (*Jour. Amer. Ceram. Soc.*, vol. 36, pp. 207–214; July 1, 1953.) Measurements made on ceramic compounds of the stannates of Zn, Cu, Cd, Fe^{++}, Fe^{+++} and Mn show that they are semiconductors. The stannates of Ba, Ca, Sr, Mg, Bi, Pb, Co and Ni were also investigated. Their dielectric constants at 1 Mc, range from 12 ($SrSnO_3$) to 101 ($NiSnO_3$) and their power factors from near zero ($CaSnO_3$) to 0.0456 ($NiSnO_3$). The use of the stannates in $BaTiO_3$ bodies is discussed briefly.

621.315.612 192

Dielectric Bodies in the Quaternary System $BaTiO_3$-$BaSnO_3$-$SrSnO_3$-$CaSnO_3$—W. W. Coffeen. (*Jour. Amer. Ceram. Soc.*, vol. 36, pp. 215–221; July 1, 1953.) Disks of various compositions in this quaternary system containing from 3 to 60 mole per cent stannate were prepared and their dielectric properties were investigated. Of those containing more than 30 mole per cent stannate only those high in $CaSnO_3$ content were investigated, the others being too refractory for normal commercial use. The effect of stannate addition to $BaTiO_3$ was to decrease the Curie temperature and broaden the Curie peak. Dielectric constant values of 2.3–2.8×10^3 at 1 kc and low positive temperature coefficients up to 55 degrees C were obtained with a 3 mole per cent stannate addition while a 6 mole per cent Ba- Sr- or $CaSnO_3$ addition gave a dielectric constant of $5–6 \times 10^3$ at 1 kc between 25 and 65 degrees C. Negative temperature coefficients were obtained with 10–60 mole per cent stannate additions.

621.315.612.4 193

Measurement of Permittivity of Mixed Barium and Strontium Titanate in 3000-Mc/s Region—H. Iwayanagi. (*Jour. Phys Soc. Japan*, vol. 8, pp. 525–530; July/Aug., 1953.) The permittivity was determined by observing the resonance of a sample inserted in a water-filled cavity resonator or in a waveguide window. The dependence of the permittivity on temperature (20–100 degrees C), frequency (2.2–3.5 kmc), and on the relative proportions of Ba and Sr, was investigated.

621.315.612.4:537.228.1 194

Ferroelectric Properties of Some Crystals—G. A. Smolenski. (*Compt. Acad. Sci.* [*U.R.S.S.*] vol. 85, pp. 985–987; Aug. 11, 1952. In Russian.) The permittivities of $SrTiO_3$ and solid solutions of $CdTiO_3$ investigated, in the region between 2 degrees K and 300 degrees K, exhibit maxima near 60 degrees K; the maximum for $PbTiO_3$ occurs near 500 degrees C. This, and the results of previous investigations (2548 of 1950 and 155 of 1952) show that piezoelectric vibrations in the solid solutions can exist over a temperature range above the Curie point, determined by the internal stresses.

546.431.824–31:621.315.612.4 195

Double Hysteresis Loop of $BaTiO_3$ at the Curie Point—W. J. Merz. (*Phys. Rev.*, vol. 91, pp. 513–517; Aug. 1, 1953.) Application of an electric field to $BaTiO_3$ shifts its Curie temperature upward; hence, when an ac field is applied to a crystal at temperatures around the Curie temperature, the crystal becomes alternately ferroelectric and paraelectric in the course of a single cycle. The hysteresis loops observed when the crystal is in the ferroelectric state are discussed, and the results used to calculate the constants in the free-energy expression proposed by Devonshire (663 of 1950) and Slater (2188 of 1950).

621.315.612.4:548.55 196

Some Electrical Properties of Strontium Titanate—A. Linz, Jr. (*Phys. Rev.*, vol. 91, pp. 753–754; Aug. 1, 1953.) Curves show (*a*) dc conductivity as a function of temperature, and (*b*) permittivity and loss angle as functions of temperature and frequency. The permittivity of the single-crystal specimen is 20 per cent higher than that previously reported for sintered specimens. The activation energy deduced from conductivity is very close to that for slightly reduced TiO_2 crystals.

621.315.612.6:536.7 197

Thermodynamic and Kinetic Properties of Glasses—R. O. Davies and G. O. Jones. (*Advances Phys.*, vol. 2, pp. 370–410; July, 1953.)

621.315.618.015.5:546.226.161-1 198

Dielectric Breakdown of Sulphur Hexafluoride—C. N. Works and T. W. Dakin. (*Elect. Eng., N. Y.*, vol. 72, p. 624; July, 1953.) Summary only. Anomalies in the breakdown-voltage/electrode-spacing and the breakdown-voltage/pressure curves are discussed.

MATHEMATICS

517.512.2 199

Summation of Fourier Series by means of the Laplace Transformation—P. A. Mann. (*Arch. elekt. Übertragung*, vol. 7, pp. 390–392; Aug., 1953.) A method is described which is particularly useful for calculating the response of a linear network to a nonsinusoidal periodic input.

517.525.2 200

The Chebyshev Approximation Method—P. R. Clement. (*Quart. Appl. Math.*, vol. 11, pp. 167–183; July, 1953.) The method is presented in a form useful for physicists and engineers. Properties of Tchebycheff polynomials are discussed in detail. Applications to the theory of aerials, filters and directional couplers are indicated briefly. 31 references.

517.9 201

The Nature of Solutions of a Rayleigh-Type Forced-Vibration Equation with a Large Coefficient of Damping—P. Brock. (*Jour. Appl. Phys.*, vol. 24, pp. 1004–1007; Aug., 1953.) A method is described in which the forced-vibration equation is converted to a form involving a cubic function which is replaced, as an approximation, by a composite function formed of linear elements.

519:53 202

A Set of Principles to Interconnect the Solutions of Physical Systems—G. Kron. (*Jour. Appl. Phys.*, vol. 24, pp. 965–980; Aug. 1953.) Procedure is indicated for solving problems involving large and complicated physical systems by splitting the system into smaller component systems; the method avoids the necessity for solving large numbers of simultaneous equations or finding the inverse of large matrices.

MEASUREMENTS AND TEST GEAR

621.317.3:621.387 203

Neon Tube Measuring Device—H. E. Styles. (*Wireless World*, vol. 59, pp. 549–552; Nov., 1953.) A simple circuit is considered comprising a high resistance R in series with a neon tube, the latter shunted by a capacitance. When a voltage (greater than a threshold value depending on the striking potential of the tube) is applied to the series combination, the tube flashes at a rate proportional to the mean current through R. The arrangement can be used for measurement of high resistance, high voltage, or leakage current at high voltage.

621.317.332 204

Measuring Impedance of High-Frequency Resistors—C. L. Wellard. (*Electronics*, vol. 26, pp. 176–179; Oct., 1953.) Methods suitable for measurements in the frequency range 2–400 mc are discussed. The unknown resistor is connected across a coaxial-type resonator tuned by a micrometer spindle, and the voltage across the resistor is detected and measured. The impedance is determined with an error ± 5 per cent. See also 457 of 1953 (Bady).

621.317.336.029.64:537.52 205

Methods of Measuring the Properties of Ionized Gases at High Frequencies: Part 4—A Null Method of Measuring the Discharge Admittance—L. Gould and S. C. Brown. (*Jour. Appl. Phys.*, vol. 24, pp. 1053–1056 Aug., 1953.) The admittance of a gas discharge is determined by measuring the ratio of the power transmitted through a microwave cavity to the incident power as a function of signal frequency near the resonance frequency of the cavity. Part 3: 463 of 1953.

621.317.373.029.4/.51 206

Phase-Angle Measurements at A.F.—R. C. Moses. (*Radio and Telev. News, Radio-Electronic Eng. Suppl.*, vol. 50, pp. 12–13, 21 July, 1953.) A precision phase-shifter, comprising an RC network in series with the circuit under test, is adjusted to give a net zero phase-shift on the cro screen. From the values of R and C the phase-angle can be calculated with an error < 2 per cent in the range 20 cps–100 kc. The circuit diagram, including component values, of the phase shifter is given.

621.317.4:538.221 207

Measuring Methods for some Properties of Ferroxcube Materials—C. M. van der Burgt, M. Gevers, and H. P. J. Wijn. (*Radio Tech. Dig., Édn. franç.*, vol. 7, pp. 115–135; 1953.) See 3060 of 1953.

621.317.7:061.4(435.3) 208

Instruments and Equipment for the Measurement of Electrical Quantities—W. Hunsinger. (*Z. Ver. disch. Ing.*, vol. 95, pp. 633–640; July 1, 1953.) A survey of new electrical measurement instruments shown at the 1953 Hanover Technical Fair. 132 references are given to papers on instrument design and measurement techniques.

621.317.7:621.372.412 209

A Polarity Indicator for Quartz Crystals—H. L. Hammatt. (*Electronic Eng.*, vol. 25, pp. 464–465; Nov., 1953.) A simple arrangement is described for testing which face of a crystal becomes positive on compression. The circuit uses two thyratrons fed through a common anode resistor, so that when one fires, on receiving a pulse from the crystal, the other is prevented from firing. The construction of a suitable crystal holder is illustrated.

621.317.7:621.397.2:621.396.822 210

Signal-Noise Meter Checks TV Links—R. Moffett. (*Electronics*, vol. 26, pp. 164–165; Oct., 1953.) Description of portable equipment for measuring the signal/noise ratio of studio-to-transmitter links, coaxial lines, etc. The test procedure is to modulate the carrier at 60–300 cps and to balance the signal as received against the signal after passage through an attenuator followed by an amplifier with low-frequency cut-off at 1 kc. Measurements on a 7-kmc relay path are reported.

621.317.7.029.65:535.417 211

The Fabry-Perot Interferometer at Millimetre Wavelengths—W. Culshaw. (*Proc. Phys. Soc.*, vol. 66, pp. 597–608; July 1, 1953.) The design and operation of an interferometer for wavelengths around 8 mm are discussed. The reflectors are constructed from $\lambda/4$ sheets of dielectric. Extremely sharp fringes have been obtained. Measurements of dielectric constants and losses made with the interferometer are in agreement with values obtained by other methods. The possible use of the interferometer for the accurate determination of the velocity of e.m. waves, and also for length measurements, is considered briefly.

621.317.727:621.317.77 212

Oblique Coordinates and the A.C. Potentiometer—J. E. Parton. (*Beama Jour.*, vol. 60, pp. 211–219; July, 1953.) An introduction to the application of oblique (120 degrees) coordinates to three-phase system calculations is given. An oblique-co-ordinate ac potentiometer excited from three-phase mains and equipment for measuring the phase angles of three-phase mains voltages, are described and discussed.

621.317.729 213

Capacitively Coupled Field Mapper—E. O. Gilbert and E. G. Gilbert. (*Elect. Eng., N. Y.*, vol. 72, pp. 600–605; July, 1953.) The mapper uses resistive Teledeltos paper and a coupling capacitor. The coupling capacitor is formed by the resistive paper and a parallel coupling electrode of Al foil. A displacement current through the coupling capacitor sends a current into the paper whose distribution represents Poisson's equation. A null-type measurement circuit and probe for plotting the equipotential lines are required. Circuit problems and errors are considered in detail.

621.317.73 214

A Valve Megohmmeter—M. G. Scroggie. (*Wireless World*, vol. 59, pp. 516–621; Nov., 1953.) A detailed description is given of an arrangement for measuring resistances up to $>10^{13}\Omega$. A known voltage is used to pass current through the unknown resistance in series with a known standard resistance, and the voltage drop across the latter is indicated by means of a two-valve voltmeter using cathode-followed connections.

621.317.738 215

Two Probe-Type [earth-] Capacitance Meters—L. Medina. (*Proc. IRE [Aust.]*, vol. 14, pp. 193–195; Aug., 1953.) The instruments described have a range of 50 pF and operate at ~2 mc. The first is designed for a substitution method of measurement, a calibrated variable capacitor being incorporated in the probe; the other gives a direct reading on a calibrated moving-coil meter, and is intended for use in the mass production of electronic equipment.

621.317.75:[621.3.018.78+621.397.822 216

A Simplified System of Wave Analysis for Production Testing—W. P. Buuck. (*Gen. Radio Exp.*, vol. 28, pp. 1–7; July, 1953.) A practical guide to the operation of the distortion and noise meter Type 1932-A in conjunction with a cro for rapid testing and adjustment of communication equipment.

621.317.755:512.99 217

Study of a Vectorial Analyser—J. van Geen. (*Radio Tech. Dig. Édn. franç.*, vol. 7, pp. 137–144; 1953.) Fuller version of paper noted in 3372 of 1953.

621.317.755:621.316.722.029.42/.43 218

Oscilloscope Voltage Calibrator—A. B. Kaufman. (*Radio and Telev. News, Radio-Electronic Eng. Suppl.*, vol. 50, pp. 15–16, 31; July, 1953.) Details are given of a unit delivering a calibrated square-wave output of up to 50 V peak-to-peak with a 6–120-v rms input in the frequency range 1c/s–20 kc. The regulator valve providing the voltage reference is operated at 105 v and the over-all error is <4 per cent.

621.317.755:621.317.61 219

A Locus-Diagram Oscillograph for Audio Frequencies—W. Dietrich. (*Funk. u. Ton*, vol. 7, pp. 405–413; Aug., 1953.) The oscillograph, designed for operation in the range from 50 cps to 20 kc, is described, and circuit diagrams, with component values, are given for the principal units. A phase-shifter circuit producing a constant shift of 90 degrees is incorporated. Measurements on two- and four-pole networks can be made to within ±4 per cent phase angles to within ±5 degrees.

621.317.755:621.317.74.018.782.4 220

An Experimental Delay Distortion Scanner—W. J. Albersheim. (*Bell Lab. Rec.*, vol. 31, pp. 256–261; July, 1953.) See 2286 of 1952 (Hunt and Albersheim).

621.317.76 221

The Periodmeter—R. G. Barker and G. L. Connon. (*Wireless Eng.*, vol. 30, pp. 274–275; Nov., 1953.) The instrument described measures instantaneously the period of each cycle of an input signal, and displays the information on a cr tube as the height of the spot above a base line, which may be offset from the screen. Pulses derived from the input signal by limiting, differentiating and delaying are used to actuate the vertical-scan generator, while a separate timebase is used for the horizontal scan. In a particular model, full-scale deflection can be obtained for a change of 60 cps in the frequency of a signal of mean frequency 10 kc. The arrangement can be applied to the detection of pulse signals in noise (see 150 above).

621.385.001.4:534.1 222

Acceleration Effects on Electron Tubes—F. W. Stubner. (*Bell Sys. Tech. Jour.*, vol. 32, pp. 1203–1229; Sept., 1953.) Instruments for simulating mechanical shock and vibration and for the measurement of their effects on valves are described; accelerometers particularly are discussed. The mechanical design requirements of reliable valves are examined and several U.S.A. test reports and specifications are noted.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

534.88:526.956.5:537.528 223

Practical Application of the Under-Water Spark as Sound-Pulse Generator for Echo Sounding—H. H. Rust and H. Drubba. (*Z. angew. Phys.*, vol. 5, pp. 251–252; July, 1953.) Sparks between sintered metal blocks (10 mm × 8 mm cross-section) 0.3 mm apart were used to produce sound pulses, of duration about 1 ms, which were directed through the bottom plates of the research ship *Gauss* of the German Hydrographical Institute. An available 30-kc magnetostriction-type hydrophone was used to receive echoes from the sea bed. A typical echogram is given for a depth of about 35 m. For spark-produced pulses a different type of hydrophone would probably be more suitable than that actually used.

621.3.029.63/.64:001.891 224

Applications of Microwaves in Scientific Research—H. H. Klinger. (*Jour. Franklin Inst.*, vol. 256, pp. 129–146; Aug., 1953.) An outline is given of the uses of waveguides, cavity resonators, transit-time tubes, etc., in spectroscopy, biology, astrophysics, physical optics and nuclear physics.

621.38.001.8:786.6 225

Simulating Piano Tones Electronically—L. Katz. (*Electronics*, vol. 26, pp. 155–157; Oct., 1953.) Description of an instrument in which phase-shift oscillators and distortion amplifiers are used to simulate the tonal effects of the piano.

621.384.622.2 226

The 15 Million Electron-Volt Linear Electron Accelerator for Harwell—C. F. Bareford and M. G. Kelliher. (*Philips Tech. Rev.*, vol. 15, pp. 1–26; July, 1953.) An illustrated account of the design and construction. The corrugated waveguide is ~6 m long and is divided into two equal sections. A 3-kmc, 1.8-mw-pulse-power magnetron is used as the power source.

621.385.833 227

Focusing of High-Energy Particles by Grid Lenses: Part 2—Lens Errors—M. Y. Bernard. (*Jour. Phys. Radium*, vol. 14, pp. 451–458; July/Sept., 1953.) Part 1: 3387 of 1953.

621.387.424 228

Measurement of the Dead Time of a G-M Counter and of the Secondary Emission from the Cathode by the Method of Delayed Coincidences—E. Picard and A. Rogozinski. (*Jour. Phys. Radium*, vol. 14, pp. 445–450; July/Sept., 1953.)

PROPAGATION OF WAVES

621.396.11:535.42 229

On the Theory of Diffraction of the Electromagnetic Wave by Mountains—K. Furutsu. (*Jour. Phys. Soc. Japan*, vol. 8, pp. 500–524; July/Aug., 1953.) Equations for the field strength of the diffracted wave are derived for the case when the change in the surface curvature of the mountain is small over a distance of one wavelength. The use of vector notation simplifies the calculations. A general field-strength equation, taking into account free-space propagation and reflected or diffracted waves, is obtained.

621.396.11:551.5 230

Conference on Radio Meteorology—(Proc. I.R.E., vol. 41, pp. 1534–1541; Oct., 1953.) Summaries are given of 61 papers for presentation at the conference at the University of Texas, November 1953.

621.396.11:551.510.535:621.317.353 231

Alternative Developments of the Theory of Radio-Wave Interaction—L. G. H. Huxley. (*Proc. Roy. Soc. A.*, vol. 218, pp. 520–536; July 22, 1953.) The new expression obtained by

Crompton, Huxley and Sutton (106 above) for ΔQ, the average energy lost by an electron in a collision, is applied to the development of an alternative theory of radio-wave interaction. Application of this theory gives results in better agreement with experimental results (1220 of 1950) than is the case when the hitherto accepted expression for ΔQ is used.

621.396.11.029.45:551.594.6 232

A Study of Individual Radio Atmospherics Received Simultaneously at Two Places—P. W. A. Bowe. (*Phil. Mag.*, vol. 44, pp. 833–840; Aug., 1953.) An extension of earlier observations (1760 of 1951) by the simultaneous recording of atmospherics at Cambridge and Aberdeen is reported. In addition to the equipment used previously, a radio direction-finder was used to locate the sources of the atmospherics. The values of the propagation factors for 1000 km for the four frequencies studied show that the attenuation at 0.6 kc is less than at 3.5 kc. The results for 7.5 and 5 kc appear to indicate that attenuation is less in winter than in summer.

621.396.11.029.64 233

Oceanic Duct and its Effect on Microwave Propagation—L. J. Anderson and E. E. Gossard. (*Nature* [*London*], vol. 172, pp. 298–300; Aug. 15, 1953.) The results of field strength measurements in the *X* and *S* microwave bands were plotted against *X*, a criterion for trapping, expressed in terms of meteorological parameters. Good agreement was obtained between the observed and predicted trends for field strength at wind speeds higher and lower than 15 mph and for positive and negative air-sea temperature differences.

621.396.81 234

Radio Propagation Survey—C. B. Wooster and F. R. Tew. (*A.T.E. Jour.*, vol. 9, pp. 74–79; April, 1953.) The importance of surveys as a preliminary to the establishment of a multichannel radio system is indicated. Path attenuation and its measurement are discussed. A mobile survey unit is illustrated.

621.396.81.029.5:621.3.012.3 235

Ground-Wave Propagation Curves for Frequencies from 150 kc/s to 10 Mc/s—G. Millington and J. C. Thackray. (*Marconi Rev.*, vol. 16, pp. 109–120; 3rd Quarter, 1953.) The curves given were used in revising those prepared by the C.C.I.R. Field-strength contours for different distances are shown on a logarithmic frequency scale, field strength being plotted in db relative to 1μV/m. A set of five graphs covers five values of earth conductivity, including that for propagation over sea. The method of constructing the curves is described.

621.396.812.3.029.64 236

A Statistical Study of Selective Fading of Super-high Frequency Radio Signals—R. L. Kaylor. (*Bell Sys. Tech. Jour.*, vol. 32, pp. 1187–1202; Sept., 1953.) The results of a statistical analysis of 50,000 path-loss/frequency records, made over a period of two months, are shown graphically. Measurements were made over a 30.8-mile path, using a frequency sweep from 3.75 to 4.15 kmc and a single frequency of 4.19 kmc. The ground reflection coefficient over the path was <0.1. Deep fading, caused by multipath transmission, is frequency selective; deep selective fading is ordinarily accompanied by a 6–10-db signal depression over a band of several hundred megacycles; correlation between the fading of two signals decreases with the frequency separation. Hence frequency diversity offers a practical means of avoiding fading effects if a sufficiently large frequency separation is chosen.

RECEPTION

621.376.332.029.6 237

Linear Discriminator for Ultra-high Frequencies—H. Familier. (*Ann. Radioelect.*, vol. 8, pp. 211–221; July, 1953.) The operation of a hybrid-T discriminator circuit for a microwave relay is analyzed and the harmonic distortion introduced by an AM component in the klystron input is calculated. Thermal expansion of the waveguide may cause an unwanted phase shift far greater than that due to carrier-frequency instability. Methods of adjustment for distortion correction are discussed and variations of the basic discriminator circuit are reviewed.

621.396.62:621.396.96 238

On the Interference of Pulse Trains—K. S. Miller and R. J. Schwarz. (*Jour. Appl. Phys.*, vol. 24, pp. 1032–1036; Aug., 1953.) Problems arising from the undesired overlapping, or time coincidence, of pulses from separate pulse trains are considered. The coincidence time fraction is determined for two pulse trains with (*a*) a fixed initial phase and (*b*) a randomly varying initial phase. The coincidence time fraction is then determined for these cases if only pulse coincidences equalling or exceeding a specified interval are considered. Some problems involving more than two pulse trains are briefly discussed.

621.396.621:621.376.56 239

Decoding Processes for Pulse-Code Modulation—K. Steffenhagen. (*NachrTech*, vol. 3, pp. 310–314; July, 1953.) Theory is given of decoding by means of a series arrangement of parallel *RC* and *RLC* circuits.

621.396.621:621.396.822 240

The Reception of Weak Amplitude-Modulated Signals with Linear Detection—G. Fontanellaz. (*Tech. Mitt. schweiz. Telegr.-Teleph-Verw.*, vol. 31, pp. 177–181; July 1, 1953. In German.) Analyses presented by Rice (2169 of 1945 and back references) and Goldman (2007 of 1949) are applied in deriving an approximate formula for the signal/noise ratio of weak signals, in terms of carrier amplitude, modulation depth, lf and hf bandwidth, and receiver noise factor.

621.396.621:621.396.822:621.395.625.3 241

The Synchronous Magnetic Recorder and its Applications: Part 1—Theory—G. Suryan. (*Jour. Indian Inst. Sci.*, vol. 35, section A, pp. 193–203; July, 1953.) The problem of detecting weak signals in the presence of noise is discussed. The theory of the drum recorder is given. By rotating the drum at a frequency which is equal to or a simple fraction of the periodicity of the wanted signal, and using a long narrow pick-up head, an improved signal/noise ratio is obtained. This ratio is evaluated and the transmission characteristics of the recorder are considered. A list of applications is given.

621.396.621:621.396.822:621.395.625.3 242

The Synchronous Magnetic Recorder and its Applications: Part 2—Experimental—G. Suryan. (*Jour. Indian Inst. Sci.*, vol. 35, section A, pp. 205–214; July, 1953.) The design of the recorder is described and its performance is discussed and illustrated by records of nuclear resonance signals.

621.396.82.029.53:621.327.43 243

Radio Interference from Fluorescent Lamps in the Medium-Wave Band—W. Brunhart, E. Rohner, and L. Soós. (*Bull. schweiz. elektrotech. Ver.*, vol. 44, pp. 617–621; July 11, 1953. In German.) Field-strength measurements were made at 550 kc, 1 mc and 1.4 mc, of the radiation from 40-w fluorescent lamps and cold-cathode lamps supplied by six well-known manufacturers. The spatial field distribution at 550 kc, the relative magnitudes of receiver interference in the range 0.1–30 mc (one lamp only) and the effect of lamp fittings were investigated. An attenuation of up to 20 db was obtained with lamps fitted with a metal-mesh screen.

STATIONS AND COMMUNICATION SYSTEMS

621.372.015.3 244

Typical Transients associated with Fundamental Transmission Functions—A. Walther and J. Dorr. (*Arch. elekt. Übertragung*, vol. 7, pp. 379–386; Aug., 1953.) The transient response is calculated and shown graphically for pulse step, square-wave and sine-squared signals, for fundamental transmission functions selected so that numerical values can be obtained by means of Bessel functions. Deterioration of the waveforms results from the repetition of the signals in long-distance systems, unless the phase characteristic is straightened by addition of an all-pass system. Phase-characteristic straightening can only be used to a limited extent for reducing distortion, because it gives rise to precursory signals.

621.376.018.78:621.396.826 245

Harmonic Distortion due to Echo Signals. Conditions for an Amplitude-Modulated Wave—E. G. Hamer and R. G. Medhurst. (*Wireless Eng.*, vol. 30, pp. 276–280; Nov., 1953.) Approximate formulas given previously [2890 of 1952 (Hamer)] are modified to take account of cases where either the ratio of echo to main signal is not small or the modulation depth m approaches unity. Maximum values of second-harmonic distortion occur when $m=0.57$, and of third-harmonic distortion when $m=0.91$. Comparison is made between fm and AM conditions; for fm exact solutions can be obtained.

621.376.3 245

A Note on the Armstrong System of Frequency Modulation—N. N. Biswas and S. K. Chatterjee. (*Jour. Indian Inst. Sci.*, vol. 35, section B, pp. 119–124; July, 1953.) The degree of amplitude modulation produced is determined as a function of the modulating frequency and modulating voltage, using Wheeler's "zero frequency-carrier" concept (3297 of 194). The relation between the degree of amplitude modulation and the percentage of harmonic distortion is shown graphically.

621.376.3:[621.395.43+621.396.41 247

Echo Distortion in the F.M. Transmission of Frequency-Division Multiplex—R. G. Medhurst. (PROC. I.R.E., vol. 41, pp. 1520–1521; Oct., 1953.) Discussion and extension of the work reported by Albersheim and Schäfer (2022 of 1952). Please note change in U.D.C. number.

621.376.3:621.396.41 248

Calculation of the Spectrum for Multichannel Directional Radio Links with Frequency Modulation—G. Bosse. (*Frequenz*, vol. 7, pp. 239–244; Aug., 1953.) Systems in which the radio carrier is modulated by the output from a multichannel telephony system are considered. To simplify the calculation, the modulating voltage is taken as equivalent to a suitably limited Gaussian noise voltage, an approximation which improves as the number of channels increases. A general method is given for calculating the spectrum of an oscillation frequency modulated by a noise voltage, and explicit formulae are derived for the case where phase deviation is the same at all modulating frequencies. Spectral energy distribution is shown graphically for usual values of phase deviation.

621.376.5 249

Some Aspects of the Spectrum of Modulated Pulses—V. N. Rao. (*Jour. Indian Inst. Sci.*, vol. 35, section B, pp. 125–136; July, 1953.) Fitch's method (2619 of 1948) of deriving the spectrum of modulated pulses is discussed and a modified method is described. Two distinct types of spectrum for pfm and pphm systems are found, the one accompanied and the other unaccompanied by pwm. These are compared with the cw frequency and cw phase modulation spectra. The distortion with

different methods of demodulation is examined and curves for the harmonic distortion in the case of pwm are given.

621.376.54 250
Pulse Duration Modulation—H. T. Peretko. (*Radio and Telev. News, Radio-Electroni Eng. Suppl.*, vol. 50, pp. 8–11, 27; July, 1953.) Pwm characteristics are discussed, and a circuit diagram and details are given of an experimental modulator and demodulator. The pulse repetition rate is 10 sec and the modulation is applied to the trailing edge.

621.39:394.4 251
The Provision of Communications for the Coronation of Her Majesty Queen Elizabeth II—S. M. E. Rousell, E. B. M. Beaumont, and B. H. Moore. (*P. O. Elect. Eng. Jour.*, vol. 46, part 2, pp. 71–78; July, 1953.) An account of the extensive network of communications set up by the Post Office to cover the organization, control, broadcasting, television and press reporting of the coronation ceremony and the associated festivities.

621.38.001.11 252
Information Theory—A. M. Andrew. (*Electronic Eng.*, vol. 25, pp. 471–475; Nov., 1953.) "The concept of information as a measurable quantity is explained and the unit of information is defined. Communication over a noisy channel is considered and Shannon's expression for the capacity of a continuous channel perturbed by white Gaussian noise is quoted and its implications discussed. The connection between information and entropy is outlined. The theory has applications in the evaluation of the efficiencies of modulation methods and in connection with nervous transmission, reaction time and sensory prosthesis."

621.39.001.11 253
The Application of Information Theory to Data-Transmission Systems, and the Possible Use of Binary Coding to Increase Channel Capacity—J. F. Coales. (*Proc. IEE*, part III, vol. 100, pp. 291–302; Sept., 1953.) Provided a small delay for coding and decoding and some loss of fine structure in the received signal can be accepted, some of the time wasted when the input signal is changing at a rate less than the maximum can be used to integrate changes in input signal over a given interval; the integrated change is then transmitted in the following interval. In this way, the bandwidth can be reduced. Alternatively, the time saved can be used for the transmission of other quantities. The number of metrons transmitted can be increased by a factor of the order of 4 for an increase of proper scale-unit of less than 50 per cent. By limiting the total change in input signal that can be transmitted in one interval, a further saving can be effected.

621.39.001.11 254
On the Capacity of a Discrete Channel: Part 1—S. Muroga. (*Jour. Phys. Soc. Japan*, vol. 8, pp. 484–494; July/Aug., 1953.) An expression is derived for the capacity of a communication channel disturbed by noise. Numerical examples are given.

621.395+621.397.24:621.315.212.4 255
The Coaxial Cable of the New Italian Communication Network—Monelli. (*See* 13.)

621.396.61/.62.029.62:621.376.3](436) 256
V.H.F.-F.M. Broadcasting in Austria—O. Koton. (*Radio Tech., Vienna*, vol. 29, pp. 228–230; July, 1953.) Technical details of a 1-kw fm transmitter and superheterodyne fm receiver, for operation in the 88–108-mc band, are given. Conventional circuits are used. A rising characteristic for the higher audio frequencies (+13 db at 15 kc) gives an improved signal/noise ratio of the radiated signal.

621.396.65 257
Path Testing for Microwave Radio Routes—R. D. Campbell. (*Elect. Eng.*, vol. 72, pp. 571–577; July, 1953.) The effect of antenna height on received field strength is discussed and actual results are compared with calculations based on knife-edge diffraction and on smooth-earth theory. Variations in atmospheric refraction are allowed for by assuming a variable effective earth radius. Cw test equipment for the range 3.8–4.2 kmc, and tests made under normal propagation conditions on paths between New York and Atlantic and between Atlantic and New Egypt, are described and discussed. Limitations of the use of cw transmissions for such tests are pointed out.

621.396.932 258
Marine-Radio and New Commercial Radio Equipment of the Nationalized Industry—E. Hüttmann. (*Nachr Tech.*, vol. 3, pp. 290–295; July, 1953.) Technical survey of marine radio equipment made in Eastern Germany.

621.396.932.029.62 259
Survey of the U.S.W. Harbour, Coastal and Waterways Radio Service with Special Reference to Present-Day Problems—W. Kronjäger (*Funk u. Ton*, vol. 7, pp. 419–424; Aug., 1953.)

621.398:621.396.712 260
Remote-Control System for F.M. Broadcasting—A. V. Tidmore. (*Electronics*, vol. 26, pp. 138–142; Oct., 1953.) Details are given of the system in use at the WPPA transmitter at Pottsville, Pa., where the operate-control unit is located at the AM transmitter site 10 miles from the remotely controlled fm transmitter. The control operates over two telephone lines, one of which handles metering arrangements, while the other carries switching pulses and audio tones for tuning the remote transmitter.

SUBSIDIARY APPARATUS

621-526 261
Design Methods for Control Systems—H. E. Weber. (*Z. angew. Math. Phys.*, vol. 4, pp. 233–260; July 15, 1953.) The Nyquist-diagram method of analyzing feedback systems is discussed; stability problems can be simplified by using a method based on the attenuation/frequency and phase/frequency characteristics. The speed control of a dc motor is considered as an example.

621.311.6:621.387 262
Fast-Response High-Current Thyratron Power Supplies for Inductive Loads—J. H. Burnett. (*Elect. Eng., N. Y.*, vol. 72, pp. 627–630; July, 1953.) "Several methods for energizing loads such as fields of large motor generators, solenoids, magnetic clutches, and electromagnets are discussed, including the half-wave circuit with an optimum value of capacitance in parallel with the load. This method provides the fastest response for the equipment required."

621.311.8:621.396.65:621.397.2 263
Standby Generating Plant for the Manchester-Edinburgh Television Radio-Relay Link—R. C. Marshman. (*P. O. Elect. Eng. Jour.*, vol. 46, pp. 89–92; July, 1953.) Description of equipment, driven by an ic engine, which automatically restores the supply voltage at unattended relay stations in case of main failure for more than two seconds. The sets are given daily test runs, remotely controlled from attended stations.

621.355 264
Electric Batteries: Recent Patents—L. Jumau. (*Rev. Gén. Élect.*, vol. 62, pp. 323–314; July, 1953.) Discussion of improvements in Leclanché-type cells, alkaline batteries, cells with dished plates or with bipolar electrodes, HgO cells and Weston standard cells.

TELEVISION AND PHOTOTELEGRAPHY

535.65 265
A Determination of Subjective White under Four Conditions of Adaptation—W. N. Sproson. (*B.B.C. Quart.*, vol. 8, pp. 176–192; Autumn, 1953.) A detailed account is given of experiments to determine the area on the chromaticity diagram which is accepted as white by the statistical average of color-normal observers; the results are important from the point of view of color balance in color television systems. Two methods were used, one involving a tricolorimeter with three projectors associated respectively with red, green and blue filters, and the other involving a single projector with appropriate color filters.

621.397.2:621.311.8:621.396.65 266
Standby Generating Plant for the Manchester-Edinburgh Television Radio-Relay Link—Marshman. (*See* 263.)

621.397.2:621.317.7:621.396.822 267
Signal-Noise Meter Checks TV Links—Moffett. (*See* 210.)

621.397.5 268
16-mm Projector for Full-Storage Operation with an Iconoscope Television Camera—E. C. Fritts. (*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 61, pp. 45–51; July, 1953.) The modifications to a 16-mm projector include a faster pull-down, operating at 24 frames/sec, and a special optical system which, in combination with a rotating-disk shutter, provides adequate illumination within blanking time. Operational facilities are also described. The problem of the 24/30 frames/sec conversion is discussed.

621.397.5 269
1953 (Issue No. 2) S.M.P.T.E. Television Test Films: Operating Instructions—(*Jour. Soc. Mot. Pic. & Telev. Eng.*, vol. 61, pp. 52–58; July, 1953.) The seven test sections are described and their use in the detection of alignment and electrical adjustment errors in receivers is discussed briefly.

621.397.5:535.623/.624 270
American Colour Television—(*Wireless World*, vol. 59, pp. 524–526; Nov., 1953.) A short account of the NTSC compatible system, in which the color signals are transmitted on a subcarrier simultaneously with the brightness signal on the main carrier, the over-all channel width (including sound) being 6 mc.

621.397.5(41):621.397.3 271
An Outline of the British Television System. Part 1: Generating the Picture Waveform—D. Wray. (*P. O. Elect. Eng. Jour.*, vol. 46, part 2, pp. 59–64; July, 1953.) The first of three articles. The general principles applied in the development of the picture signal are explained, the processes of scanning and synchronization are described and also the operation of a television camera. Studio equipment and procedure are noted briefly.

621.397.6 272
Conversion of Television Standards—A. V. Lord. (*B.B.C. Quart.*, vol. 8, pp. 108–124; Summer, 1953.) The principles and operation of a television standards converter are described fully, with particular reference to the equipment used at Cassel for conversion from the 819-line to the 405-line standard. See also 2469 of 1953 and 2902 of 1952.

621.397.6(436) 273
Experimental Television Equipment at the Technische Hochschule, Vienna—R. Just. (*Radio Tech. [Vienna]*, vol. 29, pp. 231–234; July, 1953.) General description of 625-line equipment, with block diagrams of transmitter, camera and receiver.

621.397.61 274

Low-Level-Modulation Vision Transmitters, with Special Reference to the Kirk O'Shotts and Wenvoe Stations—E. McP. Leyton, E. A. Nind, and W. S. Percival. (*Proc. IEE*, part III, vol. 100, pp. 269–285; Sept., 1953. Discussion, pp. 285–290.) The questions of choice of modulation system, and electrical position of the modulated amplifier are discussed. The increase in over-all efficiency obtainable, with consequent reduction in the amount of equipment required and in operating costs, led to the choice of a low-level modulation system. Calculations of tube performance are made, and the design of wide-band RF circuits considered. Triple-tuned coupling circuits, which give a useful increase in over-all bandwidth, were selected. The construction and operation of the Kirk O'Shotts and Wenvoe transmitters, which differ only in operating frequency, are described with block and circuit diagrams. Metal rectifiers were used throughout the power-conversion equipment. See also 568 of 1953.

621.397.62:535.88 275

A Large-Screen Television Projector—J. Haantjes and C. J. van Loon. (*Philips Tech. Rev.*, vol. 15, pp. 27–34; July, 1953.) An illustrated account of a projector designed to give pictures 3+4 m from a cr-tube image 72+96 mm. The cr tube, the optical system and the 50-kv dc power-supply unit are described briefly. The dissipation at the cr-tube screen is 25 w; air-jet cooling is used.

621.397.62:535.88 276

A Method of Increasing the Average Picture Brightness in Projection Television—L. B. Johnson. (*Mullard Tech. Commun.*, vol. 1, pp. 95–103; July, 1953.) "By applying a form of agc in the vision channel, a considerably higher initial setting of the contrast control is permissible, resulting in a substantial increase in the average picture brightness over a complete program."

621.397.621.2:621.73.43 277

A Direct-Drive Line-Timebase Circuit—K. E. Martin. (*Mullard Tech. Commun.*, vol. 1, pp. 99–103; July, 1953.) A line-scan circuit for a rectangular picture tube is described. The deflection coils are connected directly in the anode circuit (190 v) of the output pentode so that only a small autotransformer for the hv supply is required. This supply is derived from the flyback pulse and has an effective source impedance of 6.5 MΩ. A variable-resistance shunt across the boost capacitor provides a simple and effective width control. Constructional details and component specifications are given.

621.397.82:551.510.535 278

Interference to Television via Sporadic E on May 17, 1953—T. W. Bennington. (*B.B.C. Quart.*, vol. 8, pp. 169–175; Autumn, 1953.) Interference from several European stations was observed between about 1300 and 1800 G.M.T., on 45-mc television and the accompanying 41.5-mc sound. Contours are plotted showing the approximate distribution of sporadic E over Europe at hourly intervals from 1000 to 1900 G.M.T. on this day; these distributions account fairly well for the interference experienced. Estimates based on Slough records are given for the percentage of total time during which such interfering propagation is likely to occur.

TRANSMISSION

621.376.32 279

Investigations on the Serrasoid Modulator—H. Schönfelder. (*Funk u. Ton*, vol. 7, pp. 333–341; July, 1953). The theory of operation is given and illustrated by experimental results. See also 1846 of 1953 (Paulsen).

621.376.32:621.315.612.4 280

Capacitor-Modulated Wide-Range F.M. System—M. Apstein and H. H. Wieder. (*Electronics*, vol. 26, pp. 190–192; Oct., 1953.) Operation of the modulator circuit described is based on the voltage dependence of a $BaTiO_3$ capacitor in the oscillator tank circuit. The capacitor is transformer-coupled to the circuit to prevent overheating of the dielectric. Frequency deviations up to 2 per cent are obtained with carrier frequencies of 50–500 mc.

621.396.61:621.396.712 281

Unattended Broadcasting Transmitters—W. J. Morcom. (*Marconi Rev.*, vol. 16, pp. 134–140; 3rd Quarter, 1953.) The best arrangements for reliability of service appear to be: (*a*) for high-power installations, operation of two transmitters in parallel, with automatic switching so that the loss in signal strength due to failure of one transmitter is only 3 db; (*b*) for low-power stations, operation of three transmitters in parallel with no automatic switching and a consequent power loss in the balancing resistance in the event of failure. Suitable combining networks are illustrated. Special features of transmitter design and the functions of monitoring circuits are noted.

TUBES AND THERMIONICS

621.314.632:546.289 282

Some Thermal Properties of Point-Contact Germanium Diodes—J. R. Tillman and J. C. Henderson. (*Phil. Mag.*, vol. 44, pp. 677–696; July, 1953.) The isothermal inverse-voltage /current characteristics of several types of point-contact Ge diodes were determined experimentally. On the assumption, the validity of which is discussed, that these characteristics can be used to determine the temperature of the barrier layer, other thermal characteristics were investigated. The results were compared with theoretical results obtained for a simple model of the diode and qualitative agreement was found. The steady-state V/I curve is in good agreement with the observed characteristic curve and this is taken as a confirmation of the view that turnover is largely the result of self-heating, as also deduced by Bennett and Hunter (1531 of 1951).

621.314.7 283

Transistors: Theory and Application: Part 8—Small-Signal Transistor Operation—A. Coblenz and H. L. Owens. (*Electronics*, vol. 26, pp. 158–163; Oct., 1953.) Various methods of analyzing transistor operation are explained and compared. Power gain and other parameters are discussed. Part 7: 3523 of 1953.

621.314.7.002.2 284

Production Techniques in Transistor Manufacture—J. D. Fahnestock. (*Electronics*, vol. 26, pp. 130–134; Oct., 1953.) Illustrated account of processes used at the R.C.A. factory at Harrison. Temperature and timing details are given for different steps in the preparation of the Ge. A rapid method of testing barrier resistance is described.

621.383.4:[546.817.231+546.817.241 285

Lead-Telluride and Lead-Selenide Infrared Detectors—T. S. Moss. (*Research [London]*, vol. 6, pp. 258–264; July, 1953.) The method of preparation, the construction and properties of PbSe and PbTe photoconductive cells are described.

621.385.001.4:534.1 286

Acceleration Effects on Electron Tubes—Stubner. (*See* 222.)

621.385.029.6 287

Statistical Theory of the Magnetron (Static Condition)—G. Ya. Lyubarski and L. E. Pargamanik. (*Compt. Rend. Acad. Sci. U.R.S.S.*, vol. 84, pp. 491–494; May 21, 1952. In Russian.) A theory is developed by considering an electron cloud as an ideal gas in statistical equilibrium in a cylindrical magnetron with a plane anode, and the results are compared with those obtained from Hull's theory (*Phys. Rev.*, vol. 23, p. 112; 1924). Agreement with experimental results is obtained for the electron-gas temperature and the anode characteristic of the magnetron. For small currents the state of the electron gas and the magnetron working conditions can be determined for a given external magnetic field, applied potential difference and emission current.

621.385.029.6:538.691:530.12 288

Relative Electron Trajectories in the Planar Magnetron—F. Ollendorff. (*Elektrotech. u. Maschinenb.*, vol. 70, pp. 213–216; May 15, 1953.) The equations of motion of the electrons in a space-charge-free planar magnetron are developed from the Lagrangian function of relativistic mechanics, and a nonlinear differential equation of the second order is derived for the electron mass as a function of time; this equation can be solved in closed form by means of elementary functions. The electron motion depends largely on a parameter γ, which in the limiting case of small velocities can be represented as the ratio of the magnetic drift velocity to the velocity of light. Only in the case $\gamma<1$ are electron paths of cycloid type obtained. When $\gamma\geqq 1$ the electron paths differ completely from those typical of the classical magnetron. The various path types to be expected are shown graphically as dependent on the value of γ.

621.385.029.6:621.373.423 289

Investigation of an Interdigital Line used as Anode Circuit for an U.H.F. Magnetron Oscillator—A. Leblond, (*Ann. Radioélect.*, vol. 8, pp. 194–210; July, 1953.) A more detailed description is given of the construction and operation of a special interdigital structure [2187 of 1953 (Leblond et al.)]. In this a cylindrical metal wall (called a "ceiling") lies close to the outer face of the segments so that the em field developed is purely transverse. Field equations are derived from which the upper cut-off wavelength can be calculated. The dimensions of five such structures are given, with their calculated and measured wavelengths, and performance curves of a magnetron operating on 11.2-cm λ are shown.

621.385.832:621.395.625.3 290

Electron-Beam Head for Magnetic Tape Playback—Skellett, Leveridge, and Gratian. (*See* 11.)

MISCELLANEOUS

621.3.017.7 291

Temperature Prediction in Electronic Design—P. J. Selgin and B. K. Hawes. (*Elect. Mfg*, vol. 50, pp. 116–119, 376, 378; Oct., 1952.) Formulas and charts are given which are useful for calculating the temperature of electronic equipment when the power dissipation is known.

621.39:061.4 292

German Radio Show—J. E. Cope. (*Wireless World*, vol. 59, p. 471; Oct., 1953.) A brief note on the Düsseldorf exhibition. High-quality reproduction was a noticeable feature of the radio receivers, many of which were for f.m. Pre-production television receivers were also shown. See also *Fernmeldetech. Z.*, vol. 5, pp. 490–492; Oct., 1953.)

621.39:061.4(435.3) 293

Radio, Television and Electroacoustic Engineering—W. Althans. (*Z. Ver. dtsch. Ing.*, vol 95, pp. 641–646; July 1, 1953.) A survey of new apparatus shown at the 1953 Hanover Technical Fair. 94 references.

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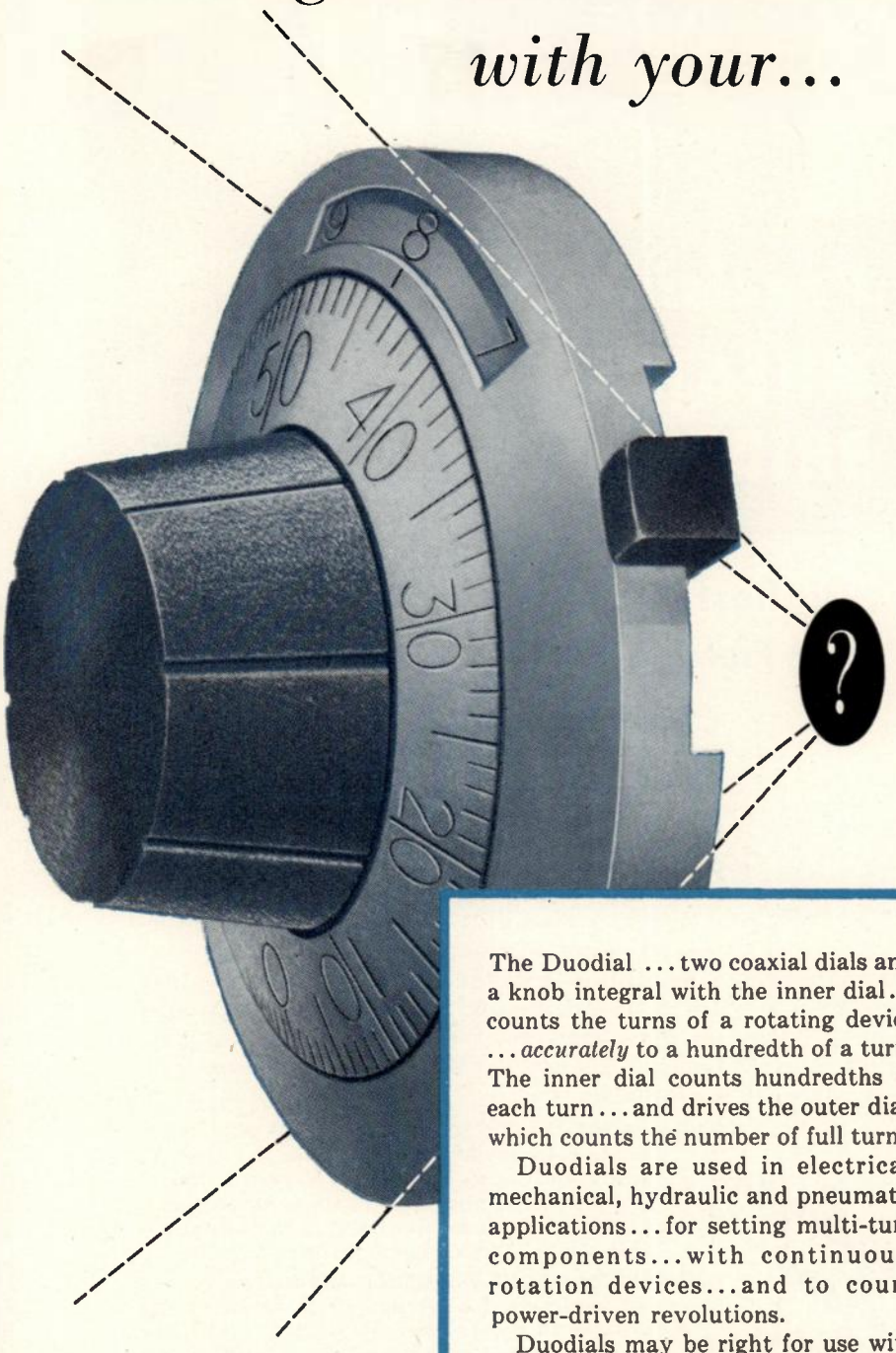

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Shallcross

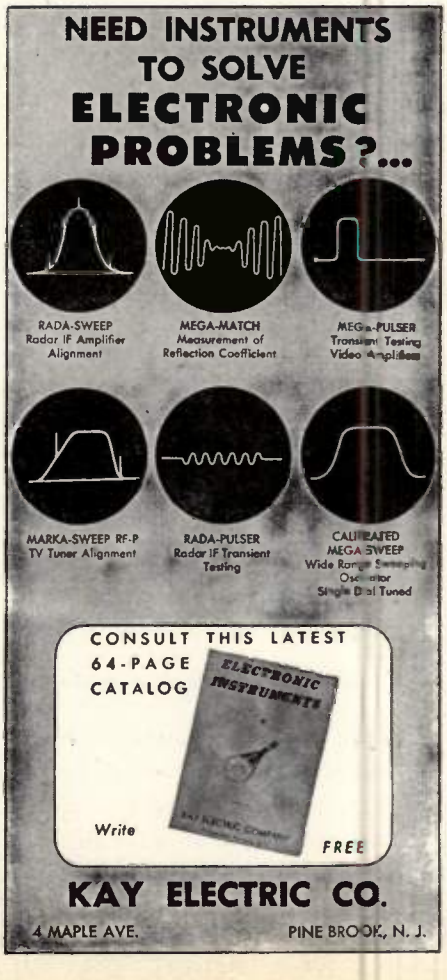
Industrial Engineering Notes

(Continued from page 36A)

TECHNICAL

A new publication, giving specifications on a large number of tubes manufactured in Europe, has been published recently in England, according to information reaching the Department of Commerce. The material is contained in *Comprehensive Valve Guide*, Book 1 and Book 2, which was compiled by Dr. Bernard B. Babani. It is reported to contain information from Dr. Babani's first book, *International Radio Tube Encyclopedia*, plus data on about 3,500 new tubes developed since 1949, including many especially developed in Central Europe on both sides of the Iron Curtain. About 1,000 of the tubes on which information is given reportedly will not be in "active production" until the end of

(Continued on page 68A)



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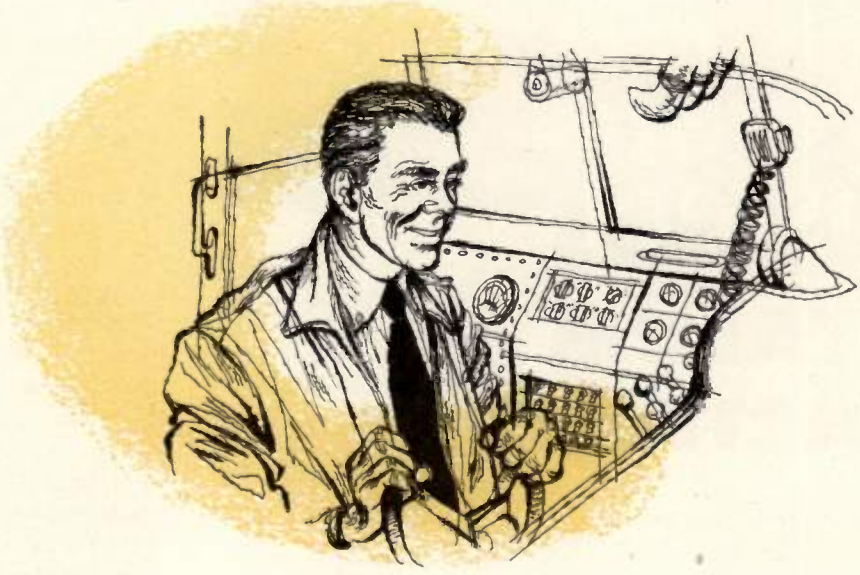

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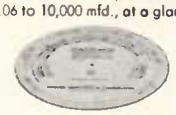

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Industrial Engineering Notes

(Continued from page 66A)

next year. The original "Encyclopedia" and the two new volumes are priced at five shillings each; all three can be obtained in a single cloth-bound volume for 15 shillings. Publication of the new book was reported by the Telecommunications Attaché at the U. S. Embassy in London. . . . **F. B. Lee, Civil Aeronautics Administrator, has announced that greater safety is in prospect this winter at 62 Instrument Landing System locations through use of new "null reference" antenna systems in transmitting glide slope signals.** The improved antenna is being installed as existing instrument landing systems are modernized and at locations where the ILS is being installed for the first time. Modernization of existing equipment includes installation of a new type of transmitter and stand-by equipment. The Commerce Department announced its plans for increasing the number of ILS operating with glide slope from the present 62 to a figure somewhere near 174. One great advantage the "null reference" antenna has over the present system is that it is virtually unaffected by snow. Normally, snow, by changing the level of the ground, causes a bend of approximately one degree on the glide path of the ILS. But with the "null reference" antenna, a snow fall of two feet will result in a change of but one-tenth of a degree in the glide path, the announcement said. . . . **The Office of Technical Services, Department of Commerce, in its November issue of the "Bibliography of Technical Reports," includes several research studies in the electronics field.** The following, of interest to the electronics industry, can be purchased from the Photoduplication Section, Library of Congress, Washington 25, D. C., for the reported price:

(Continued on page 70A)

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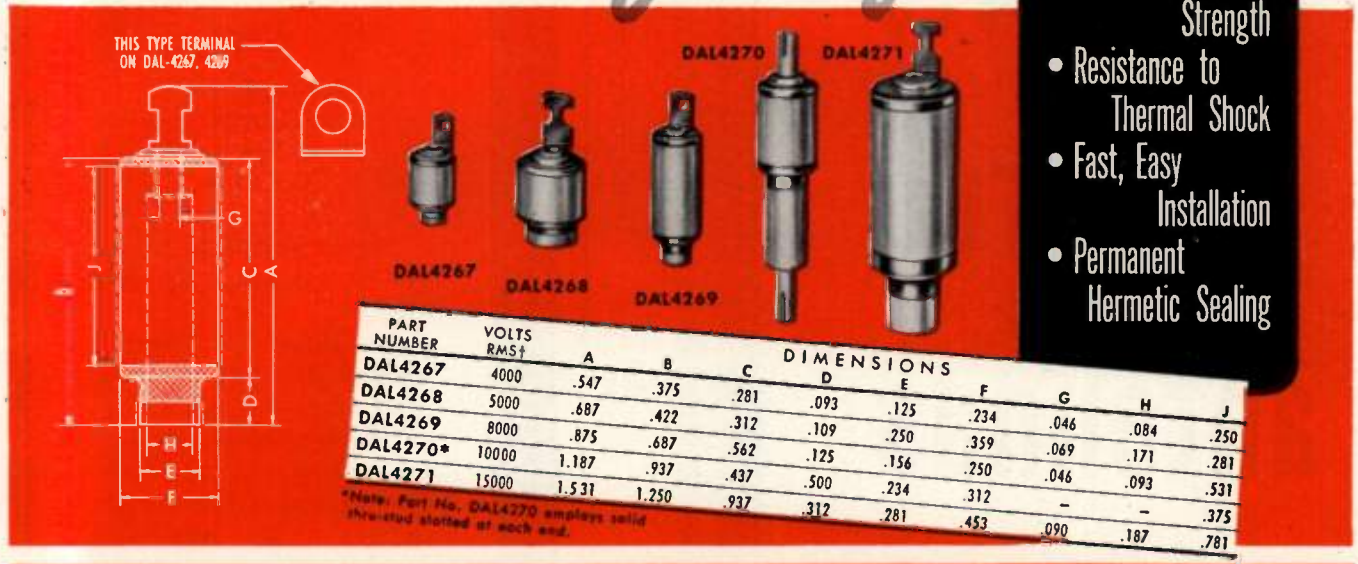
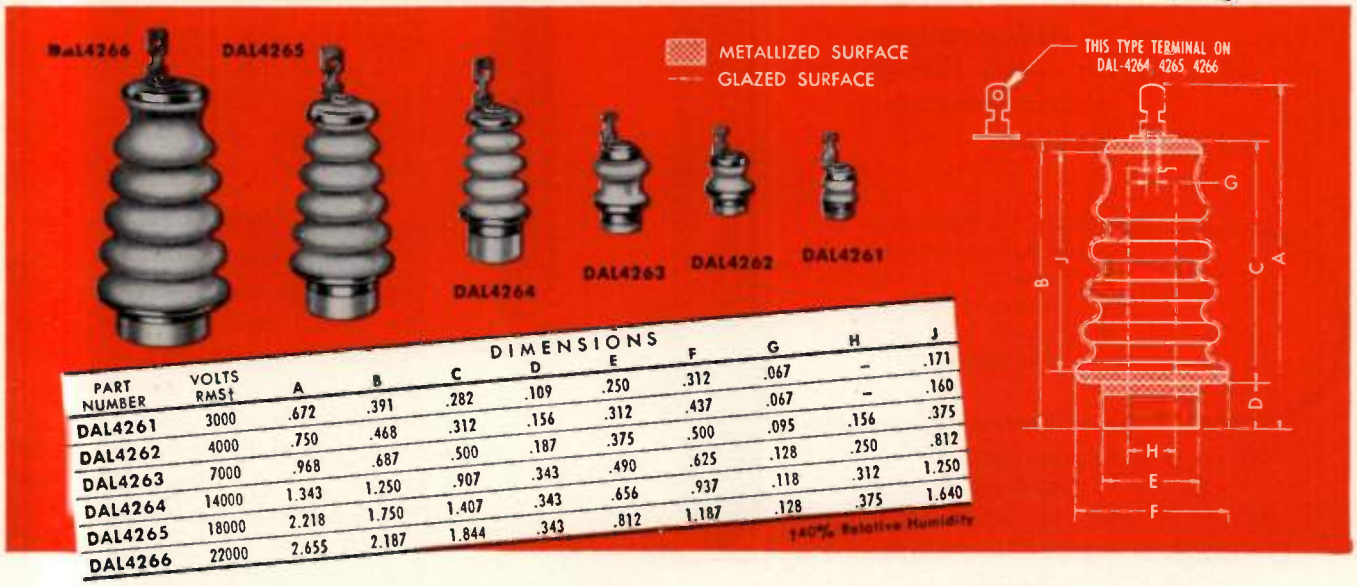


PART NUMBER	VOLTS RMS†	DIMENSIONS A	B	C	D	E	F	G	H	J
DAL4267	4000	.547	.375	.281	.093	.125	.234	.046	.084	.250
DAL4268	5000	.687	.422	.312	.109	.250	.359	.069	.171	.281
DAL4269	8000	.875	.687	.562	.125	.156	.250	.046	.093	.531
DAL4270*	10000	1.187	.937	.437	.500	.234	.312	–	–	.375
DAL4271	15000	1.531	1.250	.937	.312	.281	.453	.090	.187	.781

*Note: Part No. DAL4270 employs solid threaded slotted at each end.



PART NUMBER	VOLTS RMS†	DIMENSIONS A	B	C	D	E	F	G	H	J
DAL4261	3000	.672	.391	.282	.109	.250	.312	.067	–	.171
DAL4262	4000	.750	.468	.312	.156	.312	.437	.067	–	.160
DAL4263	7000	.968	.687	.500	.187	.375	.500	.095	.156	.375
DAL4264	14000	1.343	1.250	.907	.343	.490	.625	.128	.250	.812
DAL4265	18000	2.218	1.750	1.407	.343	.656	.937	.118	.312	1.250
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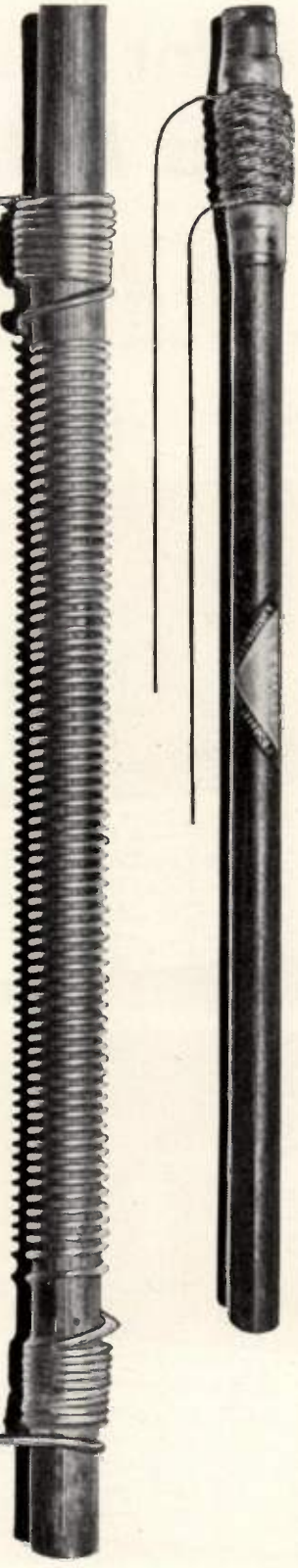

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Industrial Engineering Notes

(Continued from page 68A)

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"Nuclear Quadrupole Hyperfine Structure in Slightly Asymmetric Rotor Molecules" PB 110583, microfilm, $2; photostat, $3.75.

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"Signal Threshold Studies" PB 110769, microfilm, $3; photostat, $10.

"Simplified FM Transient Response" PB 110597, microfilm, $1.25; photostat, $1.25.

"Spectrographic Evidence for the Existence of the Pinch Effect Produced by a High-current Discharge in an Induction-excited Toroid" PB 110823, microfilm, $2.25; photostat, $5.

"Vacuum Fluctuation Effects in Circuits and Electron Streams" PB 110788, microfilm, $2.25; photostat, $5.

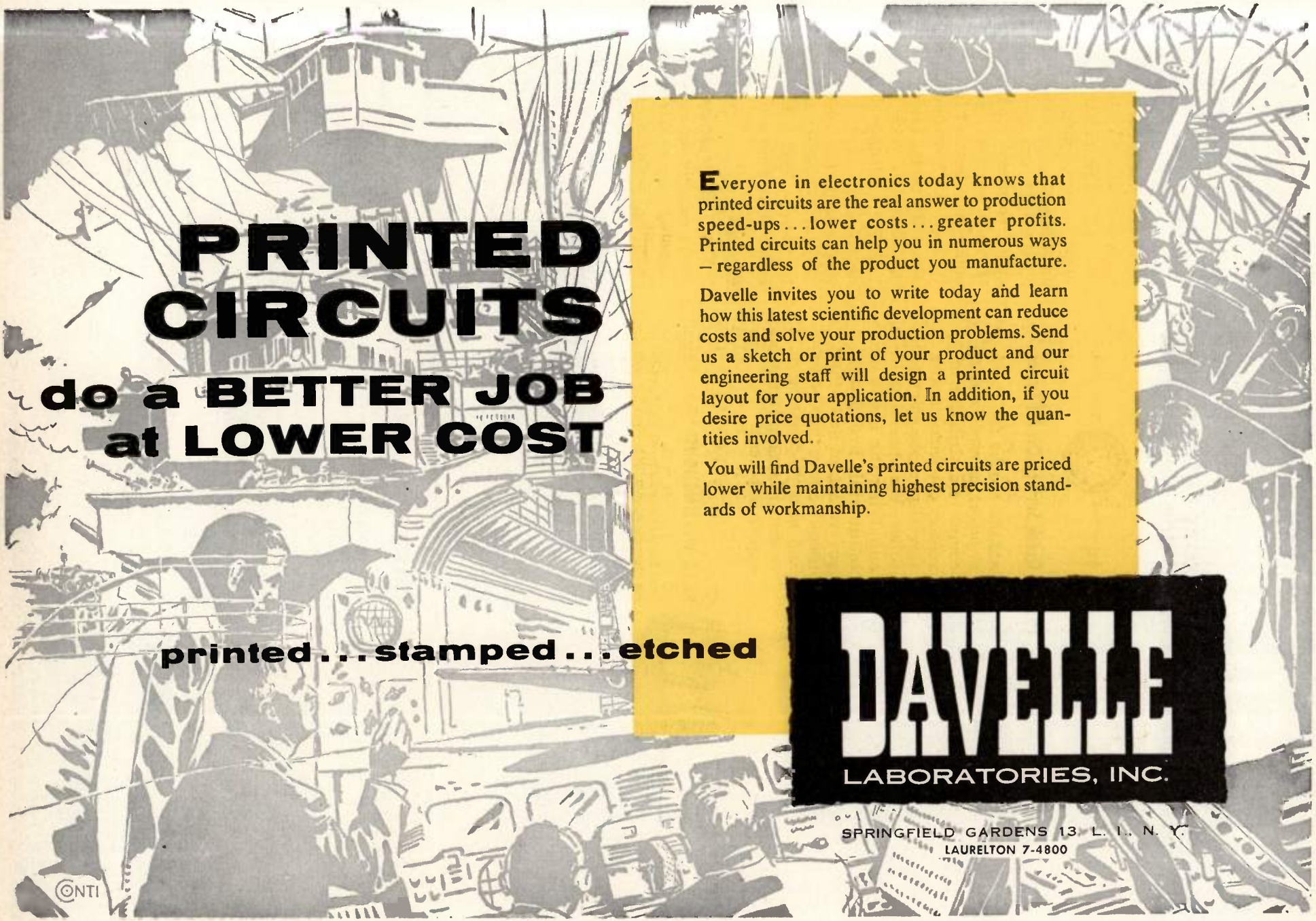
Federal Personnel

The Federal Communications Commission recently announced the consolidation of the Aural Facilities Division and the Television Facilities Division of its Broadcast Bureau, and has designated James E. Barr to be Chief of the new Broadcast Facilities Division of the Commission.

Industry Statistics

Television set production during the first 10 months set a new record for the period, while radio output remained at the highest level since 1950, according to a report from the RETMA Statistical Department. Radio and TV production in October both showed

(Continued on page 72A)



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Davelle invites you to write today and learn how this latest scientific development can reduce costs and solve your production problems. Send us a sketch or print of your product and our engineering staff will design a printed circuit layout for your application. In addition, if you desire price quotations, let us know the quantities involved.

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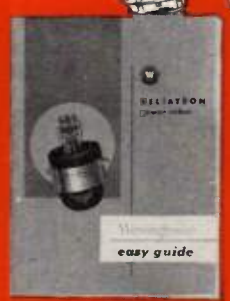

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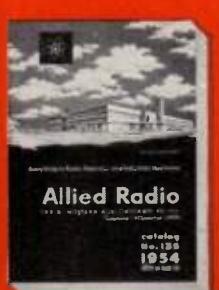

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Industrial Engineering Notes

(Continued from page 70A)

slight declines from the September level, however. During October, the report showed that 680,433 TV receivers and 1,052,493 radios were manufactured. This compares with 770,085 television sets and 1,216,525 radios manufactured in September. Total production in the first 10 months was 6,204,803 TV sets and 11,201,556 radios compared with 3,394,707 television receivers and 8,398,750 radios manufactured in the same 1952 period. TV sets manufactured with UHF tuning facilities during the first 10 months totalled 1,122,507, including 202,605 such receivers produced in October. Radios with FM circuits manufactured in October totaled 19,797. In addition, 5,002 TV receivers with FM facilities were produced.

Foreign

The U. S. Department of Commerce recently warned American manufacturers to use caution in recruiting technical workers in foreign countries for employment in America. It was pointed out that in some cases national laws govern this type of activity while in others local criticism may be voiced. In Germany, the Commerce Department noted, it is against the law for foreign nationals to attempt to recruit skilled labor for employment abroad without first applying for permission to the labor office of the province where the recruiting is to be done. It is also illegal for foreigners to advertise in German newspapers for workers without prior permission of the appropriate labor office.

Television

By Notice of Further Proposed Rule Making, the Federal Communications Commission set forth a revised proposal recently to amend the auxiliary TV broadcast rules in an effort to meet increased demand for TV auxiliary channels as a result of the addition of UHF stations. It is proposed to make

(Continued on page 74A)

FREQUENCY ANALYZER

for measurement of audio frequency phenomena

BRUEL & KJAER MODEL BL-2105

This frequency-selective vacuum tube voltmeter simplifies frequency analysis in all kinds of acoustical, electro-acoustical, and vibration studies. This versatile instrument may be used to:

Measure vibration and noise spectra... giving a spectrogram presentation of frequencies from 47 to 12,000 cycles per second.

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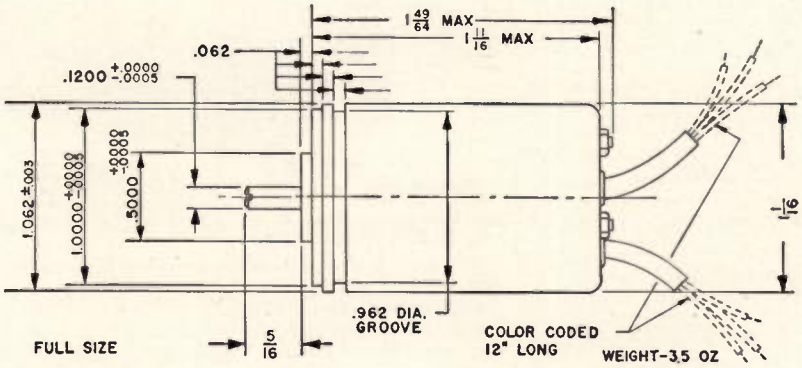
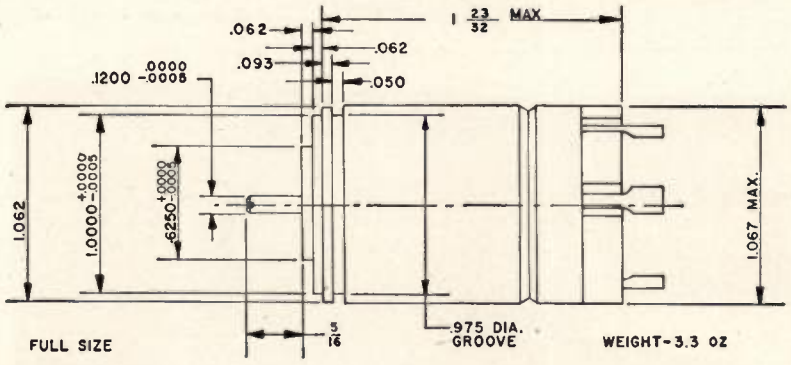


series R200 Control Transformers, Transmitters and Resolvers—maximum error spread 20 minutes of arc. Repeaters and Differentials—maximum error spread 30 minutes of arc. For 26 volt operation. Certain models for 115 volt and with variations in shaft dimensions also available.



series R500 Rugged, environmental-resistant unitized construction with guaranteed maximum error 7 minutes of arc from electrical zero. Average error spreads within 5 minutes. The accuracy of this series eliminates the need for complicated 2-speed Synchro systems in many applications. Advanced manufacturing techniques now make these performance advantages available at minimized cost.

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Industrial Engineering Notes

(Continued from page 72A)

available for TV auxiliary use 10 channels in the bands 1990–2110 mc and 2450–2500 mc, 10 channels in the band 6875–7152 mc, eight channels in the 10500–10700 mc band, and 20 channels in the 12700–13200 mc band. These auxiliary channels would not be assigned to particular broadcast station channels, but would be available for assignment on a case-to-case basis, the Commission said.

Reports from Chapters

Audio

The Houston Chapter of the Professional Group on Audio met on December 10 at the Halliburton Oil Well Survey Co., Houston, Texas, under the chairmanship of L. A. Geddes. Harry Keep, the owner of Gulf Coast Electronics of Houston, presented a paper on "Binaural Sound." The paper was discussed by Chairman Geddes, and following his discussion, the tapes and recordings were demonstrated.

Circuit Theory

The Los Angeles Chapter of the Professional Group on Circuit Theory met on November 12 under the chairmanship of J. A. Aseltine. W. R. Evans, Systems Group Leader, Electro-Mechanical Engineering Department of North American Aviation in Downey, Calif., presented a paper entitled "Further Applications of Root-Locus Methods to Circuit Analysis."

In his paper, Mr. Evans showed that the impedance of any passive network can be expressed in the form $Z=Z_0\times(1-s/q_1)(1-s/q_2)\ldots/(1-s/p_2)\ldots$ in which $Z_0=R$, Ls, or $1/Cs$. If a variable resistance is added across the terminals of the impedance, the characteristic equation will be $Z+R=-1$. A locus of roots of this equation can be sketched in the s plane versus R. One way to obtain the zeros and poles of an impedance is to start with a simple circuit and add circuit elements successively. This procedure shows some limitations on zero and pole locations and shows the effect of each circuit element. Routh's Criterion will be shown to be based upon an equivalent ladder network in which a negative L or negative C occurs for each root having a positive real part. The talk thus is a review of the basic elements of root-locus given last year with expansion of the circuit topics.

(Continued on page 76A)

NEW LOW PRICES FOR G-E ALUMINIZED TUBES

Help You In The Black-and-White TV Market!

EXTRA COST IS HALVED.
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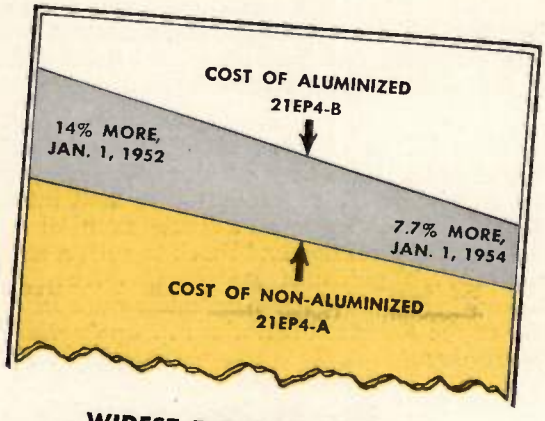
GET THE FIGURES! Phone, wire, or write for latest prices on G-E Aluminized Tubes! Learn how little more it will cost to provide pictures up to 100% brighter—*your* answer to today's stiff TV competition! *Tube Department, General Electric Company, Schenectady 5, N. Y.*

How the cost differential on one popular 21" tube—aluminized vs. non-aluminized—has been cut by General Electric large-scale production economies. Both tubes have come down in price, with the aluminized type leading.



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27EP4	*27RP4	

*Recommended types

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Vibration Control Engineers

PROFESSIONAL GROUP MEETINGS

(Continued from page 74A)

Electron Devices

The San Francisco Chapter of the Professional Group on **Electron Devices** met on December 2 at Stanford University, under the chairmanship of John McCullough, to hear James H. Anderson, Associate Professor at San Jose State College, present a paper entitled "Metallurgical Problems Connected with Vacuum Tube Fabrication."

Electronic Computers

The Philadelphia Chapter of the Professional Group on **Electronic Computers** met on November 17 at the Franklin Institute in Philadelphia, under the chairmanship of John M. Broomall. Omar L. Patterson, of the Sun Oil Physical Laboratory presented a paper on "A Precision Analogue Computer for Predicting Oil Reservoir Behavior." Mr. Patterson first gave some background information on why a computer is needed for oil field production, and also briefly described several rather primitive devices used to assist the oil field engineers. He then described a recently-completed computer of high accuracy that the Sun Oil Company has developed and is using in their oil fields in

(Continued on page 78A)

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- Q-Max is widely accepted as the standard for R-F circuit components because it is chemically engineered for this sole purpose.
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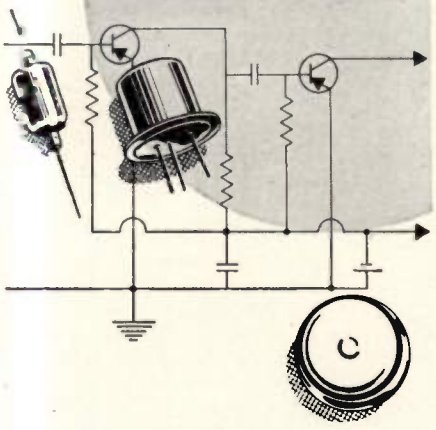

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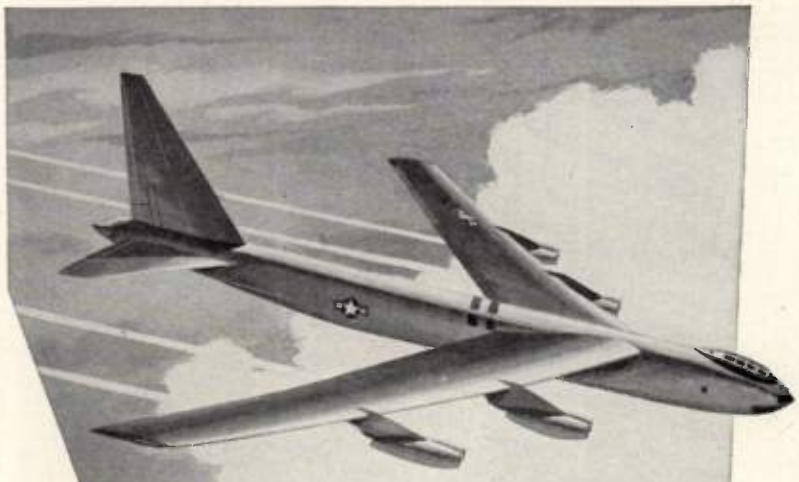

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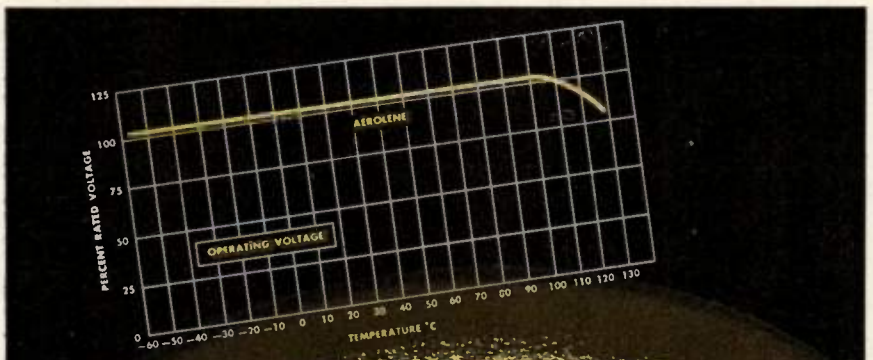

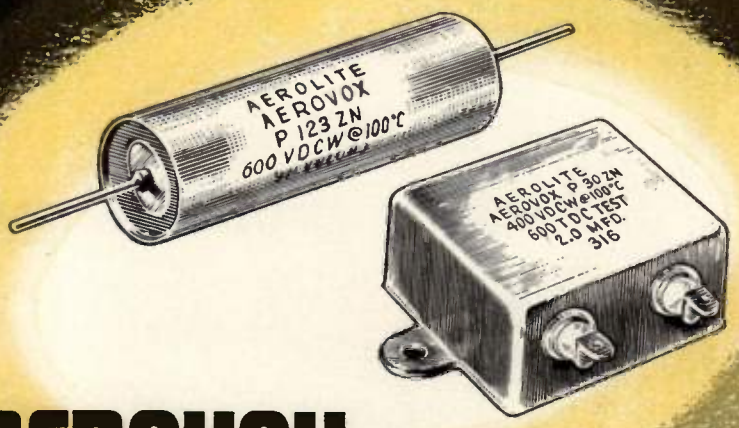

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Professional Group Meetings

(Continued from page 76A)

Texas. The computer consists of an RC section which simulates the pumping characteristics of the oil field. These in conjunction with several function generators and control circuitry form a device which solves the differential diffusion equations and results in the optimum production schedule of the oil field. Computer accuracies of .1 per cent and .02 per cent are claimed. D.C. amplifiers are used at a repetition rate of 250 cycles per second. Mr. Patterson illustrated his talk with slides showing block diagrams and oscilloscope displays. Following the presentation of the paper, there was a discussion period.

Engineering Management

The Washington Chapter of the Professional Group on **Engineering Management** met on November 23 at the Bolling Field Officers Club, Washington, D. C. Allen Schooley, Vice Chairman, was the presiding officer. The speaker was Thomas Meloy, President, Melpar Division, Westinghouse Air Brake Company, who presented a paper entitled "The Conflict of Management and Science." Afterwards, the paper was discussed by the members present.

The Los Angeles Chapter met on December 16 at the IAS Building in Los Angeles to hear Max Clarke of Hughes Aircraft Corp. speak on Methods of Establishing an Engineer's Salary.

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Accuracy Frequency Calibration—1%
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USES:

Examine pulse spectrum of magnetrons and klystrons
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SECTION MEETINGS

ATLANTA

"General Applications of Oscilloscopes," by Lee Maserian, Browning Laboratories; November 20, 1953.

BEAUMONT-PORT ARTHUR

"New Uses of Radio Waves," by A. W. Straiton, University of Texas; December 2, 1953.

BOSTON

"Matching Opportunity with Need," by J. W. McRae, President, Sandia Corporation; October 29, 1953.

"Generalized Sampling of Band-Limited Signals," by A. Kohlenberg, Melpar, Inc.; "Optimum and Ideal Nonlinear Filters," by L. A. Zadeh, Columbia University; November 19, 1953.

BUFFALO-NIAGARA

"Atmospheric Electricity," by Seville Chapman, Cornell Aeronautical Laboratories; November 18, 1953.

"Blood Volume Measurement by the Plethysmograph Method,"—Panel discussion type of meeting moderated by Karl Swartzel; December 10, 1953.

CENTRAL FLORIDA

"Television for the Indian River Area," by S. Ackerman and J. Wynn, Wynack Associates; September 10, 1953.

"Statistics and the Physical Sciences," by R. J. Monroe, Patrick AFB; October 21, 1953.

CINCINNATI

"Travelogue of African Hunting Trip," by John Zettel, Jr., Zettel Memorial Orthoptic Reading Clinic; September 15, 1953.

"The Physics of Hearing and Music," by W. E. Kock, Bell Laboratories; "Across Asia by Jeep," by Gerhard Neumann, General Electric Company; October 22, 1953.

"A High Efficiency, High Quality Audio Amplifier," by A. B. Bereskin, Faculty, University of Cincinnati; November 17, 1953.

CLEVELAND

"Reliability of Electronic Components," by Capt. T. W. Robers, U. S. Navy; November 19, 1953.

COLUMBUS

"Early History of Aeronautical Research," by J. F. Victory, National Advisory Council for Aeronautics; November 12, 1953.

CONNECTICUT VALLEY

"UHF-TV on Mt. Greylock," by Leon Podolsky, Sprague Electric Company; October 15, 1953.

"Radio Astronomy," by J. P. Hagen, Naval Research Laboratory; November 19, 1953.

DALLAS-FORT WORTH

"Our Expanding Technology," by J. W. McRae, President, Sandia Corporation; December 16, 1953.

DENVER

"Industrial Television," by F. D. Meadows, Dage Electronics Corporation; October 16, 1953.

"Ultrasonics, A New Tool in Medical Diagnosis," by W. Roderic Bliss, Denver Research Institute; November 13, 1953.

"The Collins Punched Card Integrated Flight Navigation System," by E. B. Thornley, Collins Radio Company; December 11, 1953.

ELMIRA-CORNING

"Television, Its Role in the Community," by Jack Prewitt, WECT-TV; November 16, 1953.

(Continued on page 82A)

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Input Ratings from 25 to 195 volts AC.
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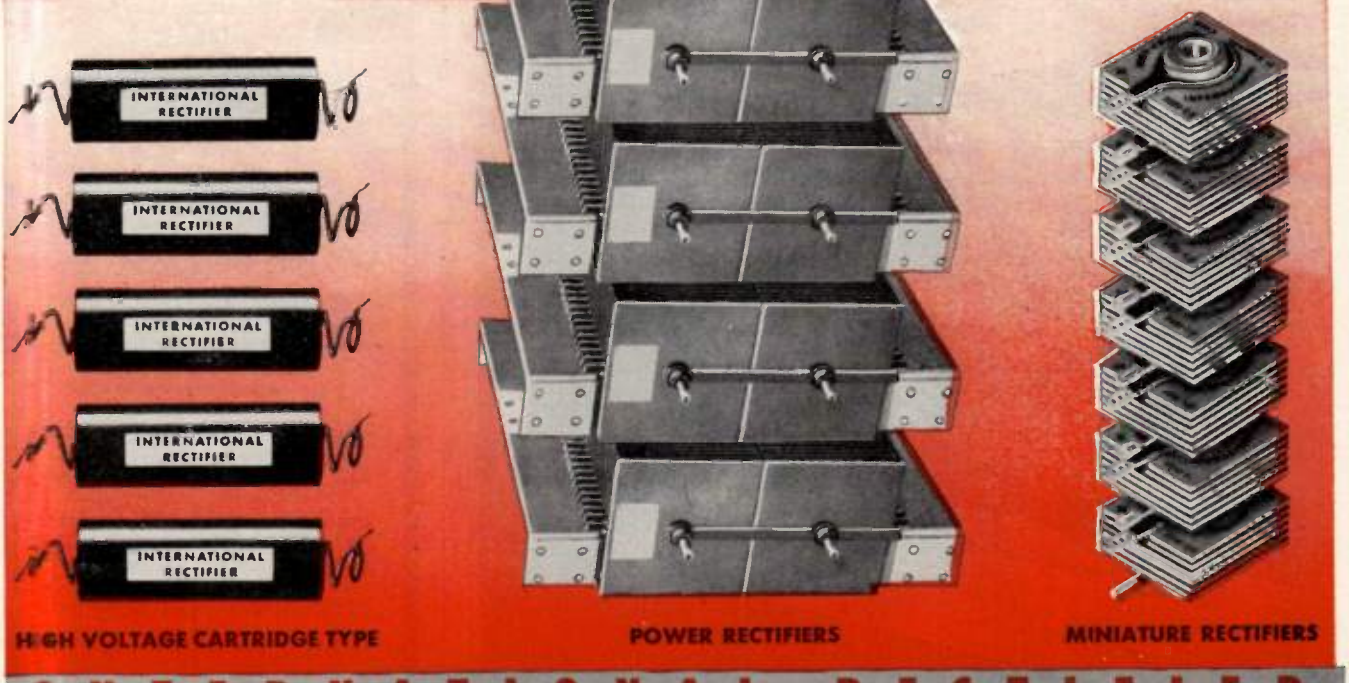

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7 North Avenue, Wakefield, Mass.

SECTION MEETINGS

(Continued from page 80A)

EMPORIUM

"FM/FM Telemetering," by Frank Shaddelman, Raymond Rosen Engineering Prod.; November 24, 1953.

EVANSVILLE-OWENSBORO

"UHF Triode and Measuring Technique Development," by R. L. Bailey, General Electric Company; December 9, 1953.

FORT WAYNE

"The Problems Encountered in Converting to Color TV," by R. N. Harmon, Westinghouse Radio Stations, Inc.; December 3, 1953.

HAMILTON

"Design of U.H.F. Tuners," by J. E. Kreps, Jr., Sarkes Tarzian Company; November 16, 1953.

HUNTSVILLE

"Uses of the Oscillosynchroscope," by Lee Maserian, Browning Laboratories; November 17, 1953.

INDIANAPOLIS

"Survey from the Beginning of UHF Broadcasting to the Present Time," by W. W. Hessler and W. D. Renner, Howard W. Sams Company; September 25, 1953.

"The Principals of Color Television," by C. N. Hoyler, RCA Laboratories Division; October 16, 1953.

"Radio Control of Model Planes and Boats," by V. C. MacNabb, Citizen-Ship Radio Corporation; November 20, 1953.

(Continued on page 84A)

ZOPHAR
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Zophar Waxes, resins and compounds to impregnate, dip, seal, embed, or pot electronic and electrical equipment or components of all types; radio, television, etc. Cold flows from 100°F. to 285°F. Special waxes non-cracking at —76°F. Compounds meeting Government specifications plain or fungus resistant. Let us help you with your engineering problems.

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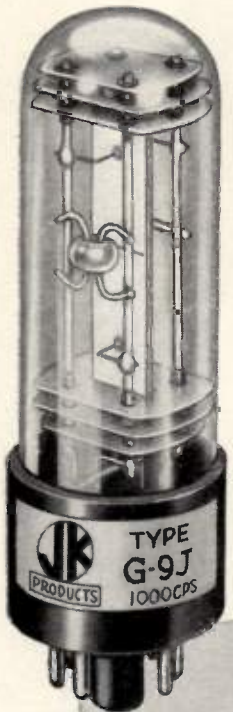

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SECTION MEETINGS

(Continued from page 82A)

LITTLE ROCK

"An Experimental Transitor Personal Broadcast Receiver," by L. E. Barton, RCA Laboratory; and film, "Nerves of the Nation"; December 7, 1953.

LONDON

"Home Music Systems," by R. A. Bruce, Northern Electric Company Ltd.; November 6, 1953.

"Recent Advances in Microwave Scanners," by H. Gruenberg, National Research Council; December 1, 1953.

LONG ISLAND

"21 cm Hydrogen Radiation from Interstellar Space," by H. I. Ewen, Ewen-Knight Corp.; November 16, 1953.

MILWAUKEE

"St. Lawrence Seaway," by Gen. S. D. Sturgis, U. S. Army; November 18, 1953.

MONTREAL

"Some Special Characteristic of Dielectrics," by Louis Kahn, Aerovox Corp.; December 2, 1953.

"Technical Arrangements for the Sound and Television Broadcast of Coronation Ceremonies on June 2, 1953," by M. J. L. Pulling, member of Institution of Electrical Engineers, speaking from London, England. Reply to Mr. Pulling and Description of Arrangements in Canada by J. E. Hayes, C.B.C.; December 3, 1953.

"Comptatible Color Television, the NTSC Standards and its Relation to the Broadcaster," by Dr. G. H. Brown and C. K. Hoyler, RCA Labs.; December 16, 1953.

NEW YORK

"Color Television—Its Challenge to the Engineer," by D. G. Fink, Philco Corp.; December 2, 1953.

OMAHA-LINCOLN

"Integrated Flight System and the VOR Radio Range," by P. A. McKinley, Collins Radio Company; October 29, 1953.

OTTAWA

"Principles and Practice of Electrical Methods of Geophysical Exploration," by J. R. Wait, Defence Research Board; December 3, 1953.

PHILADELPHIA

"The Surface-Barrier Transistor," by W. E. Bradley, Philco Corporation; "Electrochemical Techniques for the Fabrication of Surface-Barrier Transistors," by W. H. Forster, Philco Corporation and "Circuit Applications of Surface-Barrier Transistors," by J. B. Angell, Philco Corporation; December 3, 1953.

PHOENIX

"Data Processing Systems," by R. L. Sink, Consolidated Engineering Corp.; November 16, 1953.

PITTSBURGH

"A Brief Rèsumè of Nuclear Instrumentation," by V. G. Shaw, Westinghouse Electric Corp.; November 9, 1953.

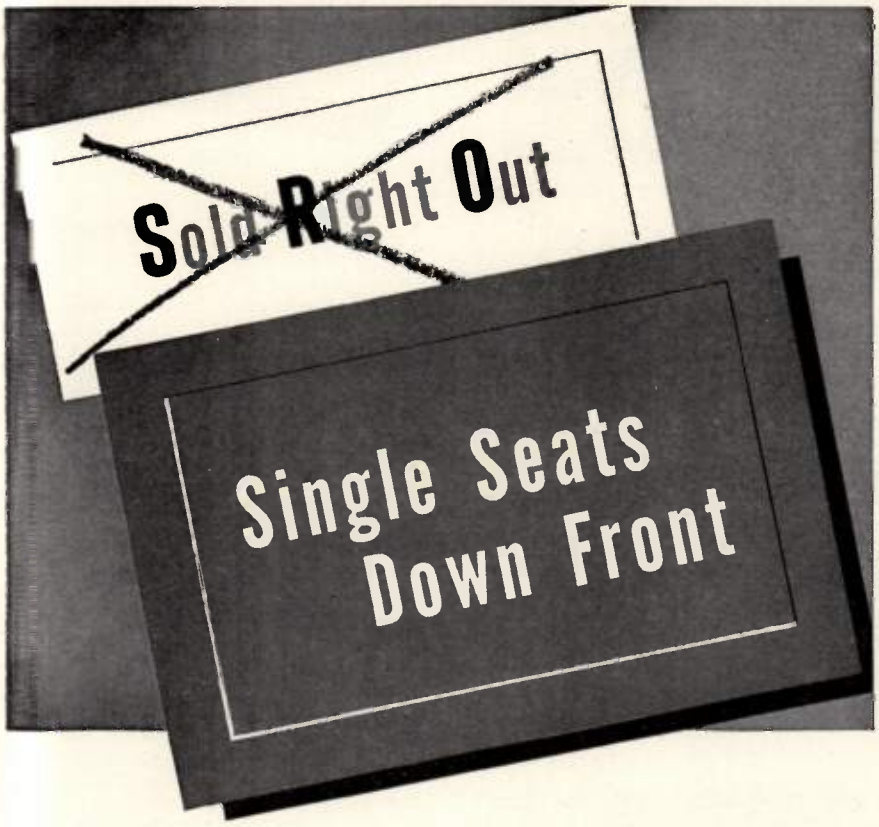
"Information Processing Machines for Science, Business & Industry," by S. N. Alexander, National Bureau of Standards; December 14, 1953.

PORTLAND

"Measurement Problems Involved in the Development of Precision Potentiometers," by E. J. Hovee, Brown Electro-Measurement Corp. and "L-C Meter," by C. H. Moulton, Tektronix, Inc.; November 23, 1953.

PRINCETON

"Rafax Bandwidth Compressor and Airport Traffic Control Recorder," by J. L. McLucas, Haller, Raymond and Brown, Inc.; November 12, 1953. *(Continued on page 86A)*



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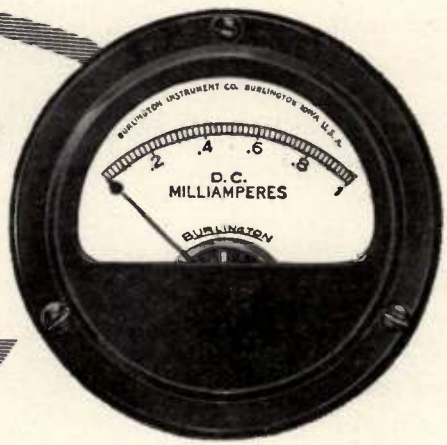

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SECTION MEETINGS

Continued from page 84A)

ROCHESTER

"Electronic Components," by Leon Podolsky, Sprague Electric Company; November 18, 1953.

"Radio and Television Progress," by William Fay, Stromberg-Carlson Company; December 3, 1953.

ROME-UTICA

"Primary Frequency Standards," by F. G. Merrill, Bell Laboratories, and "Organization of Research and Development in Britain," by Arthur Boonham, RADC; December 1, 1953.

SACRAMENTO

"Stereophonic Sound," by Jack Houser, Ampex Corporation; November 19, 1953.

"UHF TV Comes to Sacramento," by Paul Leake, KCCC; December 11, 1953.

SAN ANTONIO

"The Chemistry and Physics of Transistors," by G. K. Teal, Texas Instruments, Inc.; October 29, 1953.

SEATTLE

"Our Expanding Technology," by Dr. J. W. McRae, Westinghouse Electric Company; September 9, 1953.

"Cross Modulation in Traveling Wave Tubes," by A. W. C. Nation and J. W. Christie, Both of University of Washington; October 22, 1953.

"King—TV's New Transmitter and Antenna," by J. L. Middlebrook, KING; November 20, 1953.

SYRACUSE

"Physics of the Ionosphere," by Dr. A. H. Waynick, Faculty, Penn. State; November 16, 1953.

(Continued on page 88A)

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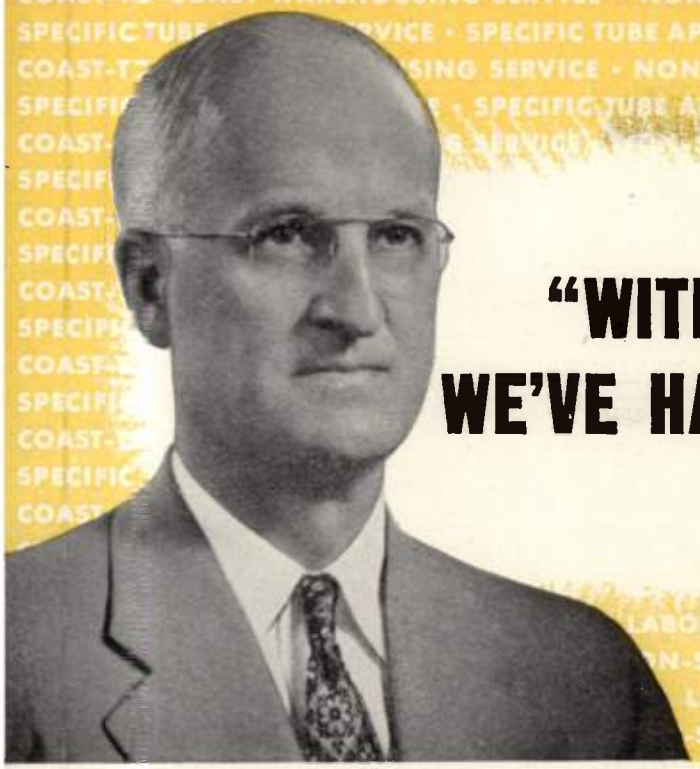

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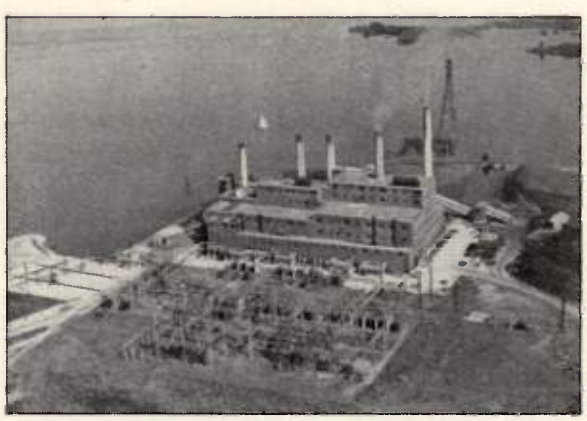

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● For duty at high voltage and high current, the Lapp Gas-Filled Condenser offers a combination of characteristics not available in any other type of capacitor . . . extreme compactness . . . low loss . . . high safety factors . . . puncture-proof operation . . . constant capacitance under temperature variation . . . and reliability of performance assured by a 15-year service record.

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Lapp

(Continued from page 86A)

TOLEDO

"Sterophonic Sound Systems," by Harold Plumador, Strong Electric Corp.; November 2, 1953.

TULSA

"The Design and Manufacture of a UHF Television Converter," by W. F. Stewart. American Radio and Television, Inc.; November 19, 1953.

TWIN CITIES

"Ultra-Sonics and Experimental Medicine," by J. F. Herrick, Mayo Clinic; November 19, 1953.

"Present Day Trends in Automatic Control with Special Reference to Non-Linear Problems," by Dr. R. H. MacMillian, Cambridge University; December 10, 1953.

VANCOUVER

"Description and Discussion of Equipment Comprising the Modern TV Transmitter," by A. A. McKerrell, Marconi's Wireless Telegraph Company; November 16, 1953.

WASHINGTON

"The Pennsylvania Turnpike as a Mass Transportation System," by G. B. Gilbert, Pennsylvania Turnpike Commission, and "Telecommunications Tools Used in the Solution of Technical Problems," by Douglas Lapp, Raymond Rosen Engineering Products Inc.; December 14, 1953.

WINNIPEG

Visit to CBC's new radio and television facilities; November 17, 1953.

SUBSECTIONS

EAST BAY

"Instrumentation for Industry and Research," by Dudley Wright, Wright Engineering; November 17, 1953.

ITHACA

"Focusing Sound Waves with Microwave Lenses," by W. E. Kock, Bell Telephone Labs.; October 30, 1953.

MID-HUDSON

"V.H.F. Impedance Meter (with applications and use)," by F. G. Marble, Boonton Radio Corporation; November 24, 1953.

MONMOUTH

"Magnetic Amplifiers—Operating Principles and Applications," by M. B. Zucchino, SCR; October 21, 1953.

"Advances in Microwave Tubes," by J. R. Pierce, Bell Telephone Labs.; November 17, 1953.

NORTHERN NEW JERSEY

"The Practical Aspects of Home Receiver Design," by R. T. Cappodanno and Walter Lukes, both of Emerson Radio and Phonograph Corp.; October 14, 1953.

PALO ALTO

"Radio Frequency Spectroscopy," by Dr. Martin Packard, Varian Associates; December 3, 1953.

UNIVERSITY OF AKRON, IRE-AIEE BRANCH

"Industrial Applications of the Edison Cell" by W. J. Irvine, Edison Storage Battery Co.; November 24, 1953.

(Continued on page 90A)

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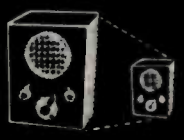

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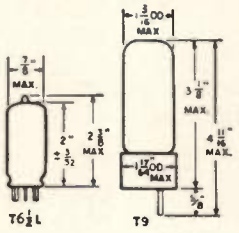

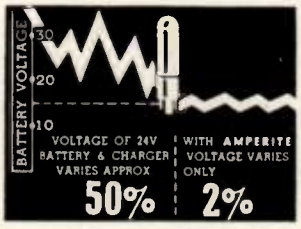

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STUDENT BRANCH MEETINGS

(Continued from page 88A)

UNIVERSITY OF ARIZONA, IRE-AIEE ELANCH
"Future Plans for the Electrical Engineering Dept." by Dr. T. L. Martin, Faculty, University of Arizona; October 15, 1953.

COOPER UNION, IRE-AIEE BRANCH
"Air Navigation" by Mr. Fred Moscowitz Rome Air Development Center; November 23, 1953

UNIVERSITY OF DENVER, IRE-AIEE BRANCH
General Meeting; November 23, 1953
General Meeting; December 11, 1953.

UNIVERSITY OD DETROIT, IRE-AIEE BRANCH
"The Development of Power Apparatus" by Mr. L. Chesnut, General Electric Corp.; October 29, 1953.

DREXEL INSTITUTE OF TECHNOLOGY IRE-AIEE BRANCH
"Student Membership in IRE," by Rev. John A. Klekotka, Faculty, Villanova College; December 1, 1953.

UNIVERSITY OF FLORIDA, IRE-AIEE BRANCH
"The Right and Wrong Way of Interviewing for a Position," by Professor W. S. Wood, Faculty, University of Florida; November 10, 1953, and "What Employers Look for in the Interviews" by Clift Burchsted, Carbide Carbon Co.

GEORGE WASHINGTON UNIVERSITY, IRE BRANCH
"Statistics In Engineering," by Dr. Frank Weida, George Washington Univ.; December 2, 1953.

IOWA STATE COLLEGE, IRE-AIEE BRANCH
Slides on "Electric Power in Greece," by Prof. B. S. Willis, Faculty, Iowa State College; November 4, 1953.

THE STATE UNIVERSITY OF IOWA, IRE BRANCH
Film, "Lightning Flashes"; December 2, 1953
"Impressions of Europe," by Dr. Hunter Rouse, Faculty State University of Iowa; December 9, 1953.
"Iowa's Electronic Scoring Machine," by Mr. R. A. Edberg, State University of Iowa; December 16, 1953.

JOHNS HOPKINS UNIVERSITY, IRE-AIEE BRANCH
General Meting; October 7, 1953.
"Importance of Resusciation," by R. H. Coleman, Consolidated Gas, Electric Light & Power Company of Baltimore; October 14, 1953.
Film, "Bottle Magic," and General Meeting; October 21, 1953.
"Image Amplifier," (film) by Mr. W. S. Lasby, Westinghouse Electric Corp.; October 28, 1953.
"Placement of Hopkins Men," by J. L. Rogers Johns Hopkins University; November 4, 1953.
"Analogue Field Plotter," by Mr. Robert Ehrlinger, Johns Hopkins University; November 11, 1953.
"A is for Atom," (film) and General Meeting; November 18, 1953.

KANSAS STATE COLLEGE, IRE BRANCH
"Carrier Current & Telemetering on Power Systems," by R. E. Phillips, Kansas Gas & Electric Co.; December 10, 1953.

LAFAYETTE COLLEGE, IRE-AIEE BRANCH
"The Transistor," by H. T. Davenport; September 29, 1953.
"History & Organization of IRE-AIEE," by Prof. L. Carnover, Faculty Lafayette College; Films, "Naturally it's F.M." & "The World's Largest Electric Workshop"; October 20, 1953.
"Facsimile," by Walter J. Schlatter, Student, Lafayette College; November 24, 1953.

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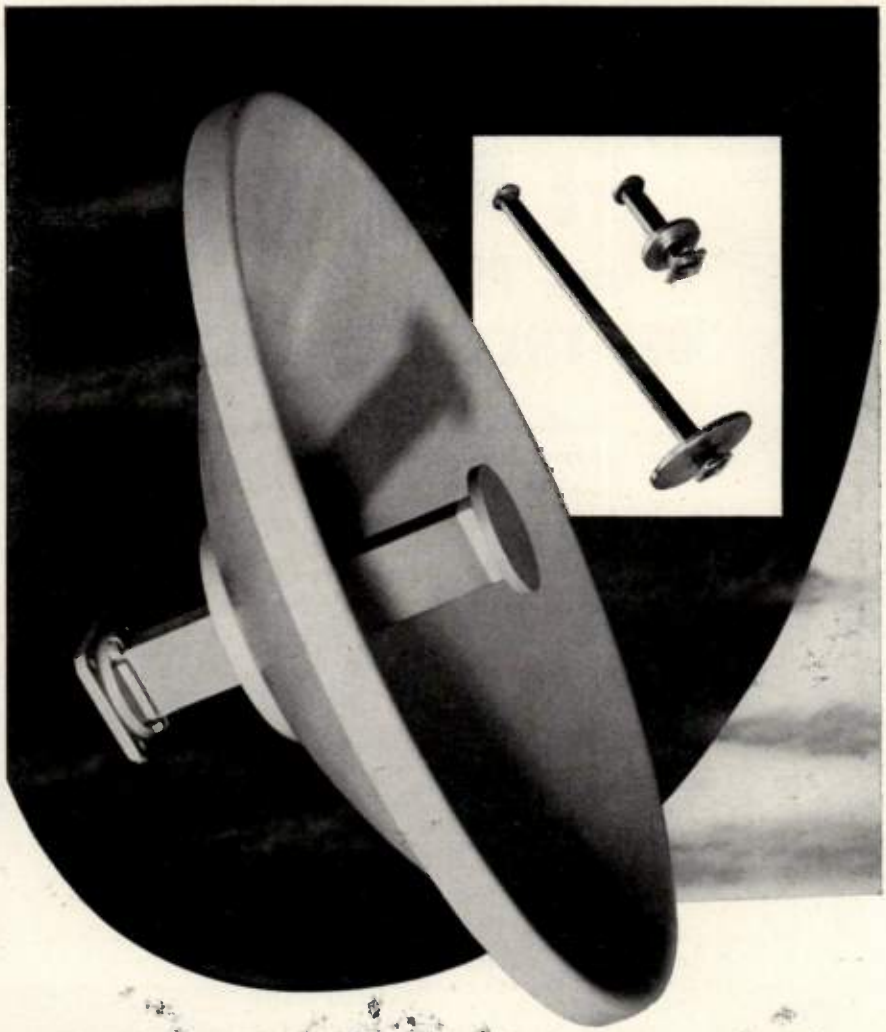

PARABOLIC ANTENNAS *for the* X-band

Quick delivery of parabolic antennas for the X-band is now possible. Designed and manufactured by The Gabriel Laboratories, these antennas meet, or better, required civilian and military specifications. Precision reflectors are illuminated by a modified Gabriel wave guide feed — the same Gabriel design which has received universal recognition in the 7000 mc commercial relay band. Large orders can be filled quickly due to the extensive manufacturing facilities of our affiliate, Gabriel Electronics Division.

These antennas are available with dish sizes of 1, 2, 3, 4, and 6 foot diameters — have a standard three or four point adjustable mounting — and are equipped with a UG-40A/U input flange which is suitable for use in pressurized systems. Feed and dish de-icers are also available for extreme weather conditions.

- Frequency coverage (two ranges) — 8900 to 9300 mc; 9300 to 9750 mc.
- VSWR — less than 1.2 : 1 throughout each range.
- Each antenna can be spot tuned to a specific frequency, with a VSWR of less than 1.05 : 1.

For analysis of your antenna or microwave problem, write or phone NEedham 3-0005.

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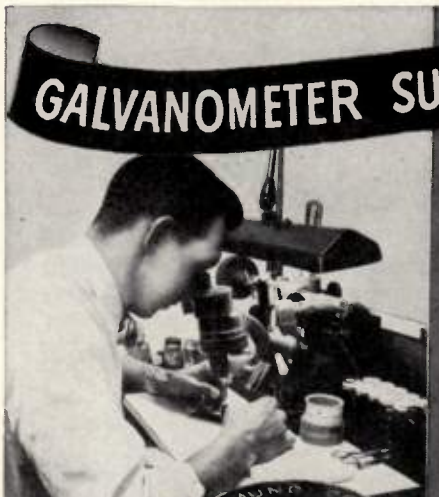

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RIBBON rolled as thin as .0001"

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CONTROL, WITHIN CLOSE LIMITS OF THE FOLLOWING CHARACTERISTICS: Size; Elongation; Tensile Strength; Torsional Properties; Electrical Resistance; Temperature Coefficient of Resistance; Weight per Unit Length; Purity . . . *Write for data and complete list of products.*

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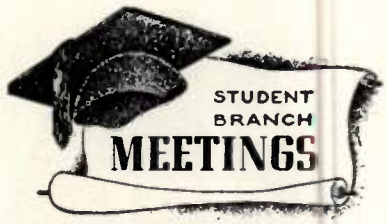

(Continued from page 90A)

LEHIGH UNIVERSITY, IRE BRANCH

"Transistor Physics," by A. D. Rittman, Philco Corporation; November 24, 1953.

"Color Television," by C. N. Hoyler, RCA; December 8, 1953.

UNIVERSITY OF MARYLAND, IRE-AIEE BRANCH

"Demonstration on T.V. Service and Repair," by Jack Star and Bill Hess, Students University of Maryland; November 11, 1953.

"The Electronics of Analog, Digital and Binary Computers," by Mr. Orem Ridgeway, Mathematician at I.B.M.; December 2, 1953.

MICHIGAN COLLEGE OF MINING AND TECHNOLOGY, IRE-AIEE BRANCH

"Electro-Osmotic Process of Solidifying Microscopic Grained Soil," by V. O. York, Faculty Michigan College of Mining and Technology; November 17, 1953.

MICHIGAN STATE COLLEGE, IRE-AIEE BRANCH

General Meeting and Introduction to AIEE and IRE for new members by I. B. Baccus and I. O. Ebert Baccus: Faculty, Michigan State College and Ebert: IRE Representative; October 7, 1953.

"New Developments in The Power Field," by P. L. Chesnut, General Electric; October 28, 1953.

"Dynamometers," by G. Calwell, Westinghouse; November 18, 1953.

MISSISSIPPI STATE COLLEGE, IRE BRANCH

Film, "Power by Which We Live," General Electric; September 17, 1953.

"Shop Accidents and Their Treatment," by Dr. Longis, College Physician; October 8, 1953.

"Effects of High Freq. Radiation on the Human Body," by Paul Hutchison, Professor, Electrical Engineering Dept. and Film "A is for the Atom." General Electric; October 13, 1953.

"Autopilots," by Mr. Lear, Chairman of the Board-Lear Inc.; November 19, 1953.

"Advances in Engineering," by William Seale, General Electric Co.; December 7, 1953.

MISSOURI SCHOOL OF MINES & METALLURGY, IRE-AIEE BRANCH

General Meeting; November 18, 1953.

"Wider Horizons for Engineers," by J. O. Grantham, Philips Petroleum Co.; December 10, 1953.

MONTANA STATE COLLEGE, IRE-AIEE BRANCH

"AIEE District Convention held in Vancouver, B. C., Canada," by Robert Norman and Films, "The Jet Story," and "The Magic of Fluorescence"; October 22, 1953.

Film, "Radio Frequency Heating," and talk by R. C. Siebel; November 12, 1953.

NEW MEXICO COLLEGE OF AGRICULTURE & MECHANIC ARTS, IRE-AIEE BRANCH

General Meeting and Election of New Vice President; December 18, 1953.

COLLEGE OF CITY OF NEW YORK, IRE BRANCH

"Guided Missiles," by Norman Sirota, Fairchild Engine Corporation; November 19, 1953.

NORTH CAROLINA STATE COLLEGE, IRE BRANCH

Film, "Exploration With X-rays," General Electric; December 2, 1953.

NORTHEASTERN UNIVERSITY, IRE-AIEE BRANCH

Business Meeting; December 1, 1953.

"Automatic Pilots," by John J. Hess, Jr., Sperry Gyroscope Company; December 3, 1953.

"Transistors," by Mr. Richard F. Snyder, David Sarnoff Research Center, R.C.A.; December 8, 1953.

Continued on page 96A)

Lab Samples from STACKPOLE

. . . assure accurate engineering plus efficient production

Laboratory samples of Stackpole components are exactly what samples should be—true and accurate reflections of the product when made in quantities and bought for use on your production line.

Stackpole was one of the earliest pioneers in strict statistical quality control. This, plus many unique manufacturing techniques and facilities evolved over years of carbon, graphite and metal powder product specialization, means that the sample you get from Stackpole matches your specifications to the fullest possible extent.

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STACKPOLE CARBON COMPANY

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Keeping Communications on the G-O!

B&W COMPLETE SYSTEMS FOR

- **Frequency-Shift Radio Teletype and Telephone Mobile or Point-to-Point**
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For dependable radio communications . . . linking stations that are fixed or mobile . . . with conventional radio gear or high speed teletype and telephone . . . B&W offers a line of complete systems and special equipment that has been proved and approved by both industry and military.

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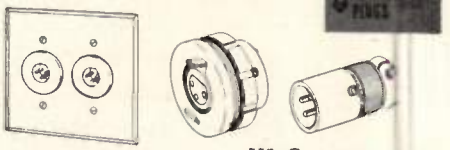
The **XL** *series*

CANNON PLUGS

Cannon's famous XL Connector Series for audio applications is available in 14 shell designs through Radio Parts Distributors and Electrical Wholesalers. These lightweight, compact connectors embody many desirable features: adaptable either to cord end or wall mounting applications...convenient latchlock coupling device...cable entry with compression gland and relief spring or integral clamp...tapped metal for insert retaining screw...provision for special grounding contact or grounding to shell. Accessories include dust caps and adapter shunts, receptacles for integral mounting on microphones and other audio components. Watch for the colorful Cannon counter display carton at your Radio Parts Distributor.

NOTE TO JOBBERS AND WHOLESALERS:

Be sure the name "Cannon" is on every plug you buy...it's your customers' guarantee of satisfaction. For complete details send for Bulletin RJC-6.



XL Connectors are available in 14 shell designs shown here.

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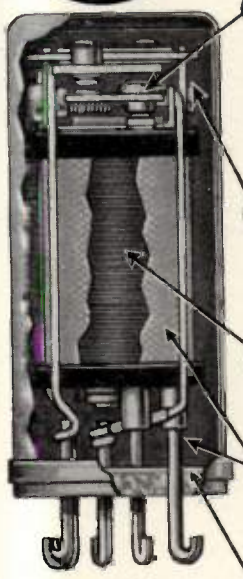
Miniature Tube Size **RELAY** with High Sensitivity

SIZE: ¾" x 1 21/32"
WEIGHT: 1⅛ oz.



1. Armature—precision balanced on needle-point bearings for friction-free movement; maximum resistance to shock and vibration.
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Versatile Mounting

● Complete information and specifications on this "PW" relay are available. Also master catalog on all Potter & Brumfield relays. Write

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From Kitty Hawk Through the Sound Barrier

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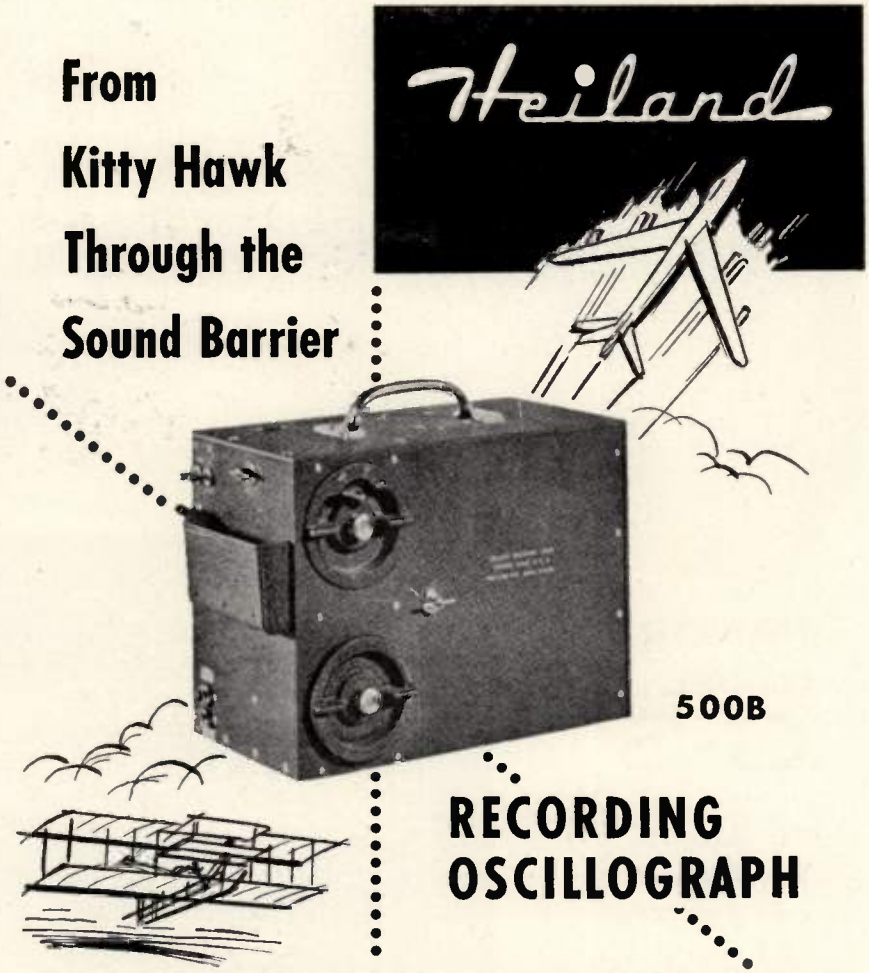

500B

RECORDING OSCILLOGRAPH

Since the Wright Brothers ushered in the age of flight, aviation has soared beyond the speed of sound and into the outer reaches of the earth's atmosphere. Heiland Recording Oscillographs are a vital factor in this dramatic conquest of the air. These highly sensitive, extremely accurate and dependable instruments enable testing engineers and scientists to solve the wide variety of aeronautical problems involving the measurement of physical phenomena such as strains, stresses, vibrations, pressures, temperatures, accelerations, impact, etc. This technical information, provided by Heiland Recording Oscillographs, makes possible the design and production of superior aircraft. Other industries the world over have proven these instruments to be invaluable in meeting similar demands in their applications.

FEATURES AND SPECIFICATIONS

Heiland Model 500B Portable Recording Oscillograph

- **Compact and lightweight**
- **Provides maximum portability and versatility where desired to record up to 12 phenomena**
- **Direct monitoring while recording**
- **4 record speeds instantly selectable**

Galvanometers: Up to 12. Record width: 4". Magazine capacity: 100'. Record speeds: 3, 6, 12 and 24 inches per second selected by levers on control panel. Other speeds, ranging from 1/32" to 24" per second, can be obtained by changing drive sprockets and motor gear box ratios. Drive: Two-speed governor controlled motor. Power supply: 23.5 to 28.25 volts D.C. Timer: Chronometric timing unit projects 100 time lines per second across record width, the 1/10 second lines being heavier. 1/10 and 1 second time lines are available. Mounts: Detachable shock or vibration mounted base. Case dimensions: 6-3/4" x 9-13/16" x 12-3/4". Weight: 33 lbs.

Write today for free catalog of Heiland equipment including 500B and A-401 Portable Recording Oscillographs, "700" series Multi-Channel Recording Oscillographs, "82-6" Bridge Balance and Strain Indicator, Amplifier System, Galvanometers and accessories.

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PANORAMIC MODEL G-3

For Research Development and Production Test Applications

DIRECT READING ULTRASONIC RESPONSE INDICATOR

Frequency Coverage:
2KC to 300KC

Center Frequency:
Variable and calibrated between 2KC and 200KC

Sweepwidth:
Variable and calibrated from 200KC down to 0KC.

Selected sweepwidth remains constant as center frequency is varied.

Amplitude Scale:
Linear or two decade log. Amplitude range 50 db overall.

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2.5 volts, flat to ± 1.5 db

Internal Source Impedance:
600 ohms

Output Attenuation:
Toggle switch operated, 75 db with steps of 40 db, 20 db, 10 db and 5 db.

Used as an adjunct to the Model SB-7 Panoramic Ultrasonic Analyzer the G-3 now makes possible high speed yet accurate visual inspection of amplitude versus frequency characteristics of networks and devices between 2KC and 300KC.
The combination, a complete package, serves as a frequency sweep source and synchronous selective detector which insures indications of *fundamental responses only*. Direct readings of frequency, amplitude, sweepwidth and center frequency are available from the calibrated dials and screen of the SB-7 Panoramic Ultrasonic Analyzer.

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STUDENT BRANCH MEETINGS

(Continued from page 92A)

OKLAHOMA AGRICULTURAL & MECHANICAL COLLEGE, IRE-AIEE BRANCH

"This is your Institute," by J. T. Woodson, W. R. Holway & Associates (AIEE); December 7, 1953.

OREGON STATE COLLEGE, IRE BRANCH

"Measurement Problems Involved in the Development of Precision Potentiometers," by Harry J. Hovee, Brown Electro-Measurement Corp. and "L-C Meter," by Clifford H. Moulton, Tektronix, Inc.; November 23, 1953.

PENNSYLVANIA STATE COLLEGE, IRE-AIEE BRANCH

"Power Installations," with color slides by I. M. McNair, Student Penn. State College; November 11, 1953.

"Engineering Education in England," by Dr. O. A. Saunders, Professor, University of London; November 23, 1953.

RENSSELAER POLYTECHNIC INSTITUTE IRE-AIEE BRANCH

"U.H.F. Television," by A. H. Chisman, Chief Engineer, WTRY & WTRI-TV; November 17, 1953.

"Characteristics of Lightning," by Mr. J. Anderson of General Electric; December 1, 1953.

RUTGERS UNIVERSITY, IRE-AIEE BRANCH

Talk by Mr. Hummel, DuMont Cathode-Ray Tube Plant; November 13, 1953.

General Meeting; December 3, 1953.

Field Trip to Philco Television Assembly Plant; December 4, 1953.

SAN DIEGO STATE COLLEGE, IRE BRANCH

"The Charactron," by D. Jackson, Convair; November 10, 1953.

"Ferroelectric Storage Devices—slides and tape recording, Trips to U.C.L.A. and to Cinerama in Los Angeles; November 24, 1953.

"Analog Computers," by Mr. Barros, Convair; December 8, 1953.

SEATTLE UNIVERSITY, IRE BRANCH

Film "A is for the Atom," and Field Trip Discussion; October 15, 1953.

STANFORD UNIVERSITY, IRE-AIEE BRANCH

"Demonstration of Sterophonic Sound" by Mr. Jack Hauser, Ampex Electric Corp.; December 4, 1953.

SYRACUSE UNIVERSITY, IRE-AIEE BRANCH

"Elements of Report Writing," by John J. Raffone, General Electric; December 3, 1953.

UNIVERSITY OF TEXAS, IRE-AIEE BRANCH

"Electric Engineering Research Lab.," by Dr. A. W. Straiton, Director region 6, October 5, 1953.

"Correlation Computer," by Dr. F. E. Brooks, Faculty University of Texas; October 19, 1953.

UNIVERSITY OF TORONTO, IRE-AIEE BRANCH

"Supervisory Control," by Ken Stock, Nicnacls Ltd.; November 20, 1953.

"Transistors," by J. R. Whyte, Canadian General Electric Co.; December 4, 1953.

Films "Air Waves" and "Radio Broadcasting"; December 11, 1953.

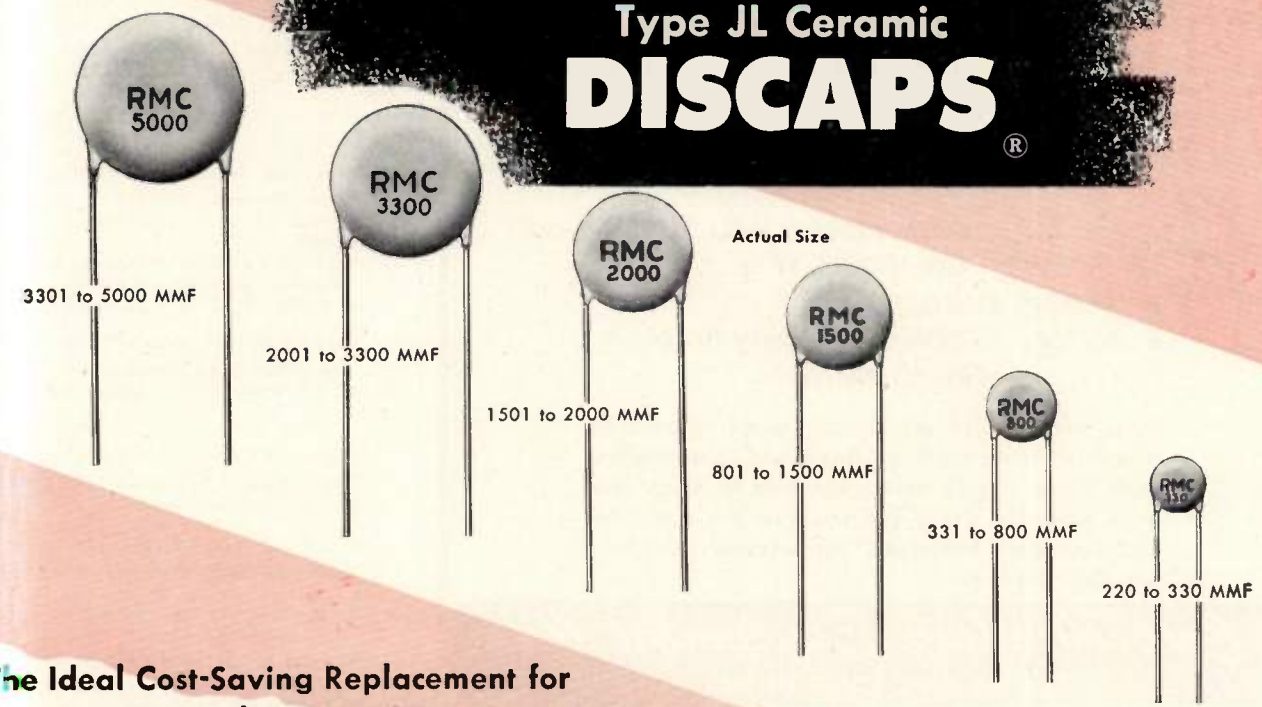
TUFTS COLLEGE, IRE-AIEE BRANCH

"Job Assignment for Junior Engineers," by Dr. Donald B. Sinclair, General Radio Co.; December 2, 1953.

(Continued on page 98A)

Another RMC First

Type JL Ceramic DISCAPS®



The Ideal Cost-Saving Replacement for Paper or General Purpose Mica Capacitors

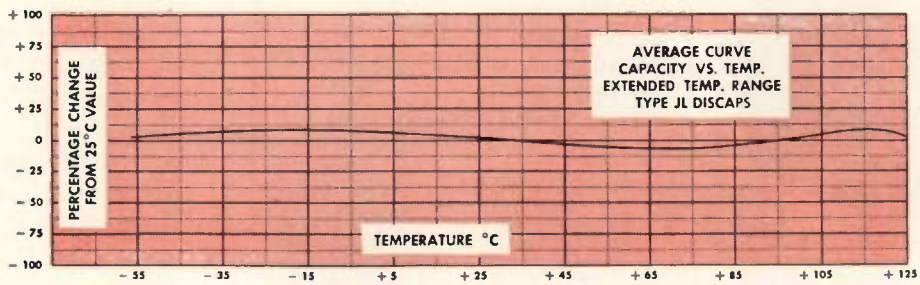
Type JL DISCAPS, the result of extensive research in the RMC Technical Ceramic Laboratories, afford exceptional stability throughout an extended temperature range. The maximum capacity change between −60° C and +125° C is only ±7.5% of capacity at 25° C. Type JL DISCAPS are available in tolerances of ±10% or ±20%. Standard working voltage is 1000 V.D.C.

Manufactured in a wide range of capacities, Type JL DISCAPS offer the advantages of longer life, dependability, and lower initial cost. Their smaller size and greater mechanical strength provide additional economies in assembly line operations.

It will pay you to investigate the advantages of using Type JL DISCAPS as replacements for paper or general purpose mica capacitors. Your inquiry is invited.

SPECIFICATIONS

POWER FACTOR: 1% max. @ 1 K C (initial)
POWER FACTOR: 2.5% max. @ 1 K C, after humidity
WORKING VOLTAGE: 1000 V.D.C.
TEST VOLTAGE (FLASH): 2000 V.D.C.
LEADS: No. 22 tinned copper (.026 dia.)
INSULATION: Durez phenolic—vacuum waxed
INITIAL LEAKAGE RESISTANCE: Guaranteed higher than 7500 megohms
AFTER HUMIDITY LEAKAGE RESISTANCE: Guaranteed higher than 1000 megohms
CAPACITY TOLERANCE: ± 10% ± 20% at 25° C



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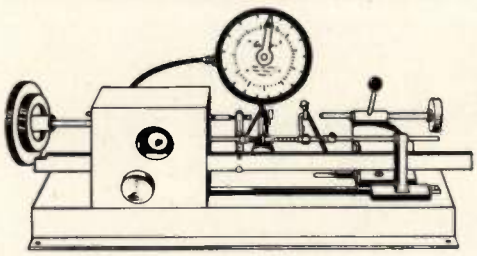
Most Versatile WINDING MACHINE *for*
LABORATORY *and* PRODUCTION USE!



The
MODEL W

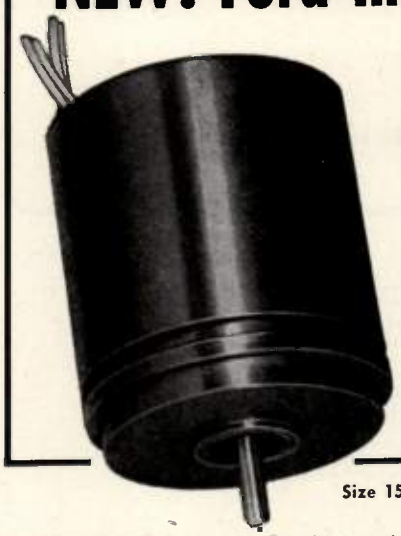
Compare.. COMPARISON WILL SHOW THAT THE MODEL W is:

- *more* RUGGED
- *more* CONVENIENT—easy to set up
  PRICED RIGHT!

Most versatile of all models is our Model W which is widely used in schools and laboratories. With a few simple accessories this machine will wind virtually every coil used in the electrical and electronics industries. Write for detailed listing and description.

Coil Winding Equipment Co.
OYSTER BAY, LONG ISLAND

NEW! Ford Instrument Size 15



TELESYN® RESOLVER

- Extremely accurate computing unit
- Resolves input voltages into sine and cosine components
- Miniature size
- Lightweight
- Precision built to meet rigid military requirements
- Rated 1-24 volts, 400 cycle input

Size 15 Resolver shown full size

WRITE FOR FULL DETAILS — this data sheet gives you the complete facts. Address Dept. IRE.

The Ford Instrument Size 15 Telesyn Resolver is available in three models with transformation ratios of 1:1, 4:1, and 8:1 . . . making it adaptable to numerous applications: analog computers, angle data transmission systems and similar equipment.

13

FORD INSTRUMENT COMPANY
DIVISION OF THE SPERRY CORPORATION
31-10 Thomson Avenue, Long Island City 1, N.Y.

Student Branch Meetings

(Continued from page 96A)

TULANE UNIVERSITY, IRE-AIEE BRANCH

"Electrical Well Logging," by L. E. Chaisson, IRE Student and General Meeting; November 17, 1953.

UTAH STATE AGRICULTURAL COLLEGE, IRE BRANCH

"Development of Fluorescent Light," by Dr. Phillip Hart, Faculty U.S.A.C.; November 19, 1953.

"Electron Microscope," by Dr. George W. Cochran, Faculty, U.S.A.C.; December 9, 1953.

UNIVERSITY OF VERMONT, IRE-AIEE BRANCH

"Power Distribution Systems," by N. Mercoe and W. Prall, both from General Electric; November 12, 1953.

"Radar, It's History, Operation and Uses," by S. F. Anderson, Student Univ. of Vermont; December 9, 1953.

WAYNE UNIVERSITY, IRE-AIEE BRANCH

"The Detroit Edison System and Its Customers," by P. A. Duker, Detroit Edison Co; November 19, 1953.

WEST VIRGINIA UNIVERSITY, IRE-AIEE BRANCH

General Meeting; October 22, 1953.
Business Meeting; November 10, 1953.

UNIVERSITY OF WYOMING, IRE-AIEE BRANCH

Film, "A Voice Shall Be Heard," by T. W. Garrod, Jr., Garrod Radio; November 24, 1953.

YALE UNIVERSITY, IRE-AIEE BRANCH

"Jim Clarke vs. The Right-Hand Rule," by James G. Clarke, Faculty, Yale University; November 24, 1953.

SOLID DELAY LINES

STOP TIME

Suitable for video integration, computers, time markers, moving target indication, etc. LFE Solid Delay Lines offer important advantages in obtaining precise delay intervals for pulse or modulated signals: Wide ranges of delay, low attenuation, low spurious response, wide band-width, smooth pass band, wide temperature range, minimum size and weight, and rugged construction.

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LABORATORY for ELECTRONICS, INC.

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Improve Tube Performance with WESGO "AL-300" ALUMINA INSULATORS

- 97% Al_2O_3 content
- Vacuum tight—extraordinary strength
- Low loss factor — High Te value
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- High strength at all temperatures up to 1500°C.
- Can be supplied in most any shape to extremely close tolerances

Our Engineering Department will gladly answer all inquiries relative to your particular problems

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Ceramic Division
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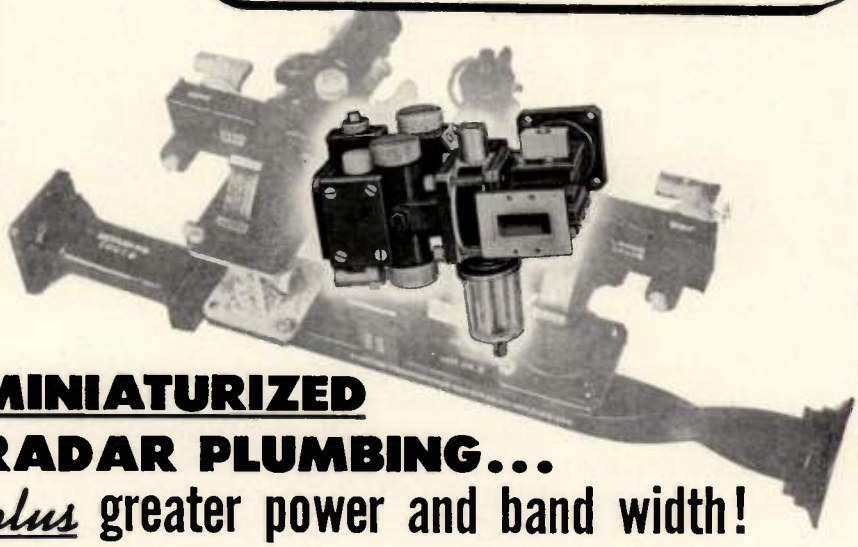
SAN FRANCISCO, CALIF.

No. 1 *in a series*

design progress through inter-company co-operation

. . . coordinated by Airtron inc.



MINIATURIZED RADAR PLUMBING... *plus* greater power and band width!

Recently, Airtron was confronted with a difficult problem — how to develop a radar plumbing system that was greatly reduced in size and weight, yet capable of handling higher power and wider band width. The solution not only necessitated redesign of the actual plumbing, but also required a completely new design of existing crystals and ATR tubes.

Accordingly, a coordinated effort, set up between the engineering staff of Airtron and those of the crystal and tube manufacturers, resulted in the development of new crystals with reverse polarization mountings . . . and new higher power ATR tubes that were inherently pressurizable and demonstrated low VSWR's at high level operation.

As a result, by taking optimum advantage of these new components, Airtron now offers radar designers and manufacturers radically improved duplexers and balanced mixers that are only one-third the previous size . . . yet capable of handling eight times more power . . . with a band width three times greater.

This is just another concrete example of how Airtron's creative engineering . . . and their close association with leading manufacturers in all phases of electronics . . . can be of assistance to you . . . *whether the components you need are new in design or so-called "standard" plumbing.*

New reverse polarized crystal and ATR tube currently used in many radar systems . . . and manufactured according to original design suggestions of Airtron engineers.

Free "Microwave Nomograms and Charts", new 20-page handbook of waveguide engineering data. Write Dept. I for your copy today.

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Manufacturers of a complete line of rigid and flexible waveguide components

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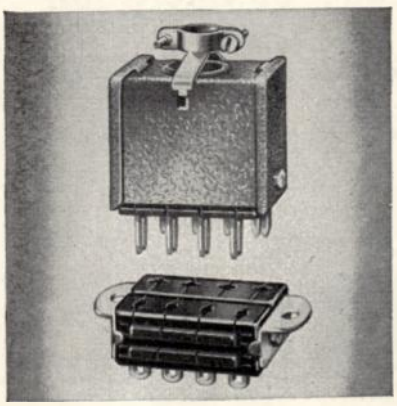

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(Continued on page 104A)

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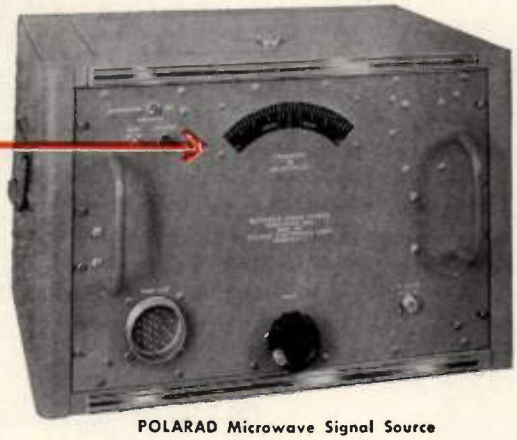

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2 NEW Twist-ear Mounting. (See Types XP45 and UPM45.)

3 PLUG-IN BLADE-TYPE TERMINALS for vertical or horizontal mounting of control to printed circuit panel. (See all photos.)

4 Threaded Bushing Mounting. (See Types XGC-45, GC-U45 and *miniaturized* U70.)

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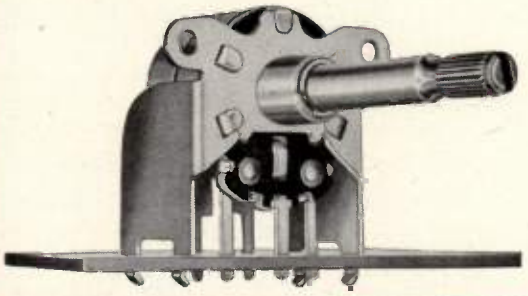
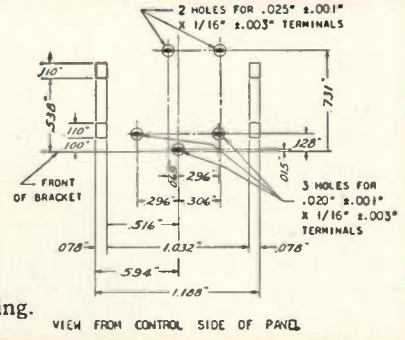
- Terminals available in 7/8" or 1-1/32" lengths from control's center.





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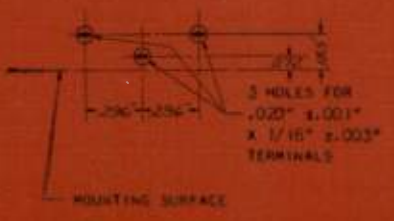
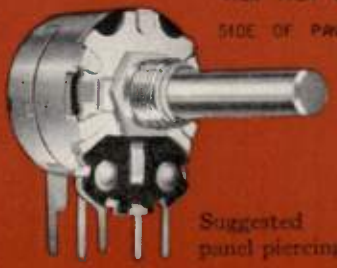
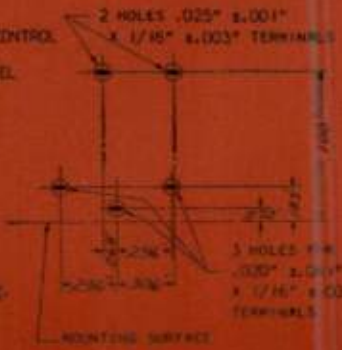
- No strain on printed circuit panel. Anchor tabs attach bracket to cabinet.
- Adequate clearance for circuit paths provided by ample spacing between terminals and by design of mounting lugs on bracket.



Suggested panel piercing.

Type XP45





Suggested panel piercing.

Type XGC-45

For TV panel control applications using a mounting chassis to support printed circuit panel. Twisting 2 ears holds control rigidly to mounting chassis. Available in finger adjusted shaft lengths of 3/4", 7/8", 1", 1-1/4", 1-1/2" and 1-3/4" panel-to-end mounting surface. Also available with recessed screwdriver slotted shaft (Type XP45).

For applications using a mounting chassis to support printed circuit panel. Threaded bushing mounting.

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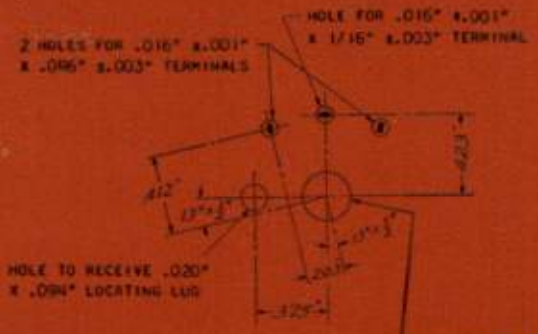
HORIZONTALLY MOUNTED

to Printed Circuit Panel. Shaft extends through panel. (Types U70, GC-U45 and UPM45.)

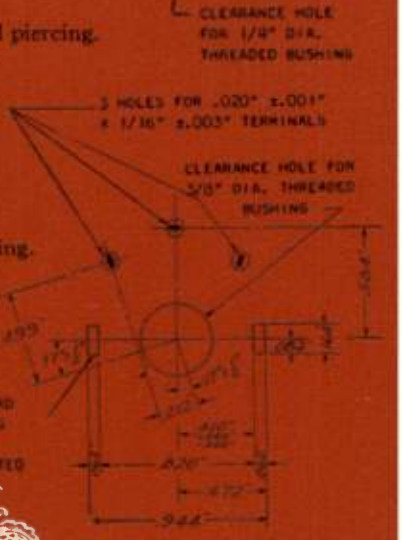
Type U70

(Miniaturized)

Threaded bushing mounting. Terminals extend perpendicularly 5/32" from control's mounting surface.



Suggested panel piercing.



Type UPM45

For TV preset control applications. Recessed screw-driver slotted shaft remains solder-free during panel dipping. Control may be held rigidly to panel before soldering by twisting 2 ears. If ears are left straight, the solder will permanently anchor control to circuit panel. Terminals extend perpendicularly 7/32" from control's mounting surface.

Suggested panel piercing.

Type GC-U45

Threaded bushing mounting. Terminals extend perpendicularly 7/32" from control's mounting surface. Available with or without associated switches.

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(Continued on page 106A)

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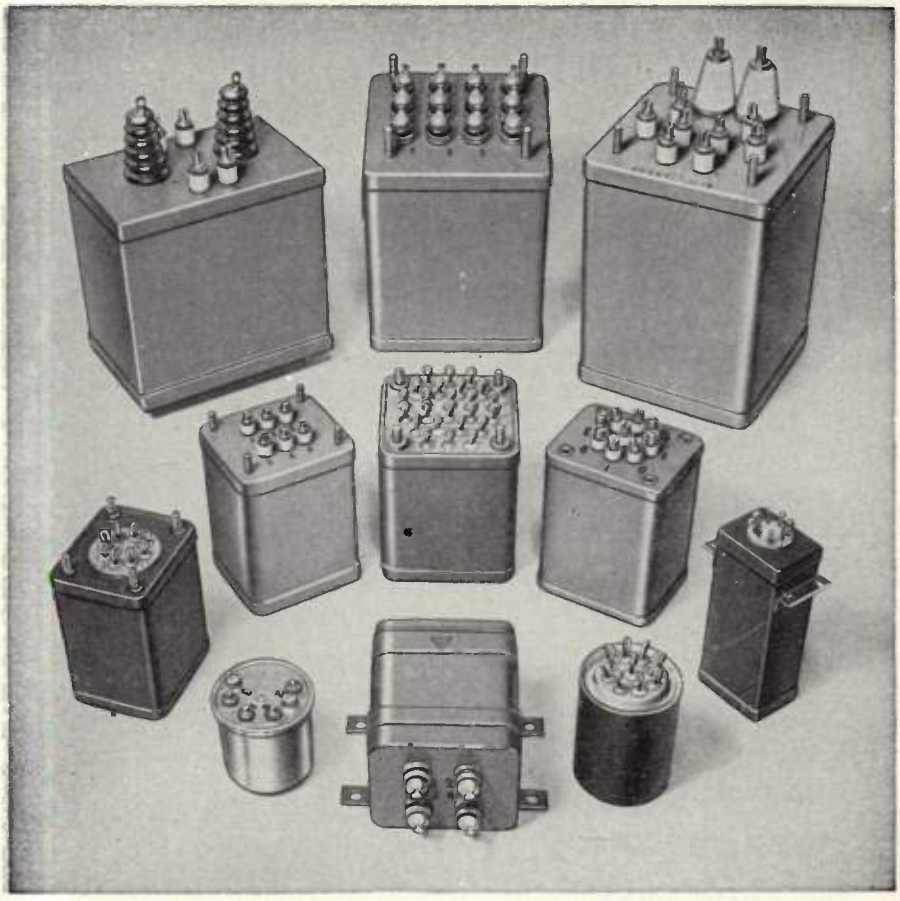

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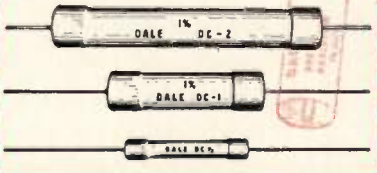

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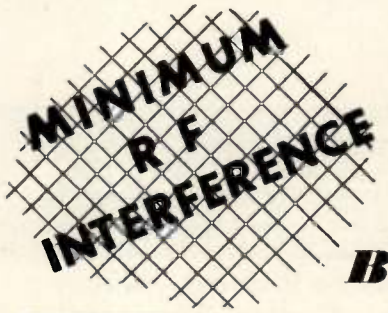

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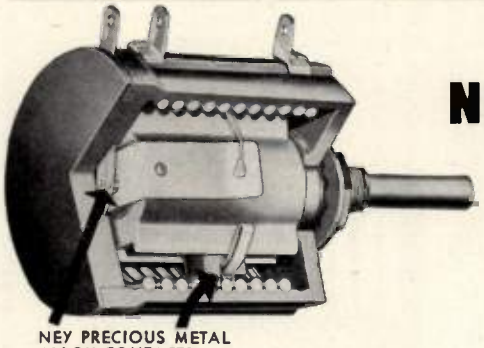

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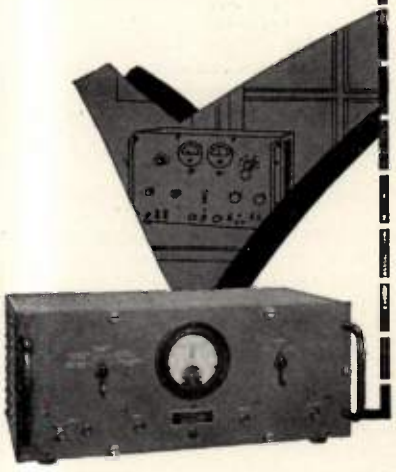
Ledbetter, R. P., 1820½ N.W. 12, Oklahoma City 6, Okla.
Leiner, A. L., 3439—17 St., N.W., Washington 10, D. C.
Leiss, W. J., 1173 S. Atherton St., State College, Pa.
McNabb, R. E., R.F.D. 3, Stamford, Tex.
Michels, L. S., 8835 Reading Ave., Los Angeles 45, Calif.
Miller, J. E., 5458 Mitchell Dr., Dayton 3, Ohio
Mitchell, D., 5816—63 Ave., Riverdale, Md.

(Continued on page 108A)

H-16 CHECKS the CHECKER



ARC Type H-16 STANDARD COURSE-CHECKER

For Omni Signal Generators

■ This newly developed instrument is a means for checking precisely the phase-accuracy of the modulation on VOR (Omnirange) Signal Generators. Now that the use of omnirange receivers and signal generators is so widespread, it is necessary to have a means of measuring the phase differences between the 30 cps envelope of the 9960 ± 480 cps reference modulation, and of the 30 cps variable modulation when that difference is required to be 0, 15, 180 or 195 degrees.

■ An important feature of the H-16 is a built-in self-checking circuit to insure .1 degree accuracy. Errors may be read directly on a 3-inch meter, calibrated to read ± 4 degrees.

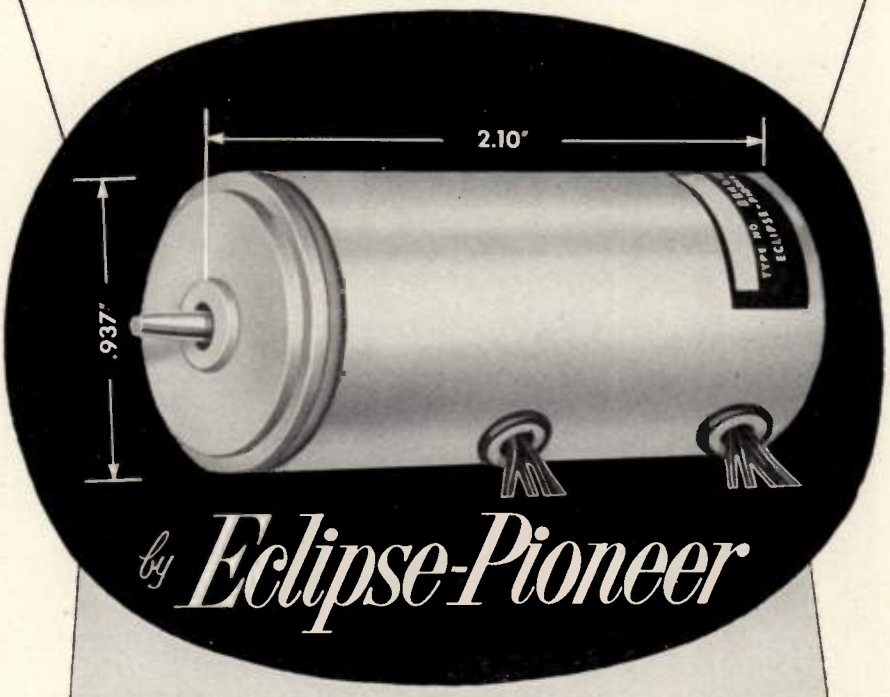
Write for detailed specifications

Dependable Airborne Electronic Equipment Since 1928

Aircraft Radio Corporation
BOONTON NEW JERSEY

LOOK!

CORROSION-RESISTANT PYGMY MOTOR GENERATORS



by *Eclipse-Pioneer*

TYPICAL ELECTRICAL CHARACTERISTICS

GENERATOR DATA

Linearity—½ of 1% to 4000 rpm
Output—300 mv/1000 rpm with 18 volt, 400 cycle excitation
Null—10 mv or less

MOTOR DATA

18 or 26 volt, 400 cycle, 2-phase low inertia motor

As one of the world's oldest and largest producers of synchro-type equipment, we are ideally qualified to provide you with the right answer to your motor generator needs. Why not take advantage of our long, practical experience? Call on us for recommendations based on handling your individual problem most efficiently.

ECLIPSE-PIONEER DIVISION OF **Bendix** AVIATION CORPORATION

For full details, write Department G.
TETERBORO, N. J.
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You're looking over the Record Player Slide most* manufacturers look up to

This is the Grant "A" Radio-Record Player Slide. You can see the features that make it so popular... continuous ball bearing action for finger-touch sliding, the silencer spring that eliminates any vibration or possible resonance, solid, dependable construction. The patented stop pin permits the phonograph shelf to be removed easily after installation.

This and other models in Grant's *complete* line operate silently with a minimum of effort, but have negligible sway or sag, even in fully extended position. And installation is fast and economical.

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Motorola
Olympic
Stromberg-Carlson
RCA
Midwest
Stewart-Warner
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Fischer
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Admiral
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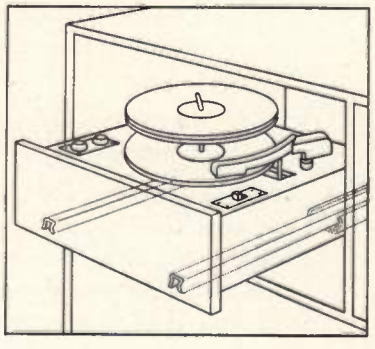

For full information, write to

Grant pulley & hardware co.

31-87 Whitestone Parkway, Flushing, N. Y.

MEMBERSHIP

(Continued from page 106A)

Morton, W. E., Jr., 5925 N. Warnock St., Philadelphia 41, Pa.
Mulligan, J. M., 819 Clairmont Ave., Elmira, N. Y.
Otis, D. C., 3215 Robinhood St., Houston 5, Tex.
Pieck, H., 81 Melville Ave., Toronto, Ont., Canada
Reynolds, R. S., 4708 Pershing Ave., S.E., Albuquerque, N. Mex.
Rice, N. O., General Electric Co., Electronics Park, Rm. 305, Bldg. 5, Syracuse, N. Y.
Riggen, T. K., 922 Bridgman St., Elmira, N. Y.
Rohr, W. E., Development Detachment, 9403 TSU, Signal Corps Engineering Laboratories, Fort Monmouth, N. J.
Sauer, R. J., 148 E. Beechwood Ave., Dayton 5, Ohio
Schildhauer, W. E., The Carborundum Co. Globar Division, Niagara Falls, N. Y.
Seide, J. H., 1020 Sherman, Holdrege, Nebr.
Sergeant, R. C., Electronic Supply Office, Great Lakes, Ill.
Singer, L. M., Dept. 70, Bendix Radio, Baltimore 4, Md.
Stocker, G. A., c/o International Aeradic (Pakistan), Ltd., David Sassoon Bldg., McLeod Rd., Karachi, Pakistan
Vogelsong, J. H., 283 Kings Rd., Madison, N. J.
Walker, J. B., 67 Broad St., Rm. 1004, New York 4, N. Y.
Watson, D. J., 465 Crescent, St. Lambert, Que., Canada
Watson, H. A., Bell Telephone Laboratories, Murray Hill, N. J.
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(Continued on page 110A)

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Read by Sight, Sound or Feel.

- Permanently Accurate
- Practically Indestructible
- Faster—Easier to use
- Automatic Release
- All Capacities

in inch ounces . . . inch pounds . . . foot pounds (All Sizes from 0-6000 ft. lbs.)

Every manufacturer, designer and production man should have this valuable data. Sent upon request.

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ADDISON (QUALITY) ILLINOIS

PRECISION *in* PINT-SIZE!

Bliley

BANTAM BX

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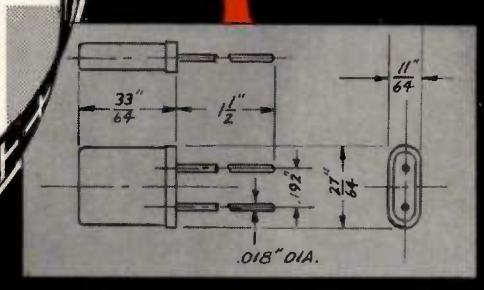

ACTUAL SIZE

BLILEY BANTAM BX
FREQ: 15mc-100mc

The new **BANTAM BX** has wire leads for mounting in a miniature selector switch or can be soldered to a printed circuit terminal board.

The new **BANTAM BX** does not sacrifice precision or dependability. It meets same performance and tolerance requirements as larger crystal units such as MIL types CR-23 or CR-32.

The new **BANTAM BX** is fully described in Bulletin 46. Now available for prototypes or in production quantities.

Bliley CRYSTALS

BLILEY ELECTRIC COMPANY
UNION STATION BLDG., ERIE, PA.

PRECISION • STABILITY • LOW NOISE LEVEL
S.S.WHITE TYPE 65X MOLDED RESISTORS

Rating—1 Watt.
Temperature Coefficient—From approx. + 0.1%/°F for 5000 ohm resistors to approx. − 0.2%/°F for 10 megohm resistors.
Voltage Coefficient—Less than 0.02%/Volt.
Upper Temperature Limit—170° F for continuous operation.
Standard Range—1000 ohms to 9 megohms. For general commercial service.
Noise Level—A low noise level is inherent in all 65X resistors. Where noiseless operation is essential, we are equipped to test and guarantee standard range resistors with "less noise than corresponds to a resistance change of 1 part in 1,000,000, for the complete audio frequency range."

EXTREME HIGH VALUE RANGE
Up to 10,000,000 megohms for specialized scientific and industrial service.

Send for Bulletin 4906. It gives full details on S.S.White Molded Resistors. Write for a copy.

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105-125 V. 60C INPUT
OUTPUT CURRENT MAY BE SET AT ANY VALUE FROM 0.2 TO 50 M.A. D.C.
AS LOAD VOLTAGE VARIES BETWEEN 0-150 V. MODEL 1A-R IS CONSTANT WITHIN 1%, MODEL 1B-R WITHIN 0.1%.
IMMEDIATE DELIVERY
MODEL 1 A-R $149—MODEL 1 B-R $286
0.2-100 M.A. RANGE $13 MORE
Other models to specifications

ELECTRONICALLY REGULATED
D.C. CONSTANT CURRENT POWER SUPPLIES
OUTPUT CURRENT IS HELD CONSTANT AS LOAD VARIES

OTHER MODELS AVAILABLE
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(Continued from page 108A)

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Bailey, G. J., 53 Vicarage La., Ilford, Essex, England
Barchenger, B. W., Box 8794, WADC, Wright-Patterson AFB, Ohio
Barnes, G. H., 455 W. Chestnut Hill Ave., Philadelphia 18, Pa.
Bartlett, R. L., 231 Green, Norfolk 13, Va.
Belford, K. R., Box 34, R.F.D. 1, Cloverdale, Calif.
Bell, B. E., 210 De Vargas, Apt. A, El Paso, Tex.
Bemis, R. C. B., 410 N. Euclid Ave., Apt. 2A, Pasadena 4, Calif.
Bettis, M. E., 406 W. 34 St., Kansas City 11, Mo.
Blake, R. E., 239 Halsted St., East Orange, N. J.
Bocchio, T. R. R., 7342 Pearson St., Lemon Grove, Calif.
Bowles, R. D., 5229 Churchward St., San Diego 14, Calif.
Boyd, R. S., 182 Lakeshore Dr., New Toronto 14, Ont., Canada
Brinkerhoff, D. E., Box 315B, State Rd. 26E, Kokomo, Ind.
Brown, R. E., 1532 S. Delaware Ave., Tulsa, Okla.
Broyles, R. S., 1980 N. Orange Grove Ave., Pomona, Calif.
Bunce, D. S., 824 Sunnyside, Chicago 40, Ill.
Bush, F. J., 51 Robin La., Levittown, L. I., N. Y.
Butsch, L. M., Jr., 23 Malcom Dr., Dayton 10, Ohio
Caffrey, T. P., 170 Hastings Pl., Syracuse 6, N. Y.

(Continued on page 112A)

PRD
TYPE 904
VHF-UHF NOISE GENERATOR

FREQUENCY RANGE (mc/sec): 10 to 1000
NOISE FACTOR RANGE (db): 20
CHARACTERISTIC IMPEDANCE 50 ohms (unbalanced)

This calibrated broadband noise source permits direct measurements of noise factors as high as 20 db for r-f amplifiers and receivers operating in the range from 10 to 1000 mc/s. Equipment is housed in an attractive metal cabinet.
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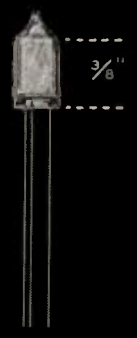
AUTOMATIC
TRANSISTOR
MACHINERY

you can produce
from 30 per hour
to 1200 per hour
with
KAHLE equipment



T2 X 3 ½" H (PHOTO ACTUAL SIZE)

Now you can manufacture transistors that are evacuated and sealed in glass at rates from 30 to 1200 per hour or more! KAHLE, the largest producer of custom machines for the glass and electronics industries, supplies equipment for every operation in making a transistor or crystal diode,— manual, semi-automatic, fully automatic.

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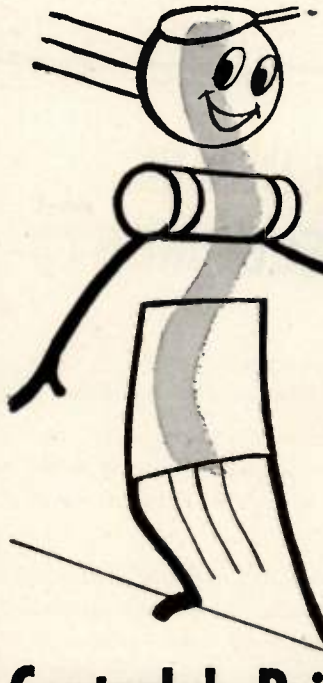
CONTI



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- Reduce stray capacities where not wanted.
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- Saves up to 80% of soldered connections.
- Combine capacitance, resistance and wiring on one plate.
- 100 or 1,000,000 plates all identical — all tested.
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"Now! Dictate a letter for a complete package of Centralab's technical sheets on PECs."

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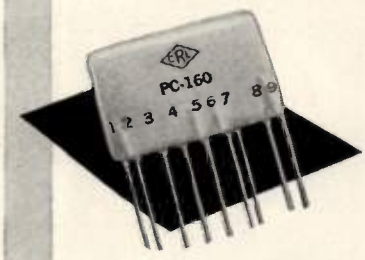

NEW PENDET

— a complete pentode detector and audio coupler circuit that replaces 9 parts . . . 9 soldered connections.

There are actually 4 resistors and 5 capacitors screened and fired to a single Ceramic-X plate. Unit is only $1\frac{5}{16}$ x $\frac{7}{8}$ x $\frac{11}{64}$" thick. Leads are $2\frac{1}{2}$" long.

Trademark —
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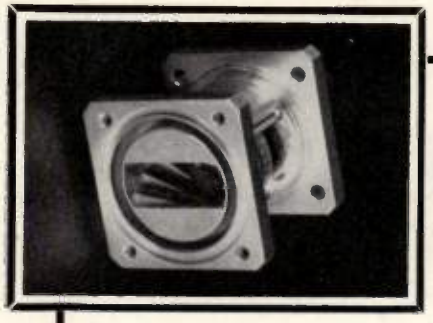
Industry's greatest source of standard and special electronic components

Y-1



MICROWAVE DEVELOPMENTS

Wheeler Laboratories is an engineering organization which offers consulting and engineering services in the fields of radio and radar.

The pictured waveguide step-twist component is a typical development by the Laboratories. This component, originally designed to fill the specific need of a client, has since been adapted for universal waveguide use.

At present, Wheeler Laboratories includes a staff of twenty engineers under the personal direction of Harold A. Wheeler, a group of designers, and a model shop; regular additions to the staff are continuing in order to keep pace with our expanding program.

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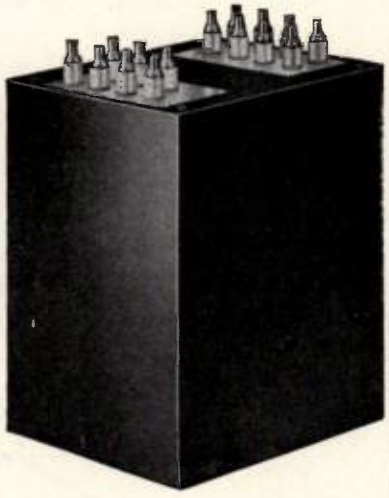
Quality COMES FIRST



● If performance and long life are the primary factors in your application of transformers, then submit your specifications to Acme Electric. Quality comes first in every Acme Electric transformer.

ACME ELECTRIC CORPORATION
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Acme ACME Electric
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Castner, R. J., 6330 Morrowfield Ave., Pittsburgh 17, Pa.
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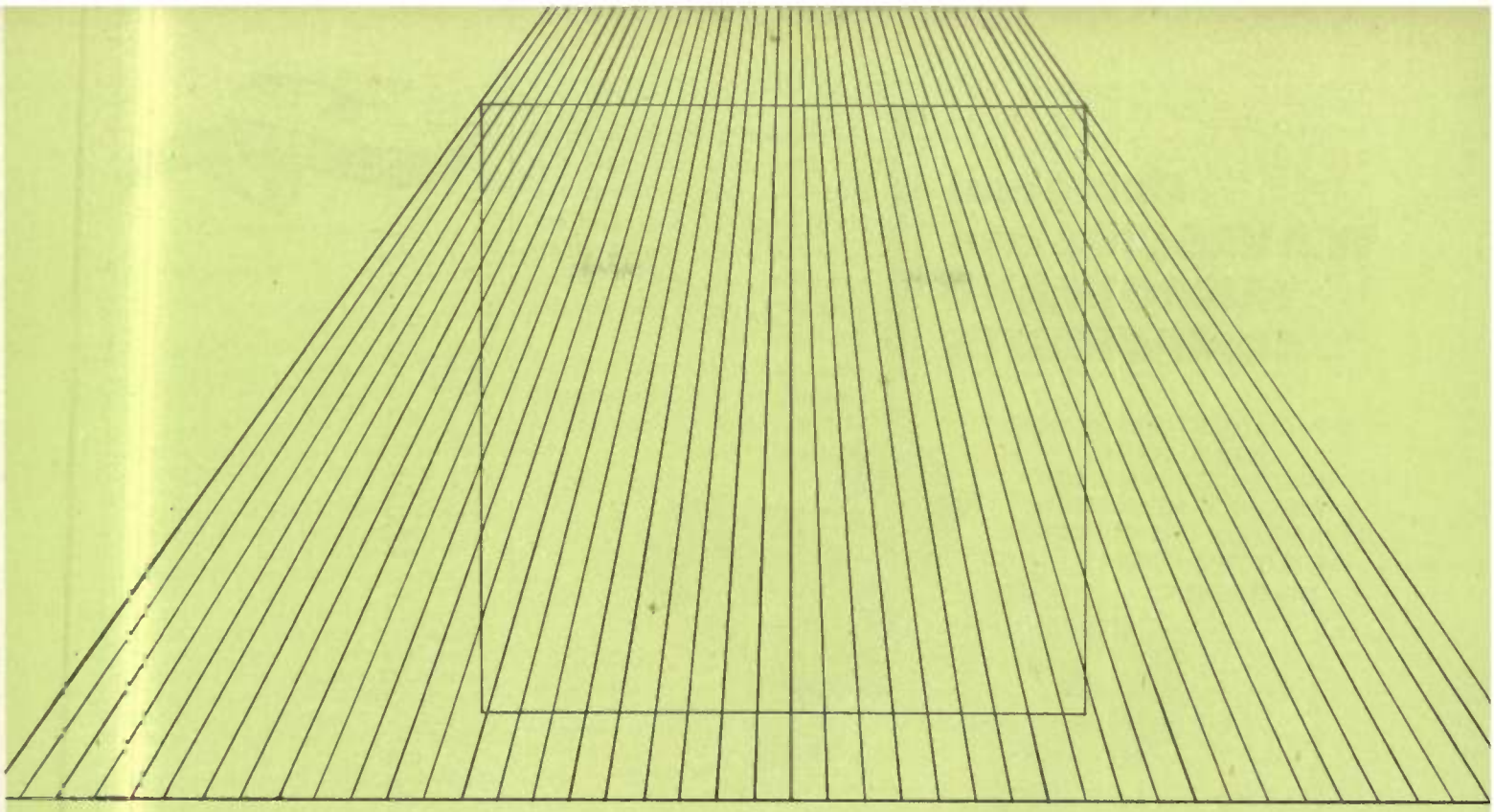
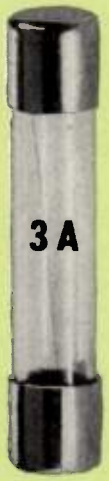
Child, F. T., 6545 S. Union Ave., Chicago 21, Ill.
Chin, H. C., 214 Canal St., New York 13, N. Y.
Chipman, J. R., 744 Loretta St., Tonawanda, N. Y.
Coburn, A. H., Oakley Ter, Cambridge, Md
Cooper, W. N., 322 Willard Ave., Phoebus, Va.
Cotton, M. L., Department of Electronics, U. S. Naval Postgraduate School, Monterey, Calif.
Cox, H. L., Jr., 1548 Michigan Ave., Columbus 1, Ohio
Crovella, E., 1012 N. Military Rd., Niagara Falls, N. Y.
Cullen, A. B., Physics Bldg., University of Mississippi, University, Miss.
Cutter, R. H., 7206 S. Champlain Ave., Chicago 19, Ill.
Davis, E. E., 8235 Briarwood Cir., Norfolk, Va.
Davis, G. W., 2225 Cromwell Rd., Norfolk, Va.
DeAgro, R. P., 5142 Klingle St., N.W., Washington 16, D. C.
Deranian, M., 7235 Githens Ave., Pennsauken 8, N. J.
Dickerson, M. L., 2630 Hazel St., Beaumont, Tex.
Dittrich, W. A., 74—24 St., N.W., Barberton, Ohio
Dulong, P. R., 425 First St., Mineola, L. I., N. Y.
Durakov, J. S., 35th Air Division (Def.), Dobbins AFB, Marietta, Ga.
Ecklund, R. E., 4845 N. Magnolia, Chicago 40, Ill.
El Meshad, Y. A., 4, Dinocrat St., Alexandria, Egypt
Esson, R. C., 9th A&E Maintenance Sq., Box 76, Mt. Home AFB, Ida.
Fitzpatrick, C. A., 41 Woodlawn Dr., Chatham, N. J.
Foley, T. P., 9419 Linden Blvd., Ozone Park 16, L. I., N. Y.
Ford, R. R., 7907 District Heights Pkwy., Washington 28, D. C.
Fucci, E. A., Box 3022, USAFIT, Wright-Patterson AFB, Ohio
Gale, W. L., 1511 W. Bryn Mawr, Chicago 26, Ill.
Gandin, P., 40 bis rue Rivay Levallois, Seine, France
Gerhart, G. J., 364 W. Market St., Akron, Ohio
Ghosh, S. R., 10, Deodar St., Calcutta 19, India
Gilmer, J. R., 2312 Proctor, Garland, Tex.
Gleason, D. H., 102 N. Thomas St., Apt. 1, Arlington 3, Va.
Glenn, L., Jr., Box 2023, USAFIT, Wright-Patterson AFB, Ohio
Goodman, K., 1230 S. Gertruda Ave., Redondo Beach, Calif.
Graettinger, W. F., 1817 Sandra, Midwest City, Okla.
Grant, H. S., 169 Brentwood Rd., N., Toronto 8, Ont., Canada
Greenberg, S., 8-03—149 St., Whitestone 57, L. I., N. Y.
Guyer, E., 951 Lancaster Ave., Syracuse 10, N. Y.
Haefeli, J. E., 371—15 Ave., Paterson, N. J.
Hageman, D. H. A., 3986 Hillcrest Dr., Los Angeles 8, Calif.
Hales, D. B., R.F.D. 1, Payson, Utah
Haring, D. R., 326 Chambers St., Phillipsburg, N. J.
Harrell, R. D., MCLI, Box 3026, Wright-Patterson AFB, Ohio
Hart, L. D., 5831 Walmer, Mission, Kansas
Haughey, J. W., 1948 Everitt St., Valley Stream, L. I., N. Y.
Hertz, S., 47-48—197 St., Flushing, L. I., N. Y.
Hill, B. E., Box 3039, MCLI, USAFIT, Wright-Patterson AFB, Ohio
Hoffman, C. H., 304½ E. Dayton, South Bend, Ind.

(Continued on page 114A)



THINGS ARE NOT AS THEY SEEM...

This *is* a perfect square.
It is an optical illusion that the sides bend.



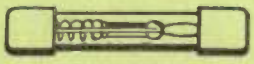
3 amps fuse will not blow at 3 amps.

Fuses are not rated by the current at which they blow. Fuses are rated by the maximum current they should carry indefinitely.

Each type of fuse blows according to the requirements of the equipment it was designed to protect.

Littelfuse has cooperated with NEC, Underwriters, Armed Forces MIL Specs Committees in establishing the characteristics of the various fuse types.

Littelfuse holds more design patents on fuses than all other manufacturers combined.



3 AG "SLO-BLO"

3 AB

8 AG U/L

1 AG

4 AG ANTI-VIBRATION

LITTELFUSE

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"ACCURACY IS A UNIVERSAL WORD!"

Specify **Universal WAVEGUIDE and MICROWAVE COMPONENTS** for

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IMPEDANCE MEASUREMENT
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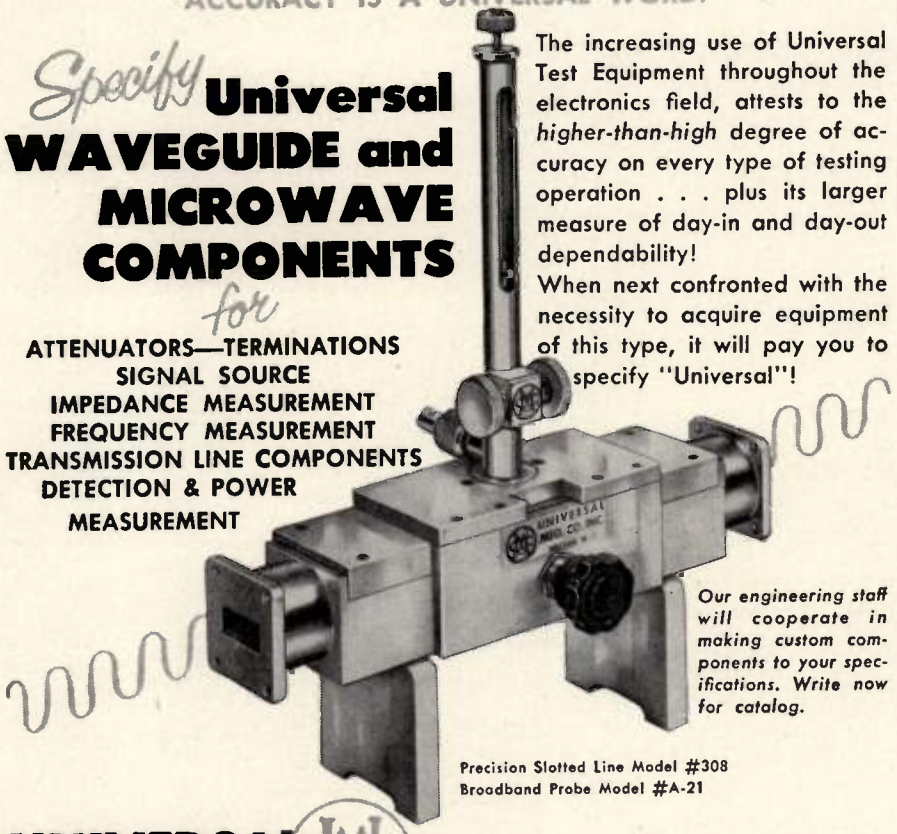
The increasing use of Universal Test Equipment throughout the electronics field, attests to the *higher-than-high* degree of accuracy on every type of testing operation . . . plus its larger measure of day-in and day-out dependability!

When next confronted with the necessity to acquire equipment of this type, it will pay you to specify "Universal"!



Our engineering staff will cooperate in making custom components to your specifications. Write now for catalog.

Precision Slotted Line Model #308
Broadband Probe Model #A-21

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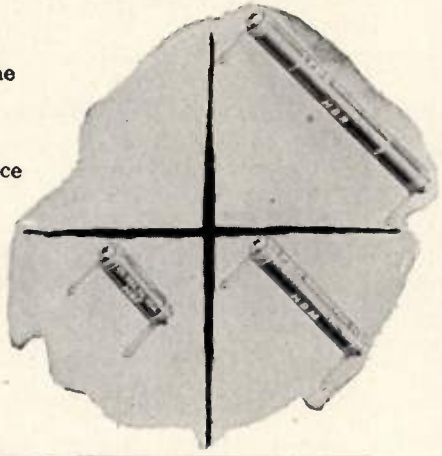
RESISTANCE VALUES up to 100 Million Megohms!

RPC Type H, High Megohm Resistors are particularly useful in applications where resistances of the *highest quality* are required. Eminently suited for electrometer circuits, radiation equipment, photo-cell circuits and high resistance standards in measuring equipment. Advanced design. High stability carbon coating on non-hygroscopic steatite rod. Can be soldered into circuit or mounted on a panel or standoff insulator. Standard resistance tolerance 10%. Closer tolerances available.

Write for additional information.



Type	Length	Resistance		Max. Voltage
		Min.	Max.	
HBF	1"	5 meg.	25 million meg.	3,500
HBM	1¾"	10 meg.	50 million meg.	7,500
HBR	3"	20 meg.	100 million meg.	15,000

rpc

RESISTANCE PRODUCTS CO.
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MAKERS OF RESISTORS—HIGH MEGOHM, HIGH VOLTAGE, HIGH FREQUENCY, PRECISION WIRE WOUND

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Horsch, J. R., 317 Springfield Rd., E. Syracuse, N. Y.
Hunting, R. A., 684 Taunton Pl., Buffalo 16, N. Y.
Ingram, J. H., 3103 Floyd, Dallas, Tex.
Jesswein, A. J., Jr., 2513 Addison Rd., S.E., Washington 28, D. C.
Johnson, L. W., 118 N. Rich St., Shawnee, Okla.
Jolkovski, J., 901 Potomac Ave., Alexandria, Va.
Jones, M. W. 82-27—268 St., Floral Park, L. I., N. Y.
Joseph, E. C., 555—23 St., S., Arlington, Va.
Jurgielewicz, S. S., Box 3040, USAFIT, Wright-Patterson AFB, Ohio
Kachurchak, C., 2839 Hudson Dr., Cuyahoga Falls, Ohio
Karagiorgis, N. D., 4506 Elm St., Bethesda, Md.
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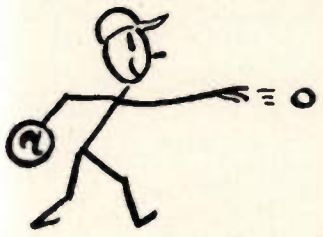
(Continued on page 116A)



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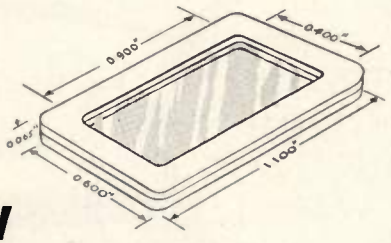
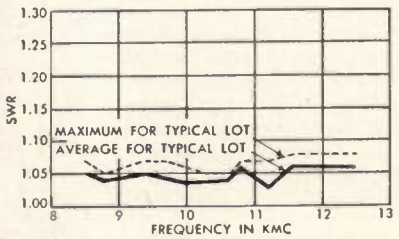


The AIL Type X-52 Nonresonant Pressure Window covers the entire band of the RG-52/U waveguide (8,200 to 12,400 Mc). Its standing-wave ratio does not exceed 1.08 over this band.



- Withstands pressure differential of 17 pounds per square inch.
- Can be used over a temperature range from −75 C. to +100 C.
- Can be soldered into a cutout of an RG-52/U waveguide flange.

$25.00 each

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An instrument consisting of two separate electrical sections, one for A.C. and one for D.C., coupled to a single D'Arsonval movement.

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20 RANGES IN A.C. Voltage (DC or AC up to 15,000 cycles) from 500 millivolts full scale up to 750 volts

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(Continued on page 120A)

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EXTRUDED TEFLON (Tetrafluoroethylene) hook-up wire is organically capable of sustained operation from +210°C to —90°C with no appreciable decomposition. This wide range of operating efficiency continually opens new applications for **EXTRUDED TEFLON** — especially where constant stability under exceptional temperature conditions is required for long periods.

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You'll want the full details on CHICAGO'S complete "Sealed-in-Steel" transformer line. There is a unit available for every modern circuit requirement. Write for your Free copy of Catalog CT-153, or get it from your electronic parts distributor.

H-TYPE. Steel base cover is deep-seal soldered into case. Terminals hermetically sealed. Stud-mounted unit. Meets all MIL-T-27 specs.

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C-TYPE. With 10" color-coded stripped and tinned leads brought out through fibre board base cover. Flange-mounted unit.

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This Brush multi-channel magnetic head (Model BK-1500 series) features precision gap alignment. When you use tape recorders incorporating these heads, you can record data on one machine, and play back on another—*with all signals remaining in perfect time-phase relationship.* As many as 14 tracks can be recorded on a single tape 1¾ inches wide.

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NAME	IMPEDANCE	FREQUENCY	MAX. VSWR	PRICE
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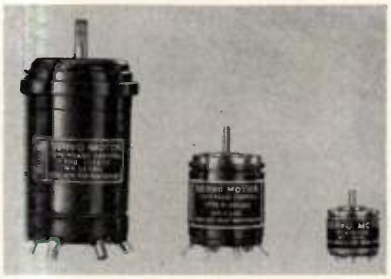
News—New Products

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

(Continued from page 62A)

Servo Motors

Kintay Mfg. Co., 55 Broadway, New York 12, N. Y., has three new servo motors available. The one at the left is a 6 watt model in a size 23 frame. It weighs 1.6 pounds, has a rotor inertia of 20.0 gm-cm², and has a theoretical acceleration of 26,500 RAD/sec². No load speed is 3,500 rpm, stall torque is 7.5 ounce/inches, and maximum output is 6.0 watts.



The center illustration is the 60 cps 115 volt model in a size 15 frame. Weight is 7.3 ounces, rotor inertia 3.03 gm-cm², and has a theoretical acceleration of 33,800 RAD/sec², No load speed is 3,300 rpm, stall torque 1.45 ounce/inches, and maximum output is 1.23 volts.

The motor on the right weighs 1.45 ounces, has a rotor inertia of 0.46 gm-cm², and a theoretical acceleration of 49,000 RAD/sec². No load speed is 6,500 rpm, stall torque 0.3 ounce/inches, and maximum output is 0.490 watts.

Thermally Tuned Reflex Klystron

A thermally tuned reflex klystron tube designed for K-band operation has been announced by **Bendix Aviation Corp., Red Bank Div.**, Eatontown, N. J. This tube, JAN 2K50 will operate over a frequency range of 23,504 mc to 24, 464 mc with a minimum power output of 8.5 mw. It will tune this frequency range in 1.2 to 2.6 seconds.

The JAN 2K50 is tuned by varying the grid bias voltage of a triode section incorporated in a metal envelepe. The plate of this triode section is attached to the klystron structure. Thermal expansion of the plate, caused by variations of plate current, is transmitted to the cavity of the klystron. This results in a change in the gap spacing and consequently the frequency.

Absorption Type UHF WATTMETER and DUMMY LOAD

FREQUENCY RANGE 2-900 MEGACYCLES
PRICE $135.00

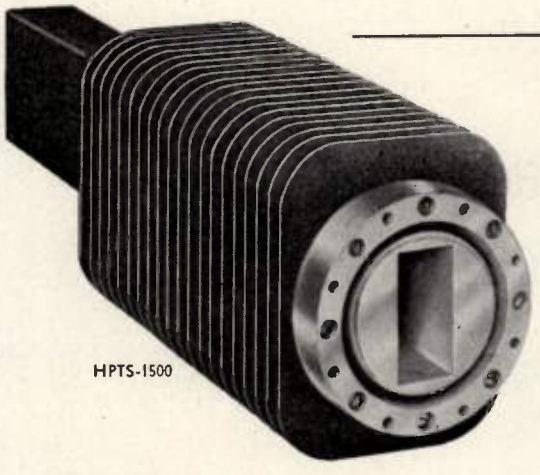
1. Oil filled, air cooled, coaxial load resistor.
2. Crystal voltmeter calibrated to read watts, measures voltage across tap on load resistor.
3. Load resistor dissipates 90 watts continuously.
4. Power ranges, 0 to 15 watts and 0 to 60 watts are standard. Other ranges between 7 to 90 watts can be supplied.
5. Calibration: May be calibrated with an accurately known power source at 60 cycles per second. Measurements of UHF power may be checked against 60 cycles per second power.
6. Impedance—50 OHMS.
   Type N female input connector.
7. Voltage Standing Wave Ratio: Less than 1.15 from D.C. to 900 MC.
8. Frequency response of wattmeter depends on VSWR only.

The chief advantage of this instrument lies in the use of a tapped load resistor for power measurement. The frequency response of the tapped load resistor is flat from DC to 900 MC, and accurate calibration of any point of the meter scale can be effected using an accurately known 60 cycle per second power source.

Hi Power U. H. F. ATTENUATOR 100-Watts

20 db, 30 db or 40 db, 51.5 OHMS impedance, type N connectors, dissipates 100 watts of Power, uses tapped load resistor as above, frequency range 1-700 Megacycles.

PRICE $95.00



HPTS-1500

HIGH POWER WAVEGUIDE DUMMY LOADS

HPTS-1500 is an S Band, waveguide termination capable of 1500 watts dissipation in still-air, and has a V.S.W.R. of less than 1.1 over the range of 2400 to 3700 mc.

PRICE $185.00

HPTXS-175 is an X Band, waveguide termination, capable of 200 watts dissipation in still-air. V.S.W.R. will not exceed 1.1 over the range of 8500 to 12,000 mc.

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ALSO AVAILABLE: Hi Power Waveguide Dummy Loads DA-18, DA-21, DA-22, TS-108, TS-231, TS-235, TS-338.

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FREQUENCY METER 810
8,200 mc to 12,400 mc

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SPECIFICATIONS:

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	Precision:	0.02%
	Loaded Q:	7000
	Reactive dip:	10% min.

Detailed catalog sheet RA-2 on request

NEW IMPROVED BOLOMETER

APPLICATIONS:
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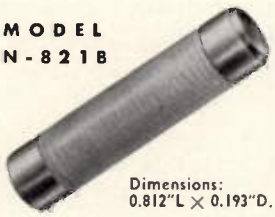
Mechanically stronger · ruggedized · silver contacts · Copolymerstyrene cartridge · completely interchangeable in all Barretter mounts · Resistance: 200 ohms at 8.75 ma · Sensitivity: 4.5 ohms per mw. Only $7.50 ppd.

MODEL N-821B



Dimensions: 0.812"L × 0.193"D.

Data sheet RB-2 on request

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Weber, F. V., 441 S. Kenilworth Ave., Oak Park, Ill.
Webster, A. L., Box 3101, USAFIT, Wright-Patterson AFB, Ohio
Weidman, J. S., Bendix Research Laboratories, 4855 Fourth, Detroit 2, Mich.
Weinhouse, N. P., 6247 Dykes Way, Dallas, Tex.
Welk, S. A., 41 Madison Ave., Toronto 5 Ont., Canada
White, T. J., 939 Riverwood Ave., Fort Garry, Man., Canada
Wiack, R. J., 227 Warren St., Jersey City 2, N. J.
Wigger, R. C., 2857 Malcolm Ave., Los Angeles, Calif.
Wilke, W. E., 5410 Thornburn St., Los Angeles 45, Calif.
Yeager, D. L., 307 Ave. G., Garland, Tex.
Zaida, W. L., 7515 W. Winona, Chicago 31, Ill.
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Zipf, A. R., Bank of America N.T. & S.A., 650 S. Spring St., Los Angeles, Calif.
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(Continued on page 124A)

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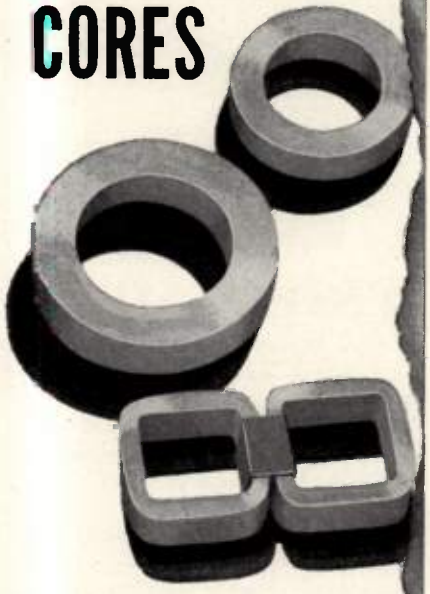

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- Pressures
- Weights
- Liquid Levels
- Excursion
- Expansion
- Acceleration

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- Derivation of components of complex motions in multiple coordinates, in electrical terms
- Generation of electrical functions
- Analog Computation

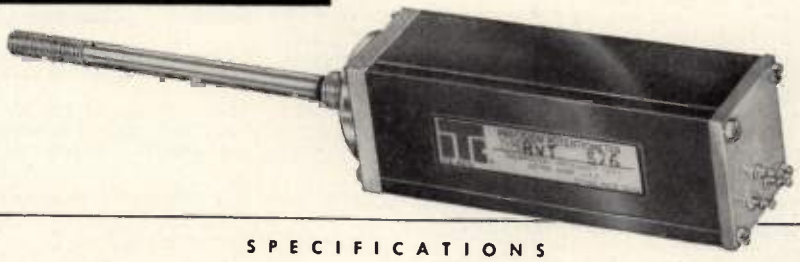
MONITORING AND CONTROL

- Detection and correction of motions exceeding pre-set limits
- Compensation of aberrated motions may be made by choice of potentiometer windings of the non-linear type

SENSING OF ANY MOTION IN HAZARDOUS AREAS

- RVT Potentiometers reliably replace the human watchman or operator in dangerous or injurious environments, and safely provide remote indication, alarm operation, or control surveillance.

The exacting standard of performance that TIC has made famous in precision rotary potentiometers has been applied to the field of straight-line or rectilineal shaft motion potentiometers. A guided centerless ground stainless steel shaft is free to move longitudinally down the axis of a machined aluminum housing. The shaft carries up to three separate dual contacts of precious metal which slide upon as many precision wound resistor strips, giving resistance variation proportional to the shaft motion. Units having multiple strips and contacts preserve the electrical individuality of each so that many wiring combinations are possible for control of sensitivity, resolution or electric current function features.



SPECIFICATIONS

Standard Stroke Lengths:	2½"	5"	8"	12"*
Dissipation @ 25°C.:	2.5w	5w	6w	8w

Independent Linearity:** ± 1% of total res. for 2½" and over.
**Under 2½": ± 2.5% to ± 5% of total res.

Resistance Range: 100 ohms/in. to 25,000 ohms/in.

Resistance Tolerance: ± 5%

Overtravel: Coin silver end tabs — no change in resistance in overtravel regions.

Ambient Temp. Range: −55° to + 80°C.

Temperature Coefficient of Resistance Wire: .00002/°C.

Life: 1,000,000 traverses over resistance element @ speed of 1"/sec. or less.

Resistance Winding: Up to three independent linear or non-linear windings may be included in one unit.

Housing: Sturdy machined aluminum with corrosion resistant anodized finish. Threaded bushing provides simple mounting. Terminals to all elements accessible at rear of housing. Housing has 1¼"x1¼" cross sectional area, length determined by stroke.

*Non standard stroke lengths ½" to 12" available.

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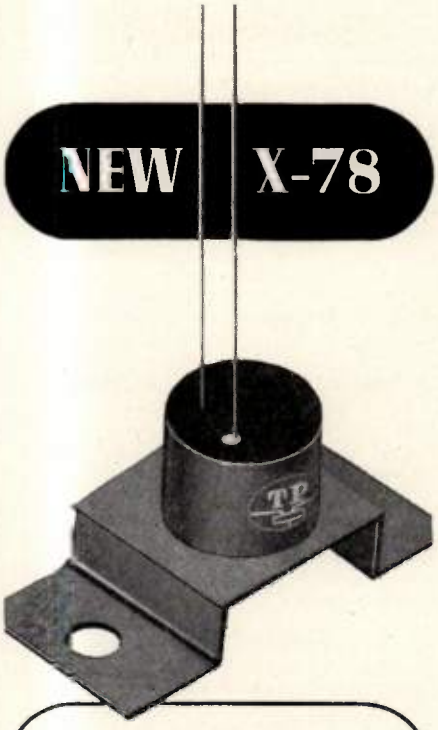
NEW X-78



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Size:
11" x 5" x 7"
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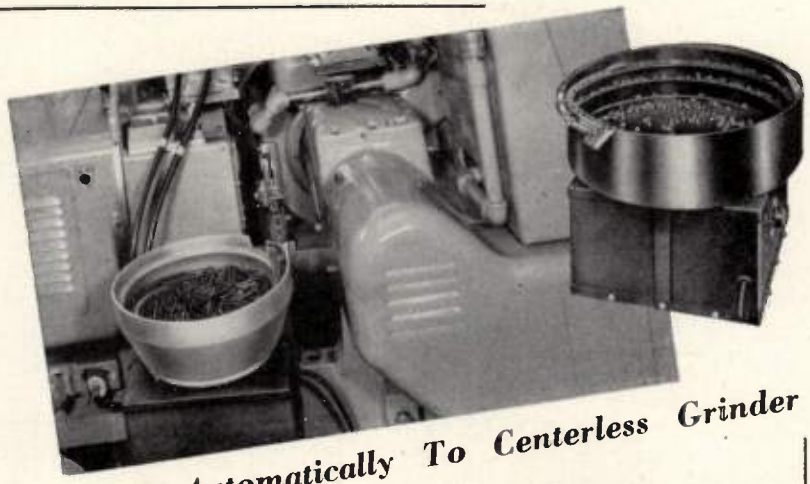

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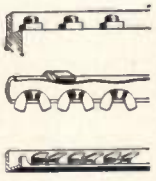

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(Continued from page 120A)

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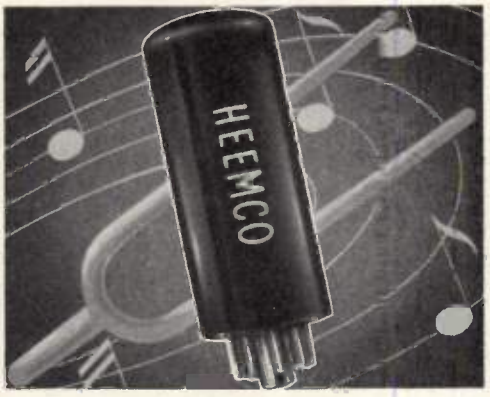
(Continued on page 128A)



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C.4	4.6	229	1.03"
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C.3	5.4	197	0.64"
C.22	5.5	184	0.44"
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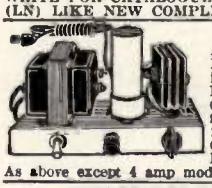

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(Continued on page 138A)

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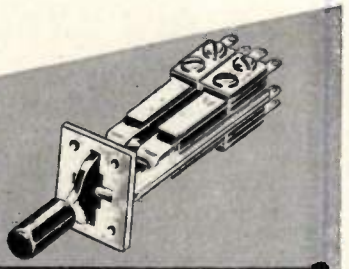

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RADAR SETS

SO-1 10 cm. with a range of 4, 20, and 80 miles. PPI presentation on a 5 inch screen. 360 deg. rotation of antenna with a pattern 8 deg. in horiz. and 18 deg. in vert. plane. Operates from 115 vdc. Set consist: of following: antenna, m/g—modulator, xmtr—rcvr, PPI unit, accessory control, and rectifier power unit.

MK 10 GUN RADAR. Extremely accurate, rugged and compact. Designed for ship board use to direct naval gun turrets. Antenna utilizes conical scanning for accurate pointing. Max. range is 20,000 yards with an error of pm 15 yds pm .1% of range. Pointing accuracy is pm .25 deg. Pulse dur. 0.5 usec. at PRF of 3600 cps. Pk. power output is 25 Kw. Primary power consumption is 1300 watts. Operates from 1.57, 60 CPS Source. BRAND NEW, COMPLETE WITH SPARES AND INSTRUCTION MANUAL ..... POR*

SQ Portable radar with type PPI, "A", or "B" Scope. Frequency is Approx. 3000 Mcs. 3 ranges: 3, 15, and 45 miles. Operates from 115v, 60 cycles $850

SE 10 cm radar for use on small surface craft for Sea-Search use. Max. range 80,000 yards, with an accuracy of 100 yards ± 1% of range. Bearing accuracy 2 degrees. Operates from 115 VDC. Complete Equipment. Brand new in original matched sets @ $850.00

SN Portable, lightweight, 10 cm set with ranges of 4 and 20 Mi. Presentation is on 5" "A" Scope. Operates from 115V, 50-60 Cy. Ideal for labs. Classrooms, and small boats $850.00

IFF SETS RC 148 RC 184 Navy BM
RC 145 RC 188

PRICE ON REQUEST

HI-POWER COMPONENTS

PLATE TRANS. Primary: 115V, 50-60 CY. Sec. 17, 600V/144 MA. Has "Built-In" Filter choke. Oil immersed $115.00

PLATE TRANS. Pri: 198/220/240V, 60 cy, 1 PH. Sec: 3650V/16.7 KVA, 30KV insulation. Oil-immersed, Less oil gauge $335.00

PLATE TRANS. AMERTRAN #31133. Pri: 110/115/120V/60cy/1 phase. Sec: 3140/1570V. 2.36 KVA $105.00

FIL TRANS. Pri: 220V/60cy/1 phase. Sec: 5 VCT/10A/30 KV TEST $37.50

PLATE TRANS. RAYTHEON UX 6801. Pri: 115V/60cy/1 PH. Sec: 22,000 V/234 MA/5.35 KVA. LO-CAP "DONUT" Construction $135.00

REACTOR RAYTHRON U-11533: 13.5H @ 1.0 AMP, 12.5 KV TEST $14.95

REACTOR, MODULATION: 50H/3A/80 OHMS DCR. Response: .03 cy-10 KC. Level: Plus 63 db. 40 KV. Test. Nominal Circuit impedance: 3000 Ohms $350.00

SWING REACTOR: 9-60 HY/.05-400 Ma. 10,000 V. Test Kenyon $14.95

MAGNETRONS

Type	Freq. Range(MC)	Peak Power Out (KW)	Duty Ratio	Price
2.2LA	9345-9405	50		$8.75
2.2[illegible]	3267-3333	265		7.50
2.2[illegible]	2965-2992	275	.002	19.95
2.3[illegible]	2820-2860	285	.002	24.50
2.3[illegible]	2780-2820	285	.002	28.50
2.3[illegible]*	3249-3263	5		16.50
2.3[illegible]*	3267-3333	8.7		24.50
2.4[illegible]	9310-9320	50	.001	24.50
2.4[illegible]	9000-9160	50	.001	59.50
2.5[illegible]*	9215-9275	50	.001	132.50
2.6M†	3000-3100	35	.002	34.50
2.6[illegible]†	2914-3010	35	.002	34.50
7*0B	690-700	40	.002	22.50
7*0D	710-720	40	.002	39.75
7*6EY	3038-3069	200	.001	32.50
7*6GY	2976-3007	200	.001	32.50
725-A	9345-9405	50	.001	Write
720-A	9345-9405	50	.001	24.50
4.3[illegible]	3550-3600	750	.001	169.45

*—Packaged with magnet.
†—Tunable over indicated range.

QK 60, 61, 62—$85 ea.

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723A ..... $12.50 | 2K25/723A/B $27.50
723A/B ..... 19.50 | 417-A (Sperry) 17.50

MICROWAVE COMPONENTS

"S Band" RG48/U Waveguide

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. .$22.50 EACH

DIRECTIONAL COUPLER. Broadband type "N" Coupling, 20 db. with std. flanges, Navy #CABV47AAN-2 (as shown) $37.50

LHTR, LIGHTHOUSE ASSEMBLY. Part 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 $49.50

BEACON LIGHTHOUSE cavity p/o UPN-2 Beacon 10 cm. Mfg. Bernard Rice, each $47.50

MAGNETRON TO WAVEGUIDE Coupler with 721-A Duplexer Cavity, gold plated $45.00

RT-30 APG-5 10 cm. lighthouse RF head c/o Xmtr.-Recvr-TR, cavity compl. recvr. & 30 MC IF strip using 60K5 (2C40, 2C43, 1B27 lineup) w/Tubes.

721A TR BOX complete with tube and tuning plungers $12.50

McNALLY KLYSTRON CAVITIES for 707B or 2K28 ..... $4.00

WAVEGUIDE TO ⅞" RIGID COAX "DOOR-KNOB" ADAPTER CHOKE FLANGE, SILVER PLATED BROAD BAND $32.50

AS14A AP-10 CM Pick up Dipole with "N" Cables $4.50

OAJ ECHO BOX, 10 CM TUNABLE $22.50

HOLMDELL-TO-TYPE "N" Male Adapters, W.E. #D167284 ..... $2.75

I.F. AMP. STRIP: 30 MC, 30 d.b. gain, 4 MC Bandwidth, uses 6AC7's—with video detector. A.F.C. less tubes $24.50

POLYROD 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" or "H" PLANE BENDS, 90 deg. less flanges $7.50

X Band— RG 52/U WAVEGUIDE

VSWR Measuring Section. Consisting of 6" straight section, with 2 pick-up, Type "N" Output Jacks, Mounted ½ Wave apart $8.50

1" x ½" waveguide in 5' lengths, UG 39 flanges to UG40 cover ..... per length $7.50

Rotating-Joints supplied either with or without deck mounting. With UG40 flanges .each, $17.50

Bulkhead Feed-thru Assembly $15.00

Pressure Gauge Section 15 lb. gauge and press nipple $10.00

Pressure Gauge, 15 lbs. $2.50

Directional Coupler, UG-40/U Take off 20db $17.50

TR-ATR Duplexer section for above $8.50

Waveguide Section 12" long choke to cover 45 deg. twist & 2½" radius, 90 deg. bend $4.50

Waveguide Section 3 ft. long silver plated with choke flange $5.75

Rotary joint choke to choke with deck mounting $17.50

90 degree elbows, "E" plane 2½" radius .. $12.50

Microwave Receiver, 3 CM. Sensitivity: 10-13μ Watts. Complete with L.O. and AFC Mixer and Waveguide Input Circuits. 6 I.F. Stages give approximately 120 DV gain at a bandwidth of 1.7 MC. Video Bandwidth; 2 MC. Uses latest type AFC circuit. Complete with all tubes, including 723A/B Local Oscillator $175.00

ADAPTER, waveguide to type "N", UG 81/U, p/o TS 12, TS-13, Etc. $27.50

ADAPTER, UG-163/U round cover to special btl. Flange for TS-45, etc. ..... $2.50 ea.

K Band—½" x ¼" W.G. 1.25 CM.

APS-34 Rotating joint $49.50
Right Angle Bend E or H Plane, specify combination of couplings desired $12.00
45° Bend E or H Plane, choke to cover $12.00
Mitered Elbow, cover to cover $4.00
TR-ATR-Section, Choke to cover $4.00
Flexible Section 1" choke to choke $5.00
"S" Curve Choke to cover $4.50
Adapter, round to square cover $5.00
Feedback to Parabola Horn with pressurized window $27.50
90° Twist $10.00

WAVEGUIDE FLANGES

UG 39/U ..... $1.10 | UG 51/U $1.65
UG 40/U ..... $1.25 | UG 52/U $3.40
UG 40A/U ..... $1.65 | UG 52A/U $3.40

PULSE NETWORKS

15A—1-400-50: 15 KV, "A" CKT. 1 microsec. 400 PPS, 50 ohms imp. $37.50

G.E. #3E (3-84-810) 8-2.24-405) 50P4T : 3KV "E" CKT Dual Unit: Unit 1, 3 sections, 0.84 Microsec. 810 PPS, 50 ohms imp.; Unit 2, 8 Sections. 2.24 microsec. 405 PPS 50 ohms imp. $6.50

7-5E3-1-200-67P. 7.5 KV "E" Circuit, 1 microsec. 200 PPS, 67 ohms impedance 3 sections $7.50

7-5E4-16-60, 67P. 7.5 KV "E" Circuit, 4 sections 16 microsec. 60 PPS, 67 ohms impedance $15.00

7-5E3-3-200-67P. 7.5 KV, "E" Circuit, 3 microsec. 200 PPS, ohms imp. 3 sections $12.50

#755: 10KV. 2.2usec., 375 PPS. 50 ohms imp. ... $27.50

#754: 10KV. 0.85usec., 750 PPS. 50 ohms imp. ... $27.50

KS8865 CHARGING CHOKE: 115-150 H @ .02A. 32-40H @ .08A, 30,700V Corona Test. 21KV Test $37.50

G.E. 25E5-1-350-50 P2T, "E" SKT, 1 Microsec. Pulse @ 350 PPS. 50 OHMS Impedance $69.50

KS9623 CHARGING CHOKE: 16H @ 75 MA. 380 Ohms DCR. 9000 Vac test $14.95

G.E. 6E3-.5-2000 50 P2T: 6 KV., "E" Circuit 0.5 usec /2000 PPS/50 ohms/2 sections $7.50

PULSE EQUIPMENT

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: 1-71B, 4-89-B, 3-'72's, 1-'73. New ..... Less Cover—$125

TPS-3 PULSE MODULATOR. Pk. power 50 amp. 24 KW (1200 KW pk): pulse rate 200 PPS, 1.5 microsec. pulse line impedance 50 ohms. Circuit series charging version of DC Resonance type. Uses two 705-A's as rectifiers, 115 v. 400 cycle input. New with all tubes $49.50

PULSE TRANSFORMERS

GE #-2748-A 0.5 usec @ 2000 Pps. Pk. Pwr out is 32 KW impedance 40:100 ohm output. Pri. volts 2.3KV Pk. Sec. volts 11.5 KV Pk. Bifilar rated at 1.3 Amp. Fitted with magnetron well $39.50

GE #K-2449A Primary: 9.33 KV, 50 ohms Imp. Secondary: 28 KV, 450 ohms. Pulse length: 1.0/5 usec @ 635/120 PPS. Pk. Power Out: 1,740 KW Bifilar: 1.5 amps $62.50

K-2745 Primary: 3.1/2.8 KV, 50 ohms Z. Secondary: 14/12.6 KV 1025 ohms Z. Pulse Length: 0.25/1.0 usec @ 600/600 PPS. Pk. Power 200/150 KW. Bifilar: 1.3 Amp. Has "bilt-in" magnetron well $42.50

K-2461A Primary: 3.1/2.6 KV—50 ohms (line). Secondary 14/11.5 KV—1000 ohms Z. Pulse Length: 1 usec @ 600 PPS. Pk. Power Out: 200/130 KW. Bifilar: 1.3 Amp. Fitted with magnetron well $39.75

UTAH X-151T-1: Dual Transformer, 2 Wdgs. per section 1:1 Ratio per sec 13 MH inductance 30 ohms DCR $7.50

UTAH X-150T-1: Two sections, 3 Wdgs. per section. 1:1:1 Ratio, 3 MH, 6 ohms DCR per Wdg. $7.50

68G711: Ratio: 4:1 6.7 Ohms. Pri: 0.23 Ohms sec. $4.50

TR1049: Ratio: 2:1 Pri. 220 MH, 50 Ohms, sec. 0.75H, DCR 100 Ohms $6.75

K-901695-501: Ratio 1:1, Pri. Imp. 40 Ohm, Sec. Imp. 40 Ohms. Passes pulse 0.6 usec with 0.05 usec rise $8.95

Ray UX 7896—Pulse Output Pri. 5v. sec. 41v. ...$7.50

Ray UX 8442—Pulse inversion—40v + 40v $7.50

PHILCO 352-7250, 352-7251, 352-7287

RAYTHEON: UX8693, UX5986 ..... $5 ea.

W.E.: D-166310, D-166638, KS 9800, KS9948.

UTAH #9262, with Cracked Beads, but will operate at full rated capacity $5.00

UX 8693 (SCS #2Z9627-54); 3 Wdgs. 32 turns #18 wire. DCR is: .362/.372/.4 ohms. Total voltage 2500 vdc. $5.00

D-166173: Input: 50 ohms Z. Output: 900 ohms 3 Wdgs. Freq. range 10 kc-2mc. P/O AN/APQ*-13 ..... $12.50

K-2450: Pulse-Inversion auto-transformer: primary 13 kv, 4 usec. Output: 14 kv @ 100 kw peak $34.50

THERMISTORS

D-164699 Bead Type DCR: 1525-2550 Ohms @ 75 Deg. F. Coefficient: 2% Per Deg. Fahr. Max. Current 25 MA AC/DC $2.50

D-167332 Bead Type, DCR is 1525-2550 Ohms. Rated 25 MA at .825-1.175 VDC $1.50

D-167613 Disk Type DCR: 355 Ohms @ 75 Deg. F.P.M. 2.5%, 1 Watt $1.50

D-166228 Disk Type 7120 Ohms @ 60°F. 4220 Ohms @ 80°F. 2590 Ohms @ 100°F.. 1640 Ohms @ 120°F. $1.50

HELMHOLTZ PHASE-SHIFTER

Stator consists of 4 loops oriented at 90 degrees to each other. Total stator inductance is 40MH. Rotor: 10MH. Total phase shift 0-360 deg. Designed for range unit of SCR-268 ..... 3.95 EACH

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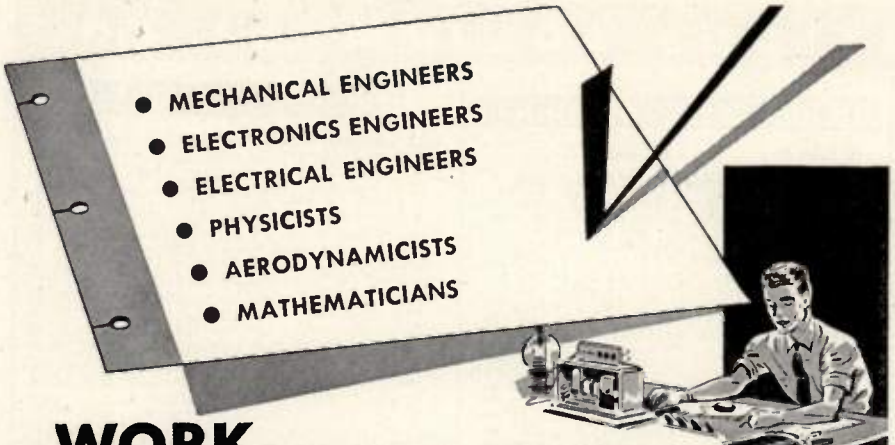

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DIVISION A

SANDIA *Corporation*

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Positions Open

(Continued from page 124A)

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The M.I.T. Instrumentation Laboratory is developing equipment for fire control, navigation and aircraft control. Several openings exist for electrical engineers who are recent graduates with outstanding academic and performance records to do electronic and servo-mechanism development and system design work followed by testing in the laboratory, in flight, and in the field. Opportunity for academic study. Send resumes to Instrumentation Laboratory, 68 Albany Street, Cambridge 39, Mass. Att: M. Phillips.

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An established manufacturer of small transformers, located in New Jersey, desires a design engineer. The position requires designing transformers to customers specifications plus technical and sales contact with customer. Applicant should have EE degree or equivalent plus experience in transformer design and specifications. Base salary to $8,000 depending on background and experience plus other benefits. Write Box 751.

ELECTRICAL ENGINEER

Successful sales and Business Manager EE, wants to start manufacturing company in the electronics field and seeks capable, experienced Electrical Engineer as partner. Applicants should be in a position to contribute full time plus sound ideas or a completely developed, marketable product. Limited investment may be required, but integrity, competence and ability to work together will be a paramount consideration. Location: New York area. Replies will be fully confidential. Box 752.

(Continued on page 130A)

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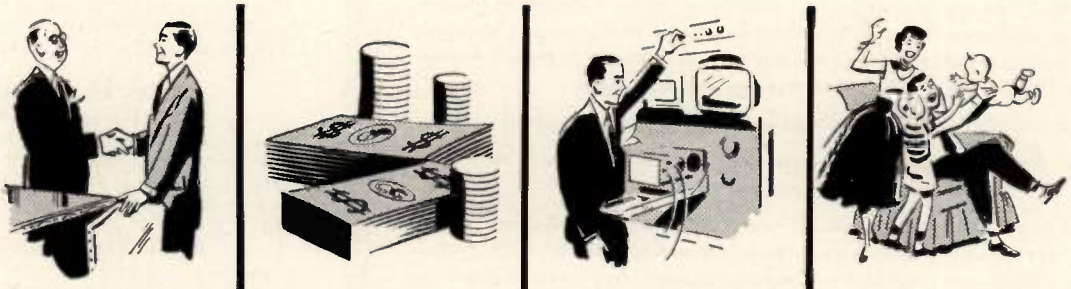

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***Positions Open In:* RESEARCH—DEVELOPMENT—DESIGN—APPLICATION**
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RADAR—Circuitry—Antenna Design—Servo Systems—Information Display Systems—Gear Trains—Stable Elements—Intricate Mechanisms

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Positions Open

(Continued from page 128A)

ELECTRONICS ENGINEER

Electronics Engineer to participate in magnetic resonance research. Capable of independent electronic investigation and design. University of Chicago, Research Institutes 5640 S. Ellis Avenue, Chicago 37, Ill. Att: Professor Clyde A. Hutchison, Jr.

ELECTRONIC ENGINEER

Do development work on electro-mechanical devices and associated equipment. Only experienced graduate engineers accustomed to work independently will be considered. Experience on tape recorders desired but not essential. Permanent position on non-defense projects. Salary open. Old established firm in N.Y. metropolitan area. Send complete resume and state salary expected. Box 753.

ELECTRICAL ENGINEERS

M.I.T. Lincoln Laboratory. Staff research positions available for exceptional electrical engineers with advanced training or experience in pulse circuitry and microwave applications. Please reply to Personnel Dept., P.O. Box 390, Cambridge, Mass.

ELECTRICAL ENGINEERS

Electrical Engineers experienced in Audio and Magnetic recording with an E.E. degree or equivalent. Laboratory assignment in Research Division. Gray Manufacturing Company, Hartford, Conn.

ELECTRONIC ENGINEER & ENGINEERING DRAFTSMAN

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(Continued on page 132A)

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Please send your resume to

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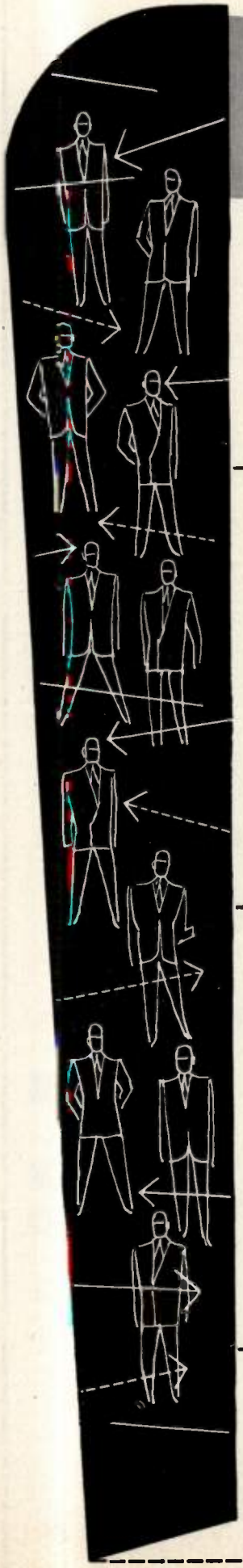

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REQUIREMENTS: At least a Master's degree in Mathematics, Physics or E.E. and 5 to 10 years experience including several years design work.

THREE OPENINGS

SENIOR RADAR INDICATOR
SYSTEMS ENGINEER

DUTIES: Systems planning, studying, proposing, carrying on applied research, development and design of advanced radar indicator systems for all types of radar systems including search, fire control and missle guidance. Coordination of 5 to 10 subordinates.

REQUIREMENTS: B.S. degree in E.E. or in Physics and 5 to 10 years design experience in the radar field with emphasis on indicators and analog devices.

FOUR OPENINGS

SENIOR RADAR ENGINEER

DUTIES: Applied research, development and design of advanced radar, fire control and missle guidance equipment. Openings exist in the fields of pulsed circuits, video, simulators, preamplifiers, linear and logarithmic post amplifiers and MTI and in analysis and synthesis of complex radar systems.

REQUIREMENTS: B.S. degree in E.E. or Physics and 3 or more years design experience.

THREE OPENINGS

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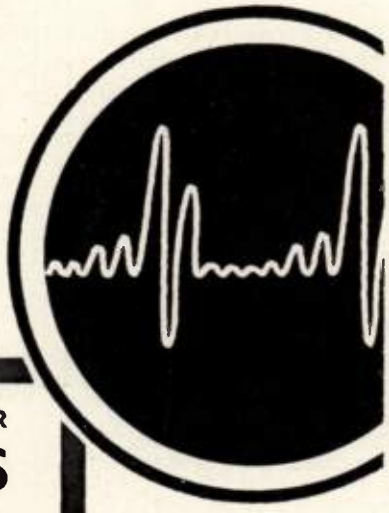
DUTIES: Preparation and editing of technical reports and instruction manuals for communications, radar and missle guidance equipment.

REQUIREMENTS: B.S. degree in E.E. or Physics and 3 years technical writing experience.

Or, if you have experience in fields of antennas and waveguides, high frequency heating applications, servo mechanisms, magnetic amplifiers, magnetic deflection yokes, magnetic materials, transformers, or components, send resume at once!

ADDRESS:
R. M. Swisher, Jr.
Employment Supervisor, Dept. F-4
Westinghouse Electric Corporation
109 W. Lombard Street
Baltimore 1, Maryland

Westinghouse
ELECTRONICS DIVISION
BALTIMORE MARYLAND



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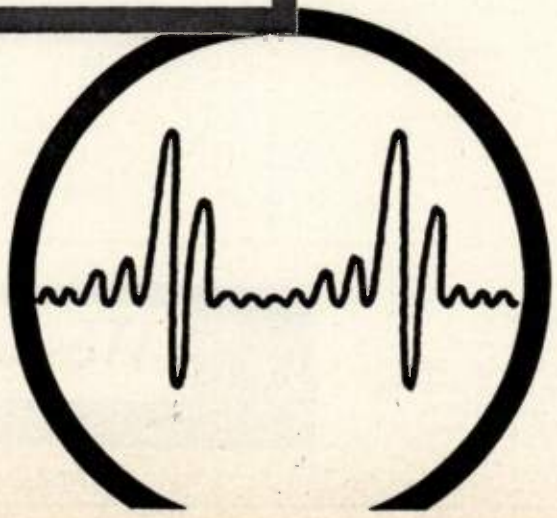

Positions Open

(Continued from page 130A)

cancies for junior electronics engineer and for engineering draftsman. Electronic Engineer at $5060 per year. Requirements: BEE degree plus 18 months professional experience in electronics.

Engineering Draftsman at $4205 per year. Requirements: Five years experience in engineering drafting. Interested individuals should apply to Industrial Relations Officer, Special Devices Center, Port Washington, New York.

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(Continued on page 134A)

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Los Angeles . . 57 . . California

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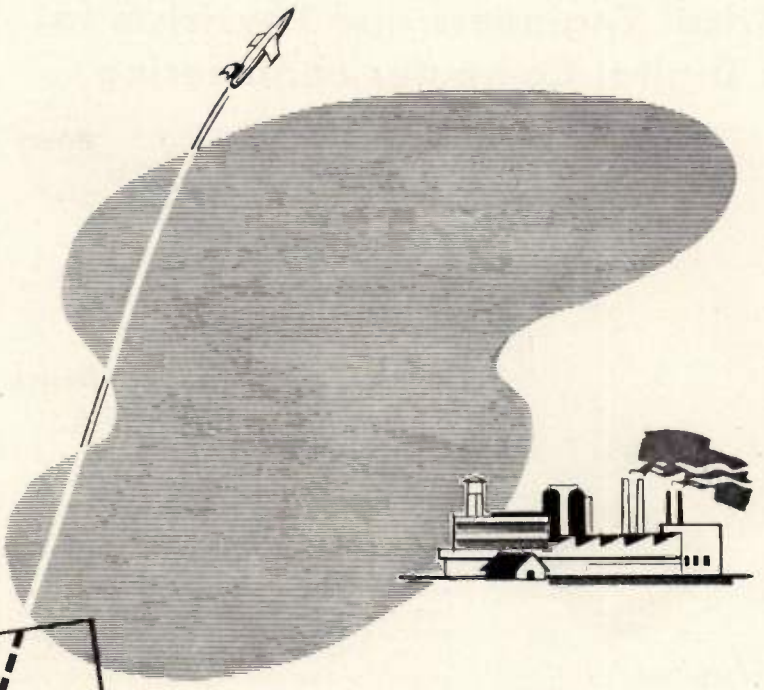

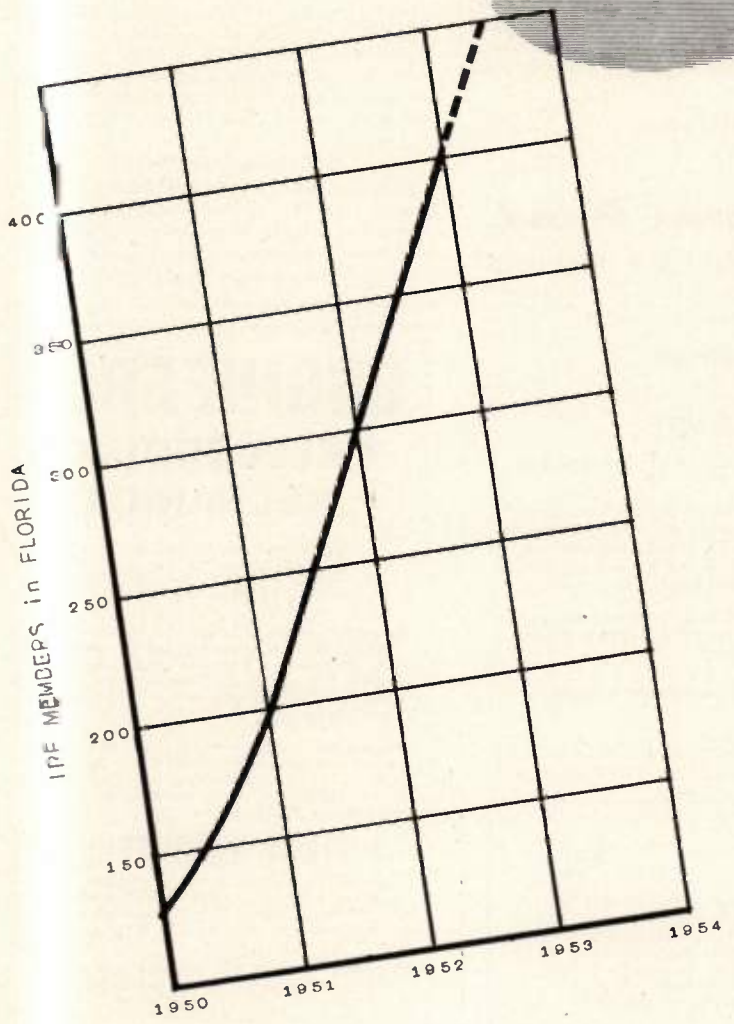

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Research Laboratories
4855 FOURTH AVENUE, DETROIT 1, MICHIGAN

Positions Open

(Continued from page 132A)

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Specializing in antenna design and analysis. Writing ability desirable. Permanent position in excellent location. Advancement commensurate with ability and results. State qualifications, salary desired and availability in first letter. Telrex, Inc., Box 879, Asbury Park, N.J.

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Nationally-known and well established midwest manufacturer of microwave components desires the services of electronic engineer to organize and head group to perform component design and de-

(Continued on page 136A)

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• MECHANICAL

Experience in design and Development of Radar and Sonar necessary.

Broad knowledge of Search and Fire Control Systems; Servo Mechanisms, Special Weapons, Microwave, Antennas and Antenna Mounts, Circuit Design, Filter Network, Precision Components (Capacitors—Resistors) and Communications.

Mechanical Engineer should also have experience in packaging of Electrical Equipment to Gov't specifications including design of complex cabinets, shock mounts and sway brace structures.

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Qualified to instruct in the operation and supervise installation, maintenance and repair of Radar, Sonar, Flight Simulators and allied electronic equipment in the Field.

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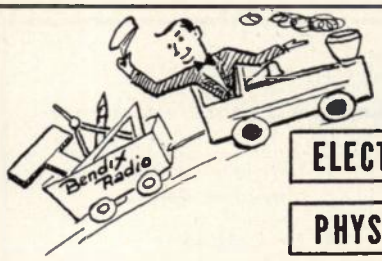

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(Continued from page 134A)

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Senior positions are available for two engineers with advanced degrees or industrial experience. One position is in the communications field, including high performance networks and special equipment in the frequency range of 100 kc to 250 mc. The work will include circuit and component design, followed by laboratory measurements and tests. The second position is in microwave engineering, including antenna design and systems and component design in the c-w radar field. Both of these jobs offer excellent opportunities with a small company located in suburban Boston. Please reply to Robert L. Blanchard, Chief Research Engineer, Trans-Sonics, Inc., Bedford, Mass.

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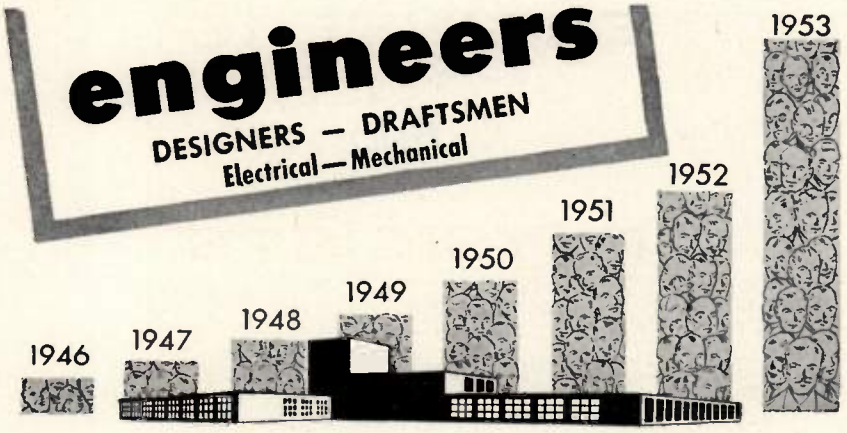

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(Continued from page 126A)

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PATENT ENGINEER

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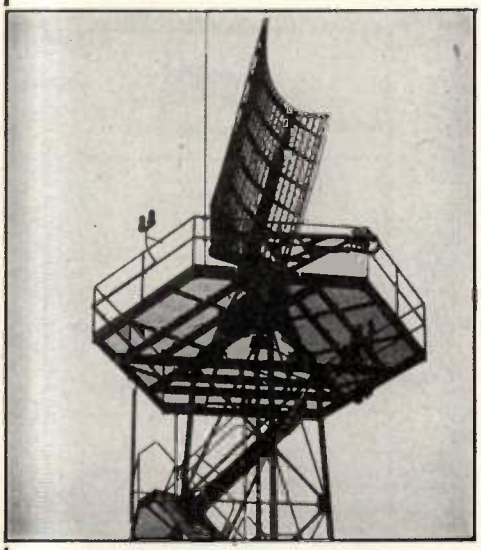

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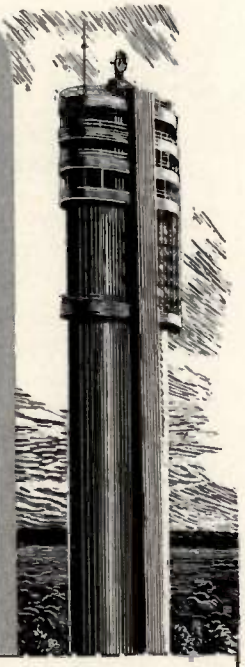

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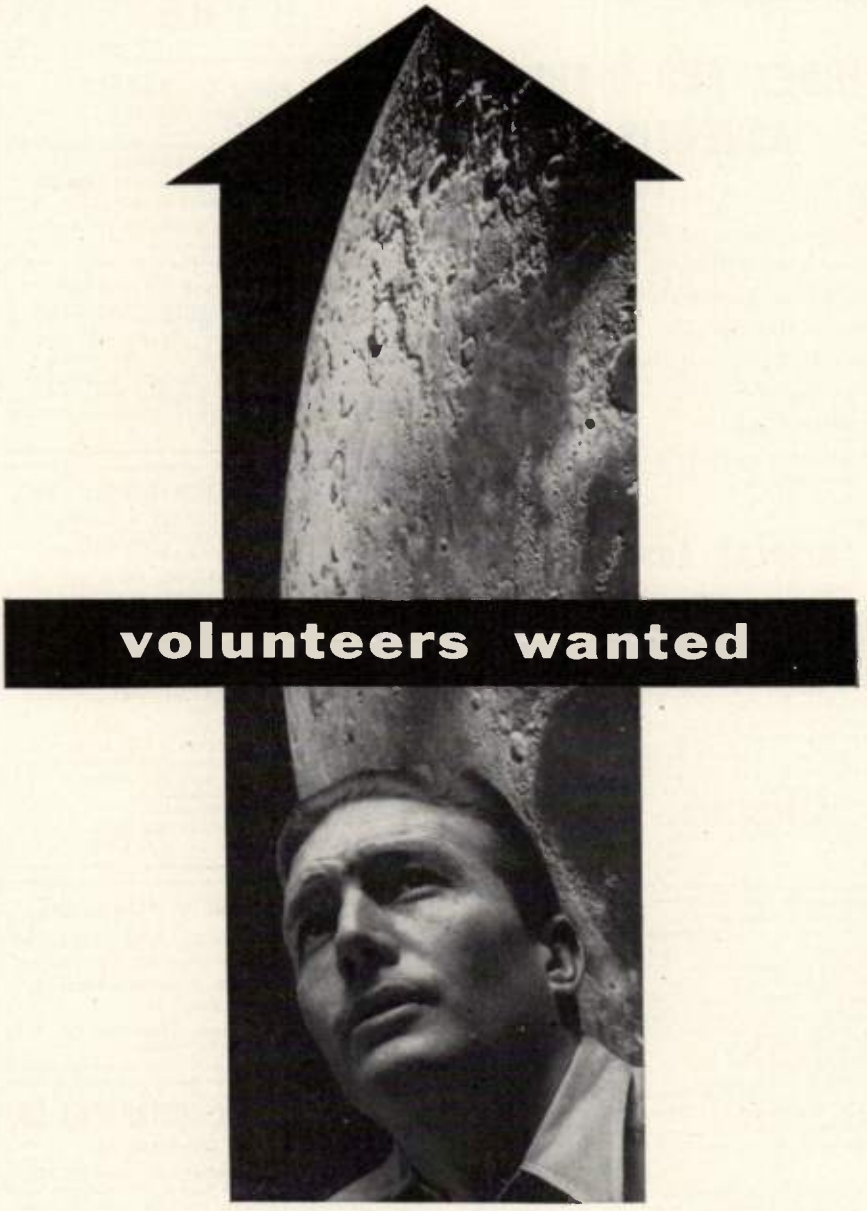

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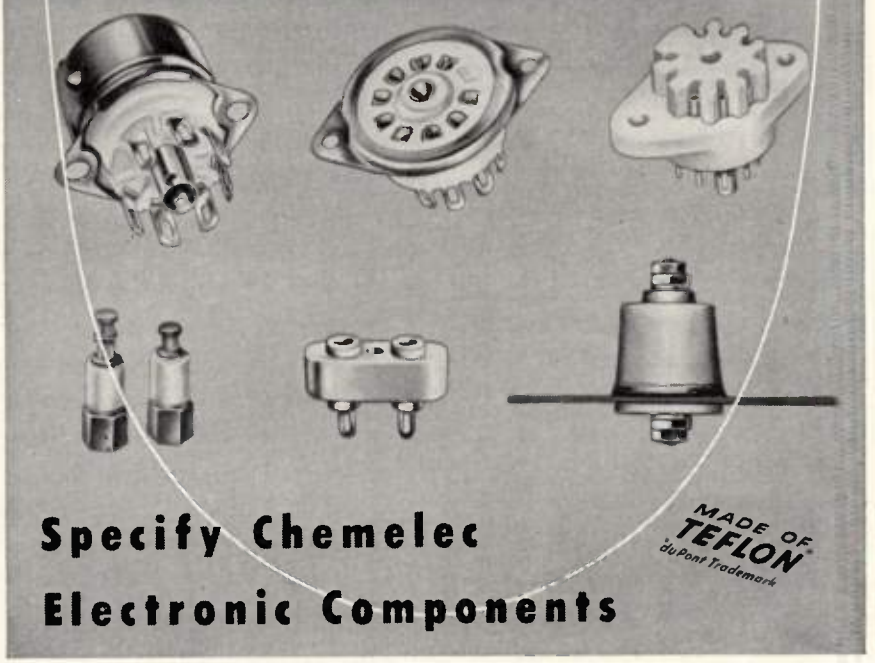

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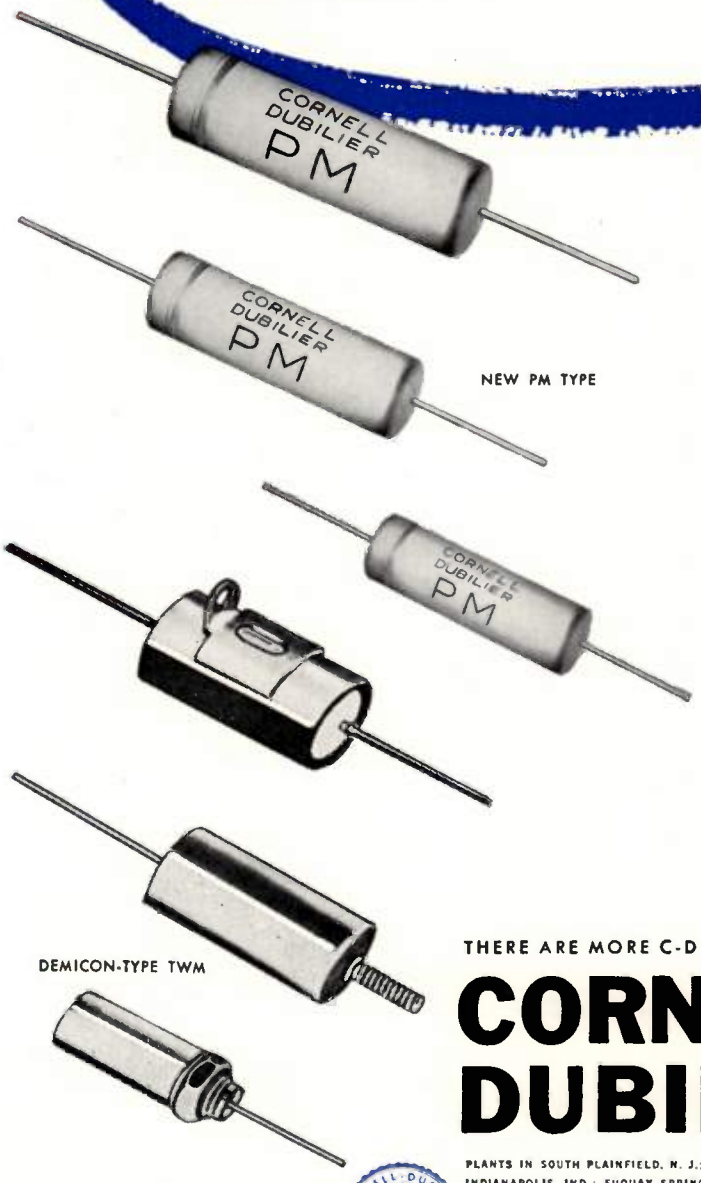

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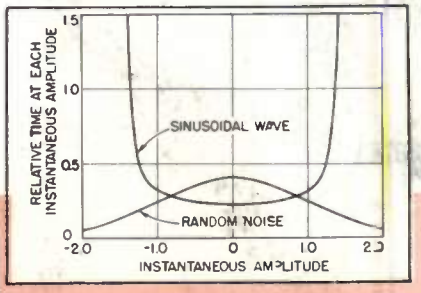

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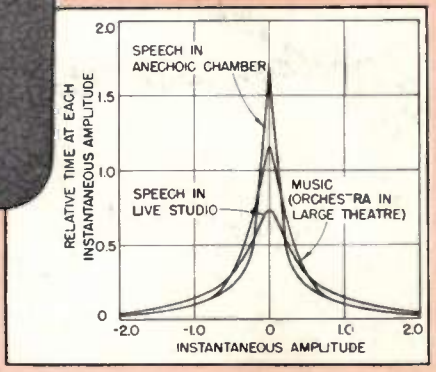

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